NONRESIDENT
TRAINING
COURSE

Aviation Electronics Technician - Intermediate

NAVEDTRA 14029

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PREFACE

By enrolling in this self-study course, you have demonstrated a desire to improve yourself and the Navy. Remember, however, this self-study course is only one part of the total Navy training program. Practical experience, schools, selected reading, and your desire to succeed are also necessary to successfully round out a fully meaningful training program.

COURSE OVERVIEW: In completing this nonresident training course, you will demonstrate a knowledge of the subject matter by correctly answering questions on the following: Servo systems, logic devices, communications, navigation systems, optic and infrared systems, television, computers and programming, waveform interpretation, and automatic test equipment.

THE COURSE: This self-study course is organized into subject matter areas, each containing learning objectives to help you determine what you should learn along with text and illustrations to help you understand the information. The subject matter reflects day-to-day requirements and experiences of personnel in the rating or skill area. It also reflects guidance provided by Enlisted Community Managers (ECMs) and other senior personnel, technical references, instructions, etc., and either the occupational or naval standards, which are listed in the *Manual of Navy Enlisted Manpower Personnel Classifications and Occupational Standards*, NAVPERS 18068.

THE QUESTIONS: The questions that appear in this course are designed to help you understand the material in the text.

VALUE: In completing this course, you will improve your military and professional knowledge. Importantly, it can also help you study for the Navy-wide advancement in rate examination. If you are studying and discover a reference in the text to another publication for further information, look it up.

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Published by
NAVAL EDUCATION AND TRAINING
PROFESSIONAL DEVELOPMENT
AND TECHNOLOGY CENTER

NAVSUP Logistics Tracking Number
0504-LP-026-7060

Although the words "he," "him," and "his" are used sparingly in this course to enhance communication, they are not intended to be gender driven or to affront or discriminate against anyone.

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CHAPTER 1

SERVO SYSTEMS

Chapter Objective: *Recall the purpose and functions of servo systems to include oscillation, zeroing synchro units, use of the synchro alignment set, antenna positioning servo systems, and hydraulic servo systems.*

As an Aviation Electronics Technician (AT), you will encounter various types of servo systems. The particular type (electromechanical, electro-hydraulic, hydraulic amplidyne, pneumatic, etc.) will depend upon the type of load for which it was designed. One of your primary jobs will be the control of radar antennas from a remote control station. We will discuss some methods of antenna control later in this chapter.

This chapter will not provide a detailed presentation of any one servo system. Instead, we will discuss the basic systems, identify their essential components, and explain the function of each component. For details concerning the theory and operation of a particular system, you should refer to the applicable technical manuals for that system. Before continuing, you should review the basic theory of synchros and servomechanisms discussed in Module 15 of the Navy Electricity and Electronics Training Series (NEETS), NAVEDTRA 172-15-00-8.

BASIC SERVOMECHANISMS

Learning Objective: *Identify the concepts and components of a basic servomechanism to include a data transmission system, servo control amplifier, and a servomotor.*

The essential components of a servomechanism are a data transmission system, a servo control amplifier, and a servomotor. These components are shown in the block diagram of figure 1-1, and are discussed in the following paragraphs.

The functions of the data transmission system are as follows:

1. To measure the servo output
2. To transmit or feedback the signal, which is proportional to the output, to the error detector (a differential device for comparing two signals)

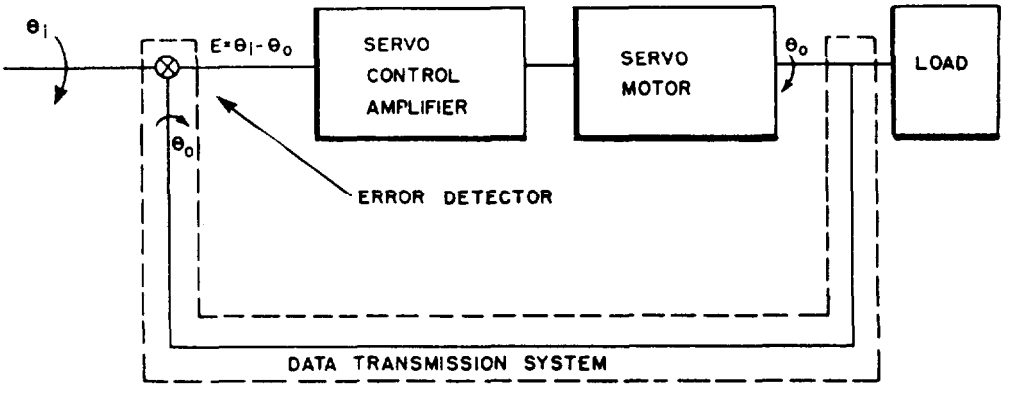


Figure 1-1.-Simplified block diagram of a servomechanism.

3. To compare the input signal with the feedback signal

4. To transmit to the servo amplifier a signal that is proportional to the difference between the input and output

The signal obtained by comparing the servo input and output is called the servo error, and is represented by the symbol E. Figure 1-1 shows that the servo error (E) is the difference between the input (θ_i) and the output (θ_o). This is stated mathematically as $E = \theta_i - \theta_o$.

In many servo systems, the physical location of the servo input device and output device are remotely located from each other, and may also be remotely located from the servo amplifier. This requires some means of transmitting the output information back to the device receiving the input command and transmitting the servo error to the servo amplifier. This system of transmission, as well as the comparing device (called an error detector), is part of an overall data transmission system. We discuss data transmission later in this chapter. The function of the servo amplifier is to receive the error signal from the error detector, amplify it sufficiently to cause the output device to position the servo load to the commanded position, and to transmit the amplified signal to the servomotor.

The servomotor positions the servo load. The motor must be capable of positioning the load within a response time based on the requirements of the system.

ERROR DETECTORS

The component of the data transmission system that compares the input with the servomechanism output is the error detector. An error detector can be either a mechanical or an electrical device. A simple form of a mechanical error detector is the differential. However, in aircraft weapons systems, most error detectors are electrical devices because of their adaptability to widely separated or remotely installed components. Most of the electrical devices used are of either the potentiometer (resistive) or one of several magnetic devices.

Electrical error detectors may be either ac or dc devices, depending upon the requirements of the servo system. An ac device used as an error detector must compare the two input signals and produce an error signal. The phase and amplitude of the error signal will indicate both the direction and the amount of control necessary to accomplish correspondence. A dc device differs because the polarity of the output error signal determines the direction of the correction necessary.

Error detectors are also used extensively in gyrostabilized platforms and rate gyros. In the stabilized platform, synchros are attached to the gimbals. Thus, any movement of the platform around the gyro axes is detected by the synchro, and the error voltage is sent to the appropriate servo system.

In rate gyros, an E-transformer (discussed later) is commonly used to detect gyro precession. It is extremely sensitive to very slight changes, but its movement is limited to a very small amount. Thus, it is extensively used with constrained gyros.

POTENTIOMETER

Potentiometer error detector systems are generally used only where the input and output of the servomechanism have limited motion. They

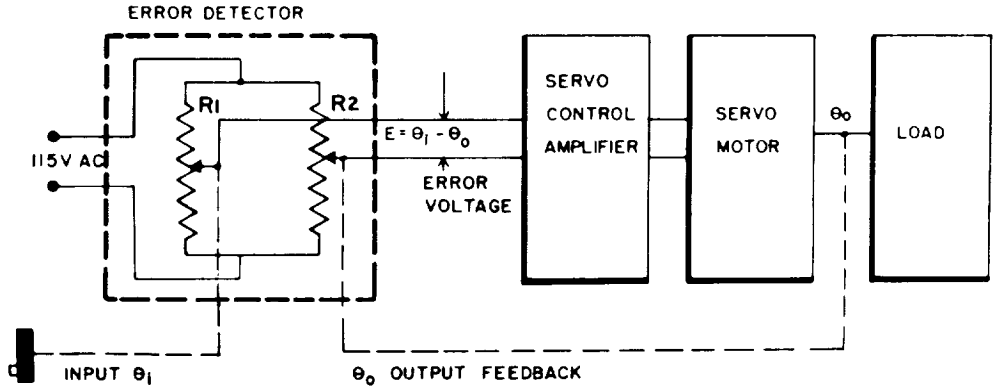


Figure 1-2.-Balanced potentiometer error detector system.

are characterized by high accuracy, small size, and the fact that a dc or an ac voltage may be obtained as the output. Their disadvantages include limited motion, a life problem resulting from the wear of the brush on the potentiometer wire, and the fact that the voltage output of the potentiometer changes in discrete steps as the brush moves from wire to wire. A further disadvantage of some potentiometers is the high drive torque required to rotate the wiper contact.

An example of a balanced potentiometer error detector system is shown in figure 1-2. As we have indicated, the purpose of the circuit is to give an output error voltage that is proportional to the difference between the input and output signals.

The command input shaft is mechanically linked to R1, and the load is mechanically linked to R2. An electrical source of 115 volts ac is applied across both potentiometers.

When the input and output shafts are in the same angular position, they are in correspondence and there is no output error voltage. If the input shaft is rotated, moving the wiper contact of RI, an error voltage is developed and applied to the control amplifier. This error voltage is the difference of the voltages at the wiper contacts of R1 and R2. The output of the amplifier causes the motor to rotate both the load and the wiper contact of R2 until both voltages are equal. When equal, there is no output error voltage.

Figure 1-2 illustrates R1 and R2 grouped together. In actual practice the potentiometers may be positioned remotely from each other, with R2, the output potentiometer, being located at the output shaft or load. The remote location of one of the components does not remove it from being a part of the error detector.

E-Transformer

The E-transformer is a type of magnetic device used as an error detector. Its application is useful in systems that do not require the error detector to move through large angles. A simplified drawing, which is one of several possible devices in this category, is shown in figure 1-3.

The primary excitation voltage is applied to coil A on the center leg of the laminated core. The coupling between coil A and the secondary windings, coils B and C, is controlled by the armature, which is displaced linearly by the input signal. When the armature is positioned so the coupling between the windings is balanced (null), the output voltage is minimum because of the series-opposing connections of the secondary windings. The phase of the output voltage on either side of the voltage null differs by 180 degrees. By proper design of the transformer, the amplitude can be made proportional to the displacement of the armature from its null voltage position. This type of error detector has the advantages of small size and high accuracy. It has the disadvantage of permitting only limited input motion.

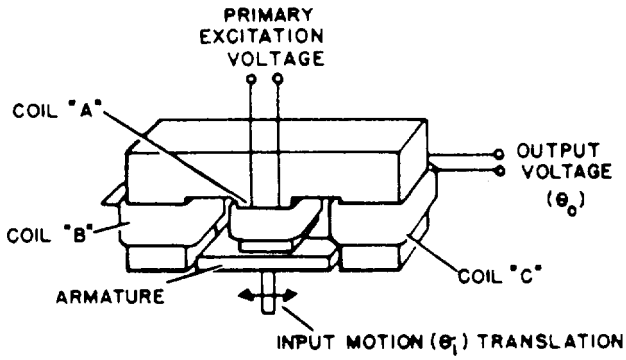


Figure 1-3.-E-transformer error detector.

Control Transformer

Synchros have been developed to a point of relatively high accuracy, low noise level, reasonably small driving torques, and long life. These qualities also apply to synchro control transformers. A primary advantage of the synchro control transformer over other types of error detectors is its unlimited rotation angle; that is, both the input and the output to the synchro control transformer may rotate through unlimited angles. Among the disadvantages of synchros (including the synchro control transformer) are the large size necessary to maintain high accuracy, the power consumed, and the output supplied to the servo control amplifier is always ac modulated with the servo error.

Alternating current may be used if the two following conditions are met:

1. The frequency of the ac used must be greater than the maximum frequency response of the measuring devices used.

2. If negative values of the variables are allowed, the devices used can be phase-sensitive.

Figure 1-4 shows a dc signal and the same function represented by an ac voltage. The instantaneous value of the ac signal does not indicate the value of the function, but the average value of the ac signal may be used to represent the value of a function. If the ac signal is the input to a servomotor, for example, the motor must not attempt to follow every variation of the ac signal, but must follow the average value. The second condition is essential because a negative ac signal does not exist. However, negative values can be indicated by a change in phase of the signal. Note that in figure 1-4, during the period when the dc signal is positive, the positive peaks of the ac signal correspond to the positive peaks of the ac reference. During the period when the dc signal is negative, the positive peaks of the ac signal correspond to the negative peaks of the reference; i.e., the signal is 180 degrees out of phase with the reference. Alternating-current servomotors are available. These servomotors will rotate in one direction when the input signal is in phase with a reference voltage, and in the other direction when the signal is out of phase with the reference voltage.

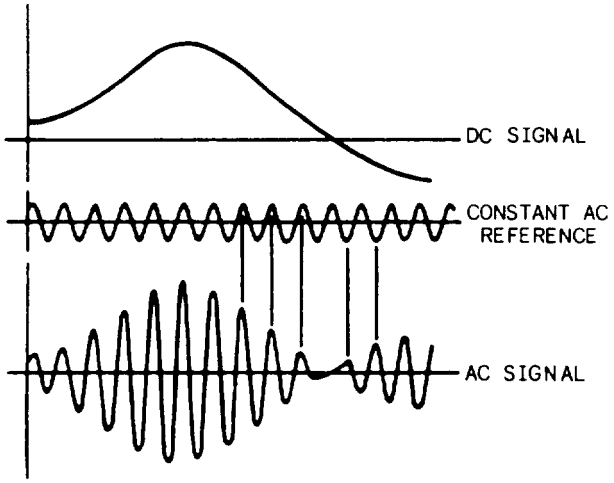


Figure 1-4.-AC modulated with the servo error.

A synchro data transmission system is comprised of a synchro transmitter, a synchro control transformer, and, in some cases, a differential transmitter for additional servo inputs. The synchro transmitter transforms the motion of its shaft into electrical signals suitable for transmission to the synchro control transformer, which comprises the error detect or (fig. 1-5).

The stator of the transmitter consists of three coils spaced 120 electrical degrees apart, The voltage induced into the stator windings is a function of the transmitter rotor position. These voltages are applied to the three similar stator windings of the synchro control transformer. The voltage induced in the rotor of the synchro control transformer depends upon the relative position of this rotor with respect to the direction of the stator flux.

The variation of the synchro control transformer output voltage as a function of the rotor position relative to an assumed stator flux direction is shown in figure 1-6. While there are two positions of the rotor, 180 degrees apart, where the output voltage is zero, only one corresponds to a stable operating position of the servo.

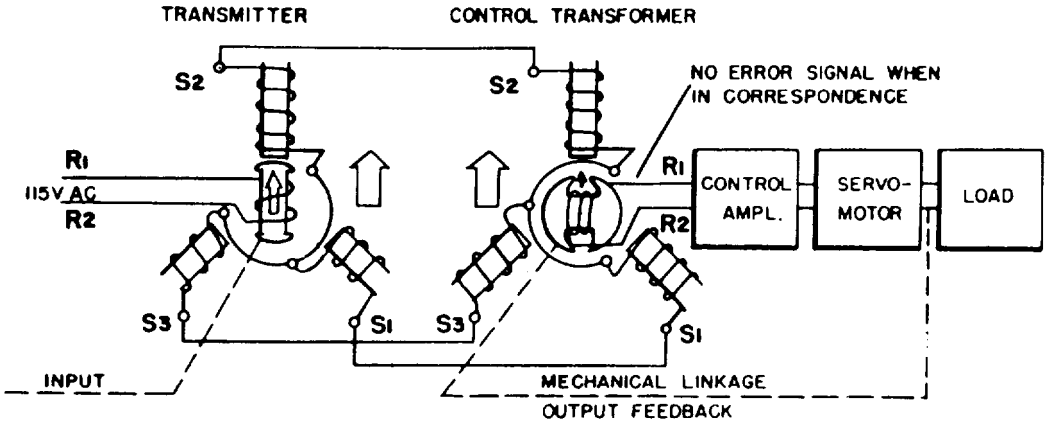


Figure 1-5.-The control transformer as an error detector.

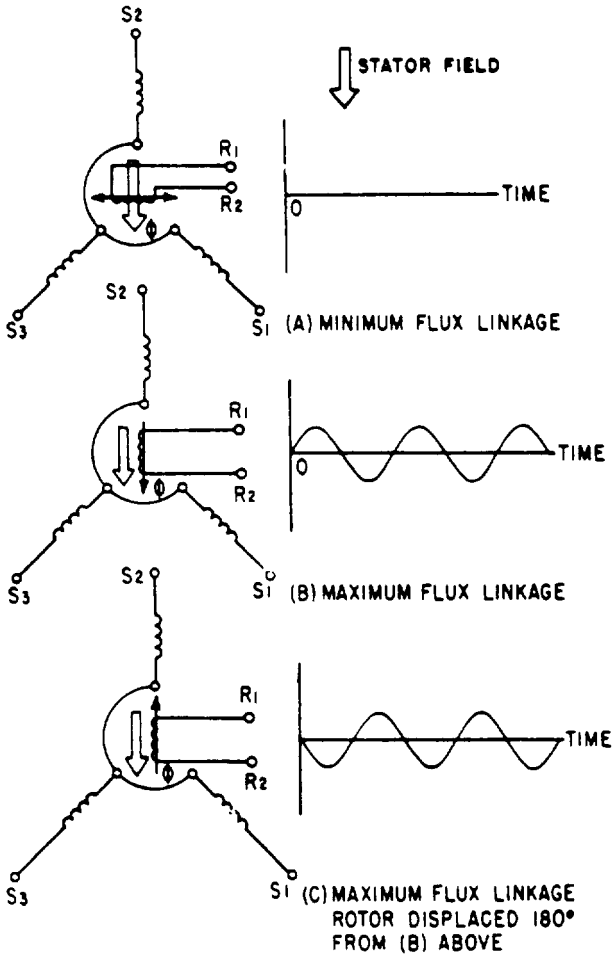


Figure 1-6.-Induced voltage in synchro control transformer rotor.

When a synchro differential transmitter is used for additional inputs to the servo system, it is connected between the synchro transmitter and the synchro control transformer (fig. 1-7). When the synchro differential rotor is in line with its stator windings, the differential transmitter acts as a one-to-one ratio transformer, and the voltages applied to the synchro control transformer are the same as the voltages from the synchro transmitter. If the synchro differential transmitter rotor is displaced by a second input, the voltages from the synchro transmitter to the control transformer are modified by the synchro differential transmitter by the amount and direction of its rotor displacement. Thus, the two inputs are algebraically added and fed to the synchro control transformer as a single input.

Flux Gate

A flux gate element may be used to drive or excite a control transformer and is usually used in compass systems. The flux gate operates on the principle of using the earth's magnetic field to produce a second harmonic current flow in the element. This, in turn, produces a voltage in the stator windings of the control transformer that is in direct proportion to earth's magnetic north. Because it is desirable to use only the horizontal component of the earth's field, a gyro is used to hold the element level with the earth's surface. Another method is to suspend the element by a spring and use the properties of a pendulum to rigidly mount it to the aircraft so that it turns in an azimuth as the aircraft turns.

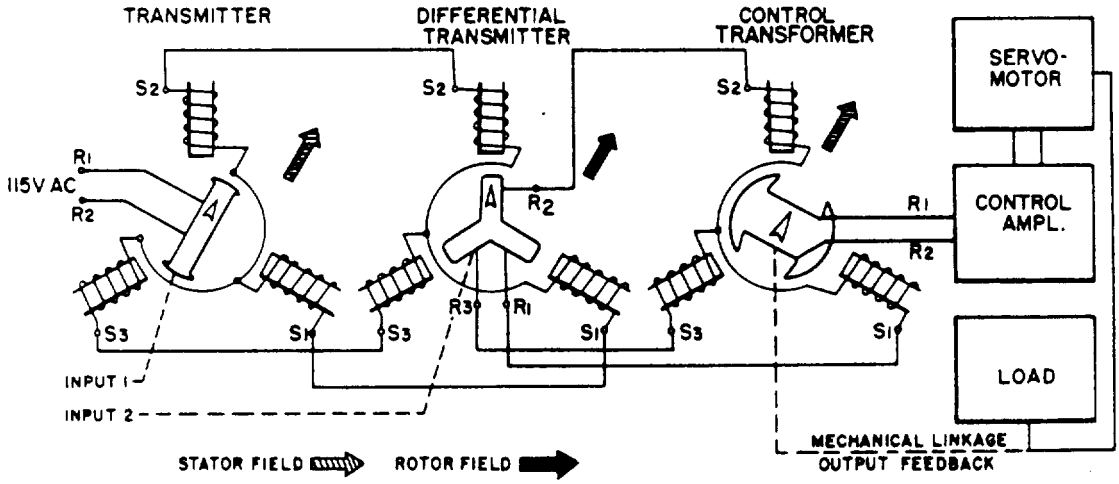


Figure 1-7.-Synchro differential transmitter used for additional input.

MULTIPLE-SPEED DATA TRANSMISSION SYSTEMS

The static accuracy (how accurately the load is controlled) of a servomechanism is frequently limited only by the accuracy of the data transmission system. The accuracy of the data transmission system may be increased considerably by employing a multiple-speed data transmission system along with a 1-speed system. The error-detector elements of the multiple-speed transmission system rotate at some multiple of the shaft being controlled. The elements of the 1-speed transmission system operate one to one with respect to the controlled shaft.

The schematic diagram of a multiple- and a 1-speed system is shown in figure 1-8. If a system can transmit data at two different speeds, it is referred to as a dual-speed system. In this example, if the input shaft turns through 1 degree, the 1-speed transmitter also is rotated 1 degree while the multiple-speed unit is rotated 10 degrees. The synchro control transformer associated with each of these transmitters is geared in similar ratios with respect to the servo output shaft. A 1-degree error between the position of the input and output shafts produces a relative rotor displacement of 1 degree in the 1-speed synchros, and 10 degrees in the multiple-speed synchros. If the relation between the rotor displacement and output voltage is linear, the error signal from the multiple-speed system is 10 times that from the 1-speed system. This amplification of the error signal in the data transmission link reduces the signal amplification required in the servo controller. If the synchro has an inherent error of 0.1 degree with respect to its own shaft, the consequent servo error introduced by a 1-speed data transmission system will be of corresponding magnitude. The consequent servo error introduced by a 10-speed data transmission system will be only one-tenth as great, or 0.01 degree.

A disadvantage of using a multiple-speed error detector lies in the possibility of the system falling out of step. If this happens, it will synchronize in a position differing from the correct position by an integral number of revolutions of the multiple-speed synchro. In the example shown in figure 1-8, if the output shaft were held fixed and the input shaft rotated 36 degrees, the 10-speed synchro transmitter would turn one complete revolution. At this point, the error signal from the multiple-speed error detector would be zero. If the output shaft were then released, the system would operate in a stable fashion with a 36-degree error between the input and output shafts. The purpose of using a 1-speed detector is to prevent this ambiguous synchronization.

An error signal selector circuit is provided that switches control of the servo to the 1-speed data transmission system. This occurs whenever the

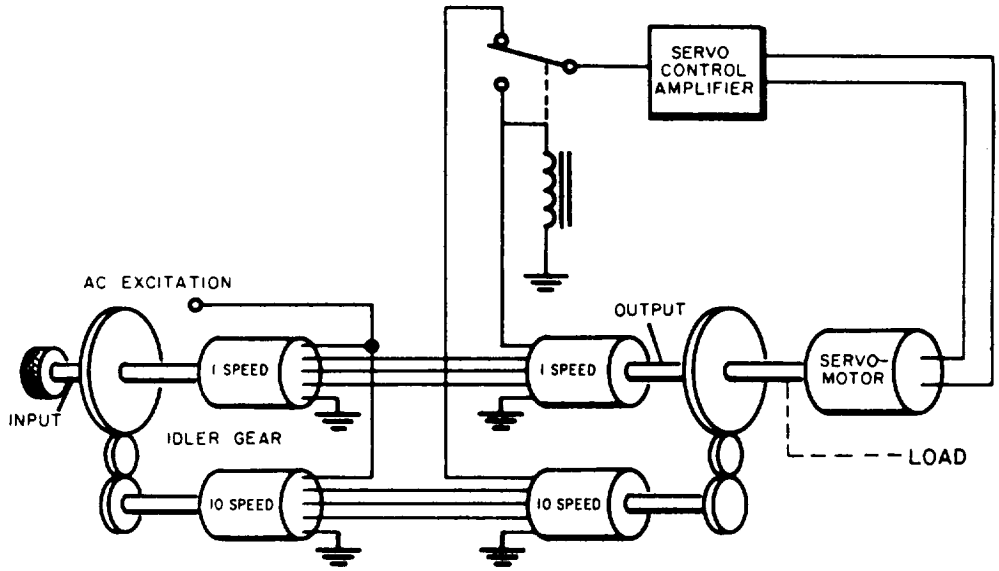


Figure 1-8.-Dual-speed data transmission system.

servo error becomes large enough to permit the multiple-speed system to synchronize falsely.

The simplest device imaginable that could control an error-selector circuit is shown in figure 1-8. It is essentially a single-pole, double-throw relay actuated by the output of the 1-speed error detector. The relay is shown in the de-energized position. When the output of the 1-speed synchro is high, the relay is energized and the 1-speed circuit controls the servomotor. When the output is low, the relay opens and the 10-speed synchro controls the circuit. Keep in mind that the synchro output is high only when there is a large error.

The relationship of the coarse (1-speed) synchro output and the fine (10-speed) synchro output is shown in figure 1-9, view A. The shaded portion represents the area where control can be switched from the l-speed circuit to the 10-speed circuit. With the selector circuit shown, it is still possible to have a single ambiguous position of the 1-speed (coarse) synchro. At this point the 1-speed (coarse) and 10-speed (fine) shafts are nulled (but are 180 degrees out of phase) and control is switched to the 10-speed circuit.

One way of eliminating this false synchronization position is to drive the multiple-speed synchro at any odd multiple of the 1-speed synchro. Figure 1-9, view B, shows the phase relationship of a 1-speed and 7-speed system. Although there is still a null of both synchros at the 180-degree position of the 1-speed synchro, their outputs are in phase. This position is an unstable one, and the servo will not remain at this point.

The system illustrated in figure 1-8 is not found in operating equipment due partly to the load the relay places on the 1-speed synchro. In actual practice, the relay could be controlled by an electronic circuit operated by the synchro voltages. A method commonly used feeds the outputs of the synchros to an electronic circuit biased so that the fine-synchro voltage is not used when the coarse-synchro voltage is high. This method does not require a relay.

The disadvantage of using multiple-speed error detectors is the need for an additional synchro system and switching circuit. This additional equipment is needed if increased servo accuracy accounts for the wide use of these multiple-speed data transmission systems. This results from the amplification of the error signal and the effective reduction of inherent synchro errors.

SERVO CONTROL AMPLIFIERS

Earlier, we stated that the output of an error detector (error voltage) can be fed to a servo control amplifier. This type of signal is small in amplitude and requires sufficient amplification to allow actuation of a prime mover. In addition to amplification, the servo control amplifier must, in some cases, transfer the error signal into suitable form for controlling the servomotor or output member. It may also include provisions for special characteristics necessary to obtain stable, fast, and accurate operation.

Servo amplifiers used in aircraft weapons systems are limited to electronic and magnetic types. The operation and explanation of electronic amplifiers and their circuits are discussed in Module 8 of the Navy Electricity and Electronics Training Series (NEETS), NAVEDTRA 172-08-00-82.

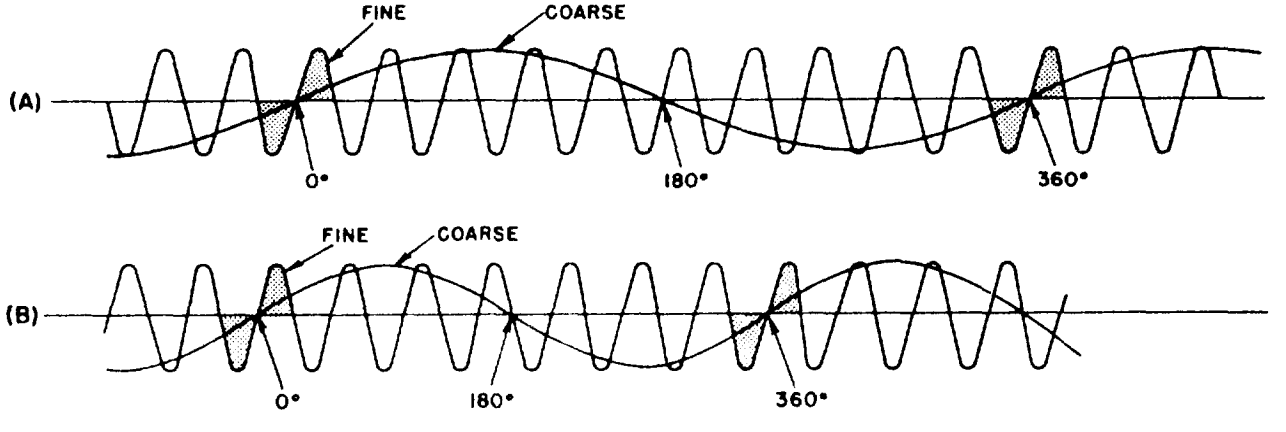


Figure 1-9.-Phase relationship of fine and coarse synchro voltages; (A) single-wed and 10-speed; (B) single-speed and 7-speed.

In addition to the requirements of basic amplifiers, servo amplifiers must also meet certain additional requirements as follows:

1. A flat gain versus frequency response for a frequency well beyond the frequency range used.
2. A minimum of phase shift with a change in level of input signal. Zero phase shift is desired, but a small amount can be tolerated if constant.
3. A low output impedance.
4. A low noise level.

Servo amplifiers may use either ac or dc amplifiers or a combination of both. The application of dc amplifiers is limited by such problems as drift and provisions for special bias voltages needed in cascaded stages. Drift, a variation in output voltage with no change in input voltage, can be caused by a change in supply voltage or a change in value of a component. Consequently, many servo amplifiers use ac amplifiers for voltage amplification.

MODULATORS

As pointed out previously, ac amplifiers are the best to use for amplifying an error signal. They do not need well-regulated power supplies and costly precision components; however, some aircraft weapons systems use a dc voltage for an error signal. The dc error voltage maybe changed to an ac signal by the use of a modulator (sometimes called a chopper). Modulator circuits used in servo control amplifiers must be phase sensitive and produce an ac output signal, whose amplitude is proportional to the dc input signal and whose phase is indicative of the polarity.

Vibrator Modulators

A modulator may be either an electromechanical vibrator or an electronic circuit. An example of a vibrator modulator is shown in figure 1-10. An ac supply voltage is employed to vibrate the contacts of the vibrator in synchronism with the supply voltage. The dc error voltage is applied to the center contact of the vibrator. Assume that the reference voltage will cause the cycle, and point B during the second half cycle. The output is represented by waveform B if the error voltage is positive, and by waveform C if it is negative.

Electronic Modulator

An example of an electronic modulator circuit is shown in figure 1-11. The circuit shown is a diode ring modulator and works by causing a changing current to flow through one-half of the primary of transformer T2, and then through the other half at a 400-hertz rate. Each half-cycle of changing current produces a half-cycle of sinusoidal output voltage. The phase of this output voltage compared to the 400-hertz carrier depends upon the direction of current through each primary half.

Diodes CR1 and CR4 are forward biased when the dc control voltage is positive. Diodes CR2 and CR3 are forward biased when the dc control voltage is negative. When two of the diodes are

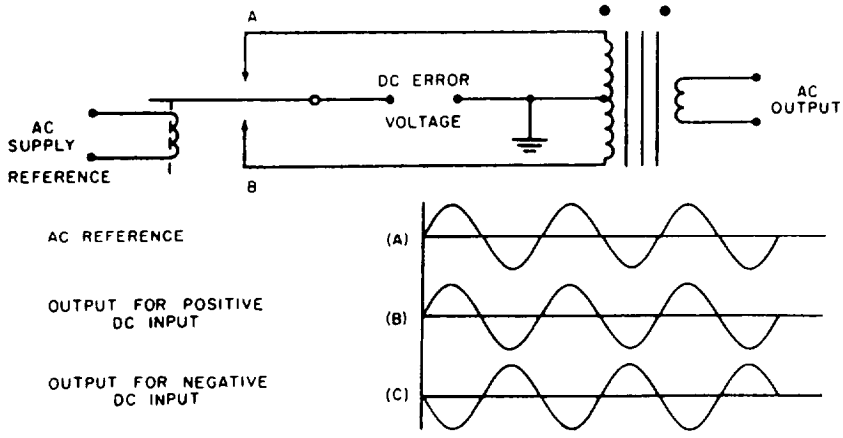


Figure 1-10.-Vibrator modulator.

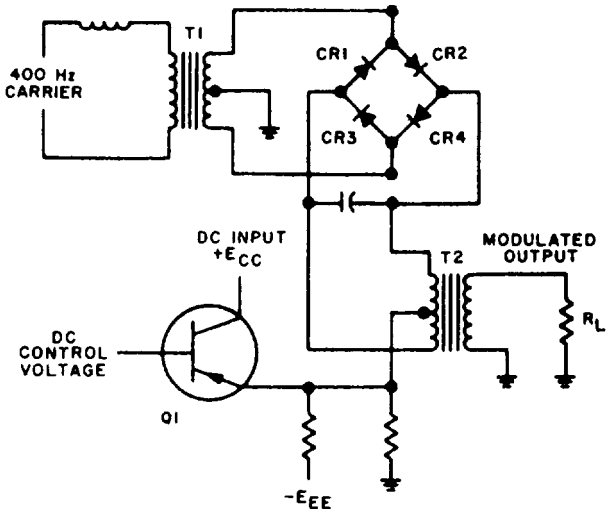


Figure 1-11.-An electronic modulator.

forward biased by the dc control voltage, the other two are back biased and cut off. As long as the instantaneous amplitude of the carrier voltage is less than the dc control voltage, the cutoff diodes remain back conducting diodes and through one of the half windings.

When one of the back biased diodes becomes forward biased (the amplitude of carrier voltage exceeds the dc control voltage), the diode conducts. This interrupts the current flowing through the half winding. The result is that the output voltage amplitude is clipped at the value it had when the current was interrupted.

The capacitor connected across the primary of T2 filters any high frequency components associated with the clipped half-cycle of the sine wave so that a nearly sinusoidal output half-cycle occurs. The output's amplitude is approximately equal to the output voltage at the time of clipping.

The capacitor operates by coupling the high frequency components of the clipped voltage through the nonconducting half windings. The high frequency components are canceled because they produce currents that flow in opposite directions in both halves of the center tapped primary windings; that is, they produce magnetic fields that cancel each other.

The amplitude of each half-cycle of the 400-hertz carrier voltage is modulated by the dc control voltage. The polarity of the control voltage determines the phase of the modulated carrier voltage output relative to the unmodulated carrier voltage input. This is done as a result of the direction of current flow through the half winding. This direction depends upon which diode is forward biased as a result of the polarity of the dc control voltage.

PHASE DETECTORS

We have stated that an ac amplifier has inherent advantages over a dc amplifier, that a dc error voltage can be changed into an ac signal, and the ac signal can be amplified and applied to an ac servomotor. However, some systems use dc servomotors, which necessitates converting the ac signal to dc. To do this, use a phase detector, sometimes called a demodulator.

Bridge Phase Detectors

Figure 1-12 displays a phase detector using a bridge circuit. With no error input signal and only the reference voltage applied, CR1 and CR2 would conduct in series when point C is on its positive half-cycle. When point C is on its negative half-cycle, CR3 and CR4 would conduct in series. Assuming the drops across the diodes and resistances to be equal, points A and B would be at ground potential on both half-cycles and the output voltage would be zero.

When an error signal is applied to the bridge in phase with the referenced voltage and points A and C are both on their positive half-cycle, electron flow will be from point G on the reference transformer T2 to point D, through CR2 to point A, from point A to the center tap on T1, and to E through R_L to G. On the next half-cycle, both points A and C will change polarity and the electron flow will be from point G to point C, through CR3 to point B, through T1 to the center tap, to

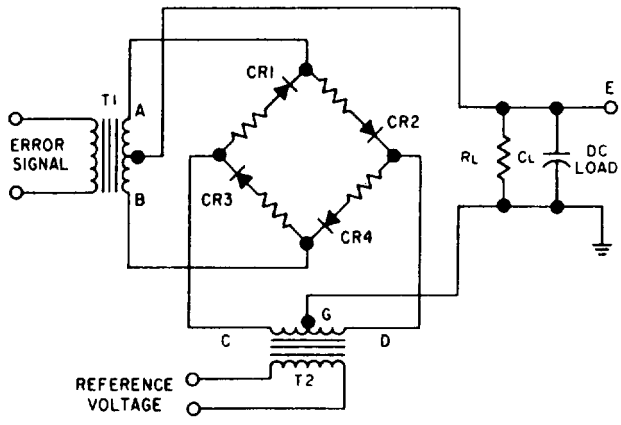


Figure 1-12.-Bridge phase detector.

the right to point E, and through R_L to G. On both half-cycles of reference and error voltage, the electron flow was down through R_L to ground, developing a negative dc output voltage.

If the error signal is applied out of phase with the reference voltage and positive at points A and D, electron flow will be from point G up through R_L, left to the center tap of T1, down to point B, through CR4, down to point D, and left to point G. On the next half-cycle, both points A and D will have G up through R_L to the center tap of Tl, up to point A, through CR1 to point C, and right to the center tap to point G. On both half-cycles of the error and reference voltages, electron flow was up through R_L, developing a positive voltage output at point E. The magnitude of the dc produced at point E in both instances was dependent on the amplitude of the ac error signal, and the polarity of the dc was dependent on the phase of the ac error signal. C_L is used to filter the pulses and provide smooth dc.

Triode Phase Detectors

A phase detector that uses npn transistors and also provides amplification of the error signal in addition to phase detection is depicted in figure 1-13. In this circuit, the collectors of the transistors are supplied with the ac reference voltage in such a manner that the collector voltages are in phase. For the purpose of explanation, assume that no error signal is present at T2. When the collectors of Q1 and Q2 are positive, the two transistors conduct equally. The collector current that flows sets up magnetic fields in the dc motor exciter windings that are equal and opposite; therefore, the fields cancel and produce no output. When the collector voltages are on a negative half cycle, C and C_2 discharge through their respective exciter windings to maintain a constant direct current through the windings.

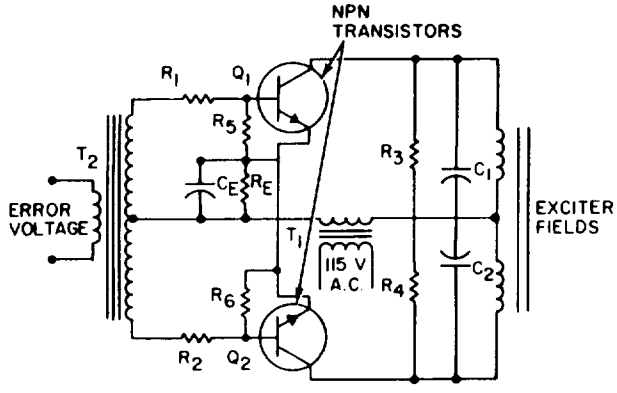


Figure 1-13.-Triode phase detector.

If an error signal is introduced into the primary of T2 with a phase relationship that causes the base of Q1 to be positive at the same instant that the collector of Q2 is positive, the following conditions exist:

1. On this half cycle the conduction of Q1 is increased above its no-error signal condition.
2. The heavier collector current causes a stronger field to be created in the upper exciter winding.
3. At this same instant, since the base of Q2 is on a negative half cycle, its average conduction is reduced to a level below that of its no-error signal condition.
4. The lower level of collector current causes a weaker field to be produced in the lower exciter winding.
5. Since the magnetic fields produced in the exciter windings are no longer of equal amplitude, they no longer cancel each other.
6. The exciter produces an output voltage of a polarity controlled by the polarity of the resultant field and of an amplitude controlled by the relative strength of this resultant field.
7. The exciter output causes the proper mechanical actions necessary to reduce the amplitude of the error to zero.
8. As the error signal is reduced to zero, the current conduction through Q1 and Q2 is again balanced. Also, the exciter fields are equal and opposite, canceling each other, reducing the exciter output to zero, and stopping the mechanical action. Resistors R1 and R2 prevent excessive base current when the error angle is large.

SPECIAL CIRCUITS

It has been shown how a servo control amplifier may have provisions for changing a dc error signal to an ac signal, and how an ac error signal may be detected to supply a dc voltage to a servomotor or controller. In the following paragraphs, other special amplifier circuits are discussed.

Two-Stage DC Servo Control Amplifier

If somewhat more power is required by the servomotor than can be supplied by the servo

amplifier (fig. 1-13), a push-pull dc amplifier can be inserted between the phase-sensitive transistors and the servomotor. In the schematic diagram (fig. 1-14), the output of the phase detector transistors is now taken across the parallel RC networks in the collector circuit.

The bias source, Ecc, for the dc amplifier is connected with its positive terminal on the base side. This positive voltage subtracts from the highly negative voltage across the capacitor to give a resulting negative voltage, which allows the transistor to operate on the linear portion of its characteristic curve.

When there is no signal input from the error detector, the collector currents of the phase-sensitive rectifiers are equal. The outputs of Q1 and Q2 are applied to the base of Q3 and Q4, respectively. Equal output from Q1 and Q2 causes equal currents to flow in Q3 and Q4. With R5 and R6 equal in resistance and current, the voltage across the motor is zero. Consequently, the motor does not turn.

For an analysis of a signal output from the error detector, assume that the error signal makes the base of Q1 positive and the base of Q2 negative. The collector current of Q1 increases and the collector current of Q2 decreases. An increasing collector current in Q1 increases the charge on capacitor C1; conversely, a decreasing collector current in Q2 decreases the charge on capacitor C2. As a result of the change in error signal, the voltage on the base of Q3 is now more negative than the voltage on the base of Q4. This increased negative voltage on the base of Q3 decreases its collector current and the voltage e3 decreases. The decreased negative voltage on the base of Q4 increases its collector current, and the voltage e4 increases. As a result, a voltage difference appears across the motor armature and the motor rotates.

When the output signal from the error detector reverses in phase, the sequence of events that follow causes the motor to reverse its direction of rotation.

Magnetic Amplifiers as Servo Control Amplifiers

The servomotor used in conjunction with the magnetic amplifier shown in figure 1-15 is an ac type. The uncontrolled phase may be connected in parallel with transformer T1 by using a phase shifting capacitor, or it may be connected to a different phase of a multiphase system. The controlled phase is energized by the magnetic amplifier, and its phase relationship is determined by the polarity of the dc error voltage.

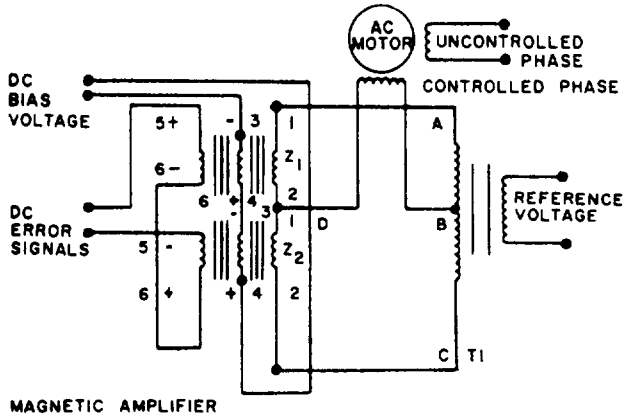


Figure 1-15.-Magnetic amplifier servo control amplifier.

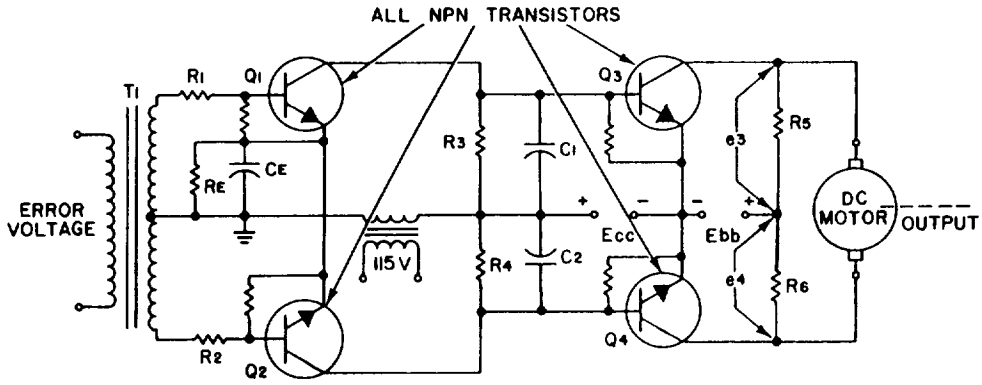


Figure 1-14.-Two-stage dc servo control amplifier.

The magnetic amplifier consists of a transformer (Tl) and two saturable reactors, each having three windings. Note that the dc bias current flows through a winding of each reactor, and the windings are connected in series-aiding. This bias current is supplied by a dc bias power source. A dc error current also flows through a winding in each reactor; however, these windings are connected in series-opposing.

The reactors Z_1 and Z_2 are equally and partially saturated by the dc bias current when no dc error signal is applied. The reactance of Z_1 and Z_2 are now equal, resulting in points B and D being at equal potential. There is no current flow through the controlled phase winding.

If an error signal is applied, causing the current to further saturate Z_2, the reactance of its ac winding is decreased. This current through Z_1 tends to cancel the effect of the dc bias current and increase the reactance of its ac winding. Within the operating limits of the circuit, the change in reactance is proportional to the amplitude of the error signal. Hence, point D is now effectively connected to point C, causing motor rotation. Reversing the polarity of the error signal causes the direction of rotation to reverse, since point D is effectively connected to point A.

The basic magnetic servo amplifier discussed above has a response of approximately 6 to 20 Hz. In some applications, this delay would be excessive, creating too much error. However, this delay can be reduced to about 1 Hz by using special push-pull circuits.

Amplifier Integrator

A servo system in a steady-state condition will have a constant positional displacement between input and output, which is called the error. The only way to reduce this error is to increase the drive torque. Thus, a new signal must be introduced that is related to the error. The error is not changing; therefore, it cannot be a derivative signal, nor can it be proportional to the error, because it would then decrease as the error decreases and a new condition would be met without removing the error. The only alternative is to produce a signal proportional to the integral of the error. Then, if a torque proportional to the time integral of the error is added to the normal torque that is proportional to the error, the error will eventually be reduced to zero. A circuit that is used for this purpose is called an amplifier integrator.

A simple and commonly used integrator consists of two circuit elements: a resistor and capacitor. (See fig. 1-16.) The voltage across the capacitor is proportional to the integral of the charging current. It can be explained by considering that the voltage across a capacitor is

$$E = \frac{Q}{C}.$$

For any given capacitor (C), the voltage depends directly on the charge (Q), which is the imbalance of electrons on the two capacitor plates. The amount of this charge depends on the current flow and the time that this flow exists.

Because the voltage is proportional to the integral of the charging current, it allows the RC circuit to be used as an integrator output. Provision must be made to supply a charging current that is proportional to the input information. The purpose of the resistor is to produce this proportional current from an input signal voltage (e_i). At the instant this voltage is applied, the charging current becomes

$$i = \frac{e_i}{R}.$$

Unfortunately, this proportionality does not continue to exist. As the capacitor becomes charged, the capacitor voltage opposes the charging current, and the charging current becomes less proportional to the input signal. This results in an error in the output. The ideal output for a constant input signal is a steadily increasing output. This steady increase is attained only when the signal voltage is first applied and the capacitor has not become appreciably charged.

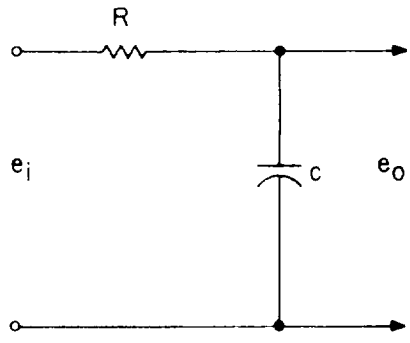


Figure 1-16.-Simple integrator.

A remedy to this error in the RC integrator is to use a circuit with a long time constant. Such a circuit delays the charging of the capacitor. The result is a more accurate integration of an input signal. The ideal output would be a perfect triangular wave. Although a long time constant produces more accurate results, it also provides a much lower output for the same input signal, Better integration is possible by the use of a high gain, feedback amplifier.

An amplifier integrator is illustrated in figure 1-17. The circuit arrangement uses a high gain amplifier and is known as the Miller integrator. The amplifier produces an output that is not limited by the input signal as it is in the simple RC integrator. The amplifier also supplies any energy that is required in the output. The function of the input signal is to control the charging current.

The operation can be explained by assuming a constant input, as shown in figure 1-17, view A. At the start, assume the initial condition is zero, that is,

$$e_i = e_g = e_o = 0.$$

Also assume that the capacitor is discharged. The positive voltage to be integrated, e_i, is then applied. The capacitor charges with a polarity as shown, since electrons are attracted from the left plate. The charging path is shown in figure 1-17, view B.

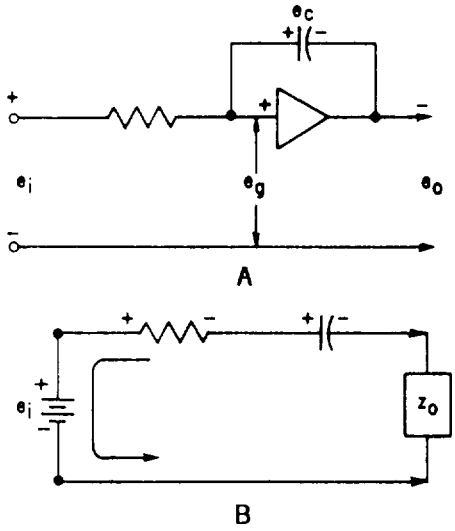


Figure 1-17.-Amplifier integrator.

A voltage measured at the amplifier input, e_g, tends to rise in the positive direction since this point is directly coupled to e_i. However, this rise tends to be opposed by the degenerative feedback voltage from the output. The output will be $-Ae_g(e_o)$. The letter *A* stands for the amplifier gain. The minus sign indicates that the output polarity or phase is opposite to the input. The output changes *A* times faster or steeper than e_g. The output voltage is negative and aids the charging of the capacitor.

For a certain input voltage, the charging current is limited to a particular value that tends to keep e_g practically zero. If the current should exceed this value, e_g would decrease a small amount due to the increased voltage drop across R. The e_o would decrease, and the charging current would decrease to the original value. If the initial charging current should decrease, the opposite action would occur. The value of the charging current is therefore stabilized to a specific value proportional to the input voltage. This eliminates the error caused by e_i and the charging current not remaining proportional in the fundamental RC integrator.

This constant charging current must be produced by e_o despite the fact that the steadily increasing capacitor voltage opposes the charging current. To do this, e_o must also steadily increase. This steady increase in e_o is exactly the integrator output voltage desired for a constant signal input.

Similar action would be produced for a condition in which the input signal suddenly became negative. Polarities would then be in reverse to those shown in the example given. Remember that simple examples are used for explanation on the assumption that the desired result will also be produced for a more complicated signal input. Removal of e_i would produce little effect upon the output that existed at that instant, since the amplifier output would oppose the tendency for the capacitor to discharge.

The limits for e_o are determined by the amplifier and not by e_i or the range of e_g. The output range would be designed to produce an increasing output for any probable input amplitude and period of application. The exception to this would be an integrator that was designed to function also as a limiter.

OUTPUT DEVICES

The output of the servo control amplifier is fed to an output device. The functions of this

device, usually a servomotor, are to supply torque, power, and dynamic characteristics required to position the servo load. Ideally, the power device should require small power from the control amplifier, accelerate rapidly, be of small size and weight, be of lasting endurance, have small time lags, and have an adequate speed range. In aircraft weapons systems, the electric motor is most frequently used as an output device. However, electromagnetic clutches, hydraulic devices, and pneumatic devices are also used.

Electric Motors

In aircraft weapons systems, electric motors are primarily used to drive the servo load. The type of electric motor used within a particular equipment is determined by power factors such as type of power available, output power, speed range, inertia, and electrical noise.

ALTERNATING-CURRENT MOTORS.— Alternating-current motors are frequently used in low power servo applications because of their sparking and rapid response. However, they have a disadvantage of having a narrow speed range characteristic. The theory of operation of ac motors is discussed in Navy Electricity and Electronics Training Series (NEETS), module 5. We briefly discuss the types of motors used with servo systems in this chapter.

The two-phase induction motor is the most widely used ac servomotor. The stator of the motor consists of two similar windings that are positioned at right angles to each other. The rotor may be wound with short-circuited turns of wire or it may be a squirrel cage rotor. The squirrel cage rotor is the type most frequently encountered. It is made up of heavy conducting bars, which are set into armature slots, the bars being shorted by conducting rings at the ends.

Two ac voltages 90 degrees out of phase must bc applied to the stator windings for the motor to turn. These out-of-phase voltages generate a rotating magnetic field, which induces a voltage in the rotor. This induced voltage generates a magnetic field in the rotor that is displaced 90 degrees from the stator magnetic field. The interaction of these two magnetic fields causes the armature to rotate.

As stated previously, the voltage to the two stator windings must be 90 degrees out of phase to cause the rotor to turn. The direction of rotation is determined by the phase relationship of the stator windings, which, in turn, is determined by the servo error detector. One phase is connected directly to one of the stator windings while the other phase is used to energize an error detector. The resulting error voltage is either in phase or 180 degrees out of phase with the signal applied to the error detector. This will cause the controlled phase to either lead or lag the uncontrolled phase by 90 degrees.

Most induction motors have low starting torque and high torque at high speed. For servo applications, it is desirable to have high starting torque so that the system may have a low time lag. This may be accomplished by increasing the armature resistance with the use of materials such as zinc for the conducting bars. This increased torque at low speed results in decreased torque at high speed. However, increased stability of the servo system is a desirable result of this change.

Split-phase ac motors are similar to the two-phase induction motor. It differs only in that a phase shifting network is used to shift the phase of the voltage supplied to one of the windings by 90 degrees. This is usually accomplished by connecting a capacitor in series with the uncontrolled winding of the stator. Direction of rotation and reversal is accomplished in the same manner as in the two-phase motor discussed above.

Other types of motors that may be used with an ac power supply are shaded pole, universal, and repulsion motors. They use various methods of obtaining rotation reversal. However, they are seldom found in aircraft weapons systems.

DIRECT-CURRENT MOTORS.— Direct-current motors have an advantage of having higher starting torque, reversing torque, and less weight for equal power than ac motors.

Series motors are characterized by their high starting torque and poor speed regulation with a change in torque. Higher torque can be obtained on reversal of direction with a series motor than any other type. However, it is a unidirectional motor and requires special switching circuits to obtain bidirectional characteristics. This is normally done by switching either the armature or field connections, but not both.

A variation of the series motor that has bidirectional characteristics is the split-series motor. The motor has two field windings on its frame, only one of which is used for each direction of rotation. This reduces the number of relay contacts required for reversing by one-half. This double winding also reduces the torque

capabilities of the motor as compared to a straight-series motor wound on the same frame.

The most frequently used dc servomotor is the shunt motor. Its direction of motion is controlled by varying the direction of flow of either the armature or field current. The uncontrolled current is usually maintained constant to preserve a linear relationship between the motor output torque and the voltage or current input. The field windings are usually two differentially wound coils to aid in direction control of the field current by the servo control amplifier. The field current is usually controlled with receiving type vacuum tubes. The larger armature currents require thyratrons or generators as current regulators, but are not normally found in aircraft weapons systems.

Magnetic Clutches

Any device using an electrical signal that may be used to control the coupling of torque from an input shaft to an output shaft is a magnetic clutch. This coupling may be accomplished by the contact between friction surfaces or by the action of one or more magnetic fields. A magnetic clutch is used only to couple the input torque to the output shaft. Thus, it is capable of controlling large amounts of power and torque for its size and weight. The magnetic clutch may be used with a large flywheel driven at high speed by a small motor. This allows the flywheel to impart very large acceleration to the load when the magnetic clutch is energized.

There are two distinct types of magnetic clutches. Some transmit torque by physical contact of frictional surfaces. Others use the action of magnetic flux produced by two sets of coils, or one set of coils and induced eddy currents resulting from rotating the one set of coils near a conducting surface. The eddy current type of clutch offers smoother operation and has no problem of wear because of friction. Both types have suitable control characteristics and are found in servomechanisms.

HYDRAULIC DEVICES

Hydraulic components used in servomechanisms are frequently found in aircraft weapons systems. Hydraulic power devices, such as motors and associated control valves, have the advantage of a response much faster than the best electric motors and equal to that of a magnetic clutch system. They also require a minimum of maintenance, have very high accuracy, and are adapted to heavy loads.

The essential components of a hydraulic system are as follows:

1. A source of high-pressure oil and sump to receive discharge oil
2. A control valve and means of employing an actuating signal
3. An actuator (motor or cylinder)

The theory of operation of a hydraulic system is discussed in *Fluid Power,* NAVEDTRA 16193 (series).

The source of high-pressure oil serves as a source of power to operate the actuator. However, this source of power is controlled by the control valve. This valve is actuated by the output from the servo control amplifier. This control is normally accomplished by feeding the error signal to a solenoid-controlled valve. However, the error signal could be used to drive an electric motor, which, in turn, would actuate the control valve. The actuator is usually in the form of an axial motor, which must be a reversible and variable speed type. Some applications may employ a cylinder where linear motion is required for positioning.

SERVOMECHANISM OSCILLATION

Learning Objective: *Identify factors affecting servomechanism oscillations to include damping, integral control, and the relationship of gain, phase, and balance.*

In aircraft weapons systems, servomechanisms are used for various functions and must meet certain performance requirements. These requirements not only concern such things as speed of response and accuracy, but the manner in which the system responds in carrying out its command function. All systems contain certain errors; the problem is keeping them within allowable limits.

As discussed previously, the servomotor must be capable of developing sufficient torque and power to position the load in a minimum of time. The servomotor and its connected load have sufficient inertia to drive the load past the point of command position. This overshooting results in an opposite error voltage, reversing the direction of rotation of the servomotor and the

load. The servomotor again attempts to correct the error, and again overshoots the point of correspondence, with each reversal requiring less correction until the system is in correspondence. The time required for the oscillations to die out determines the transient response of the system and can be greatly reduced by the use of damping.

DAMPING

The function of damping is to reduce the amplitude and duration of the oscillations that may exist in the system. The simplest form of damping is viscous damping. Viscous damping is the application of friction to the output load or shaft that is proportional to the output velocity. The amount of friction applied to the system is critical and will materially affect the results of the system. When just enough friction to prevent overshoot is applied, the system is said to be critically damped. When the friction is greater than that needed for critical damping, the system is over-damped. However, when damping is slightly less than critical, the system is said to be slightly underdamped, which is usually the desired condition. The application of friction absorbs power from the motor and is dissipated in the form of heat.

A pure viscous damper would absorb an excessive amount of power from the system. However, a system having some of the characteristics of a viscous damper with somewhat less power loss is used in actual practice. The first of this type of damper to be discussed uses a dry friction clutch to couple a weighted flywheel to the output drive shaft. A flywheel has the property of inertia, which maybe defined as that property of matter by which it will remain at rest or in uniform motion in the same straight line or direction unless acted upon by some external force.

Since the flywheel is coupled to the output shaft with a friction clutch, any rapid change in velocity of the output member causes the clutch to slip. This slipping effectively disconnects the flywheel, instantaneously, but allows sufficient power to be coupled to the flywheel to overcome its inertia. As the inertia is gradually overcome, the flywheel gains speed and approaches the velocity of the output member. As the point of correspondence is neared and the error signal is reduced, the inertia of the flywheel gives up power to the system, causing the load to increase its overshoot. When the system attempts to correct for this overshoot, the inertia of the flywheel adds to the output load, reducing the effect of the correcting signal. The effect dampens the oscillations in the system, reducing its transit time.

Another type of damper used is the eddy current damper. This damper uses the interaction of induced eddy currents and a permanent magnet field to couple the output shaft to a weighted flywheel.

The effect of damping is shown in figure 1-18. The solid line shows the action of the load without damping. The time required to reach a steady-state condition without damping should be noted. This time is greatly reduced although the initial overshoot is increased.

As shown in figure 1-18, a viscous damper effectively reduces transient oscillations, but it also produces an undesired steady-state error.

How well the load is controlled is a measure of the steady-state performance of a servo system. If the load is moved to an exact given position, then the servo system is said to have perfect steady-state performance. If the load is not moved to an exact position, then the system is not perfect and the difference in error is expressed as the steady-state error. Steady-state error may be either velocity lag or position error. Velocity error is the steady-state error due to viscous drag during velocity operation. Position error is the difference in position between the load and the position order given to the servo system. Since the friction damper absorbs power from the system, its use is normally limited to small servomechanisms.

To overcome the disadvantages of the viscous dampers and still provide damping, error-rate damping is used. This type of damping consists of introducing a voltage that is proportional to the rate of change of the error signal. This voltage is fed to the servo control amplifier and combined with the error signal. Figure 1-19 shows the

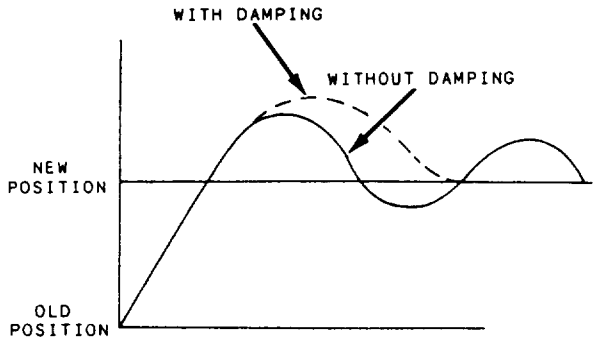


Figure 1-18.-Effect of friction damper.

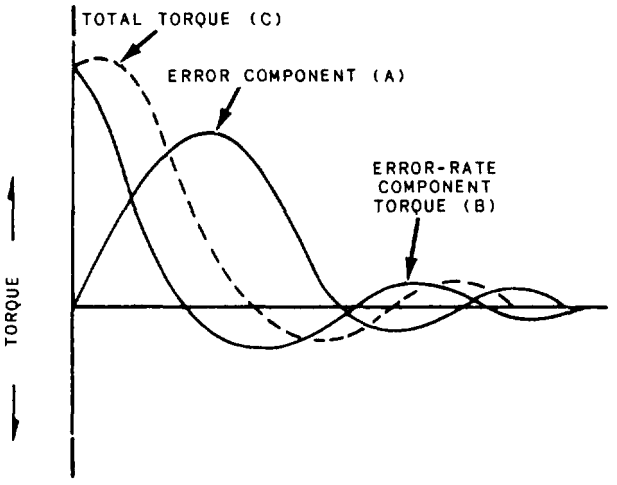


Figure 1-19.-Torque variations using error-rate damping.

effect of error-rate damping on the torque output of the servomotor. Curve A shows the torque resulting from the error voltage, curve B shows the torque resulting from the error-rate damper, and curve C depicts the resultant of curves A and B. It should be noted that torque resulting from the damper increases the total torque as long as the error component is increasing. Once the error component starts to decrease, the error-rate damper produces a torque in an opposite direction, reducing the transit time of the system.

There are two methods of generating an error-rate voltage normally found in aircraft weapons systems—tachometer and electrical networks. The tachometer error-rate damper uses a device that is essentially a generator having an output voltage proportional to its shaft speed. The tachometer is connected to the shaft of the output member, giving a voltage proportional to its speed. The output voltage is fed to a network that modifies this voltage so that it is proportional to a change in input voltage. This voltage is fed back to the servo control amplifier and added to the error signal, as shown in figure 1-19.

Electrical networks used for error-rate damping consist of a combination of resistors and capacitors used to form an RC differentiating network. For a detailed explanation of RC circuits, refer to Navy Electricity and Electronics Training Series (NEETS), module 2, NAVEDTRA 172-02-00-85. These networks, sometimes referred to as phase advance or lead networks, vary in design, depending on the type of error signal. However, in practice, networks are normally limited to the dc type (fig. 1-20) because

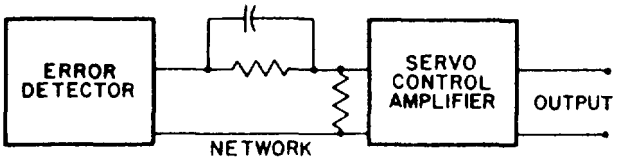


Figure 1-20.-Error-rate stabilization network.

of the unstable results that would be caused by a small change in frequency of the power source. An ac system may use a dc network by first using a demodulator (detector) prior to the network. However, the output of the network must be modulated for use in the remainder of the ac system. Like the tachometer, the output of the network is fed to the servo control amplifier.

INTEGRAL CONTROL

Servomechanisms used in aircraft weapons systems are sometimes required to follow an input function, the magnitude of which changes at a constant rate with time, such as an antenna system tracking a target. Thus, if the input is the angle of a shaft, the velocity of the shaft may be constant for a substantial percentage of time. The servomechanism may be required to respond to this type of input with substantially zero error. The error that characterizes the servo response to a constant velocity input is known as the velocity error.

To correct for velocity error or an inaccuracy due to a steady-state error, an integral control may be used. This control modifies the error voltage in such a manner that the signal fed to the servo control amplifier is a function of both the amplitude and time duration of the error signal. This is accomplished by the use of a variable voltage divider, whose output is increased with time for a constant input. As in all voltage dividers, the output is only a portion of the input that effectively reduces the amplitude of the error signal. To compensate for the loss of amplitude, additional amplification must be used either in the form of a preamplifier or a higher gain servo control amplifier. With the overall gain of the system now increased to give a normal output for transient error signals, small velocity or steady-state error signals of long duration will result in somewhat increased output to the servomotor due to the action of the integral control.

The integral control (fig. 1-21) consists of a combination of resistors and capacitors connected to make an integrator circuit for a dc error signal. The value of the components are such that the capacitor does not have sufficient time to change with fluctuations in error voltage. Only that portion of the transient error signal developed across R1 is impressed on the amplifier. However, with a velocity error or steady-state error of longer duration, the capacitor (C1) charges, increasing the amplitude of the amplifier input.

Networks shown in figure 1-21 are not limited to dc systems, as a demodulator maybe used prior to the integrator and its output modulated for easier amplification.

GAIN, PHASE, AND BALANCE

The overall system gain has a most important effect on the servomechanism response characteristics and is one of the more easily adjustable parameters in electronics servo controllers. Increasing the system gain reduces the system velocity errors and those steady-state errors resulting from restraining torques on the servo load or misalignment in the system. An increase in system gain also increases the speed of response to transient inputs. Excessive gain always decreases the rate at which oscillatory transients disappear. Continued increase in the system gain eventually produces instability.

Servo systems using push-pull amplifiers must be balanced to ensure equal torque in both directions of the servomotor. This adjustment should be checked periodically as a change in value of a component may cause an unbalanced output. Balancing is accomplished by adjusting the system for zero output with no signal applied.

A phase control is included in some servo systems using ac motors. The two windings of the ac servomotor must be energized by ac signals that are 90 degrees apart. A phasing adjustment is normally included in the system to compensate for any phase shift in the amplifier circuit, resulting in unstable operation of the system. This adjustment may be located in the control amplifier, or in the case of a split-phase motor, it may be in the uncontrolled winding.

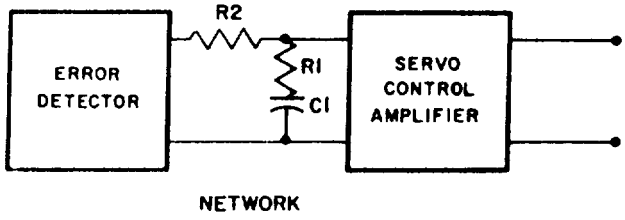


Figure 1-21.-Integral stabilization network.

ZEROING SYNCHRO UNITS

Learning Objective: *Recognize the importance of zeroing transmitting and receiving synchro units.*

In this chapter, we have stressed the importance of accuracy with servomechanisms. In any

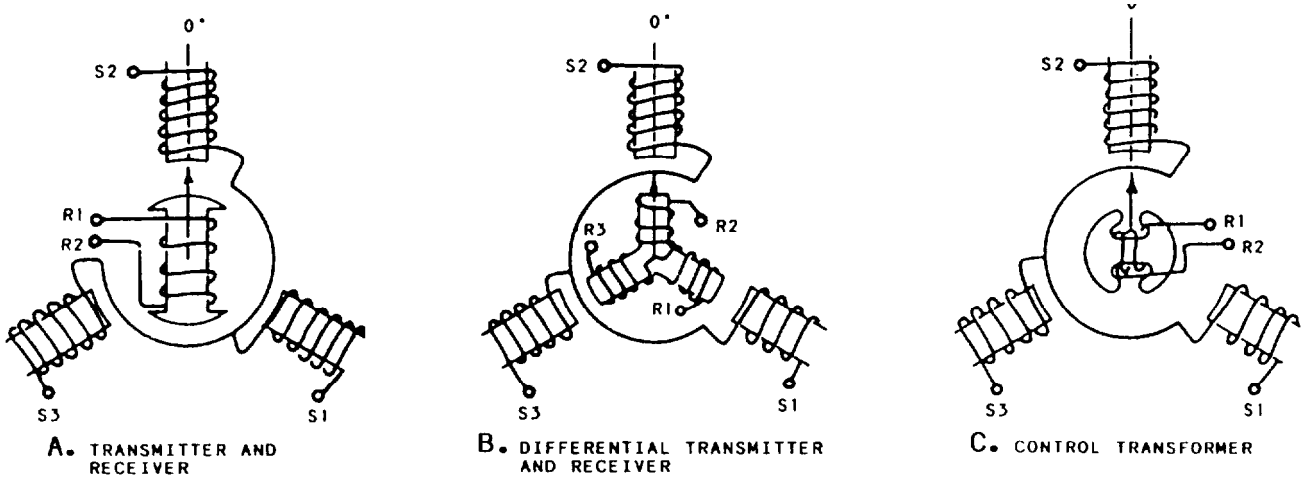


Figure 1-22.-Synchro electrical zero positions.

servomechanism using synchro units, it is also very important that the units be zeroed electrically (fig. 1-22, view A).

For a synchro transmitter or receiver to be in a position of electrical zero, the rotor must be aligned with S2, the voltage between S1 and S3 must be zero, and the phase of the voltage at S2 must be the same as the phase of the voltage at R1.

The most common methods of zeroing synchro transmitters and receivers are the ac voltmeter method and the electrical lock method. The method used to zero a synchro depends upon how the synchro is used. Where the rotor is free to turn, the electrical lock method can be used. This is accomplished by connecting S1 and S3 to R2 using a jumper wire and connecting S2 to R1 (fig. 1-23). When power is applied, the rotor will position itself in the zero position. After the synchro is zeroed, the pointer is adjusted to indicate zero.

The great majority of synchros used in aviation systems have their rotors gear driven or mechanically coupled to a driving member. In these cases it is necessary to use the ac voltmeter method, zeroing the synchro by rotating the stator or housing until its electrical zero is reached. Before you zero the synchro, the mechanical unit that positions the synchro must be set to its indexing or ZERO position. This is done by aligning the unit to its index and installing its indexing pins in the holes provided for this purpose. The pins hold the unit to its index and keep it from moving.

The ac voltmeter method is done by connecting the meter and jumper wires, as shown in figure 1-24, view A. Rotate the energized synchro until a zero reading is obtained on the voltmeter. Since rotor positions of 0 degree and 180 degrees produce this zero reading, it is necessary to determine if the phase of S2 is the same as that of R1. Make the connections as shown in figure 1-24, view B. If the proper polarity relationship exists, the voltmeter indicates less than the excitation voltage being applied to the rotor. If the indication is greater than the rotor excitation voltage, the rotor (or stator) must be rotated 180 degrees and the previous step must be performed again.

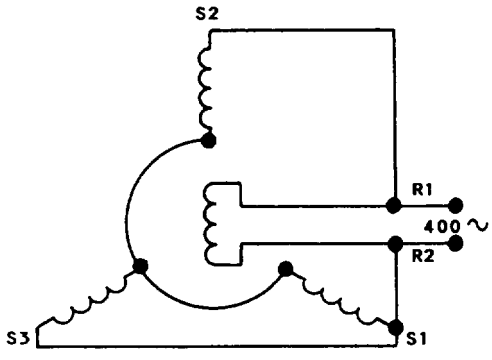


Figure 1-23.-Electrical lock method of zeroing a synchro.

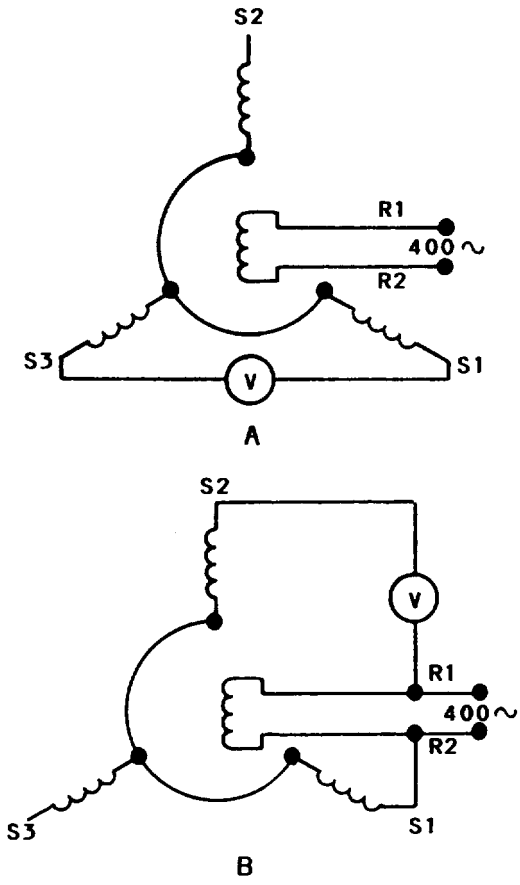


Figure 1-24.-Ac voltmeter method of electrically zeroing synchro receiver or transmitter.

DIFFERENTIAL TRANSMITTER

The electrical zero position of a synchro differential transmitter or receiver is when the three windings of the rotor are in correspondence with their respective stator windings and their respective voltages are in phase (fig. 1-22, view B).

Because the differential transmitter synchro is normally used to insert a correction into a synchro system, it is usually driven either directly or through a gear train. Before you zero the differential transmitter synchro, the unit whose

position the differential synchro transmits should first be zeroed. After this has been accomplished, connect the differential synchro as shown in figure 1-25, view A. Turn the synchro in its mounting until the voltmeter shows a minimum indication. After you complete this step, make the connections shown in figure 1-25, view B. Again, turn the synchro slightly in its mounting until a minimum voltage is indicated by the voltmeter.

DIFFERENTIAL RECEIVER

Electrical zero for a differential receiver is illustrated in figure 1-22, view B. To zero a differential receiver synchro, make the connections shown in figure 1-26. As soon as the power is applied to the synchro, the rotor assumes a position of electrical zero. The dial can then be set at zero, and the unit reconnected to its circuit.

CONTROL TRANSFORMER

The synchro control transformer is normally zeroed by using the ac voltmeter method. You should remember that the electrical zero position of the control transformer is 90 degrees from that of a receiver, since the rotor winding must be perpendicular to the stator's resulting magnetic field to have a zero output (fig. 1-22, view C). The coarse adjustment is made by connecting the meter and unit as shown in figure 1-27, view A. The rotor is rotated to give a minimum or null reading on the voltmeter. The final adjustment is made by connecting the unit as shown in figure 1-27, view B, and displacing the rotor a few degrees in both directions to determine the null or electrical zero position. Once the zero position has been determined, the unit must be locked, as discussed previously.

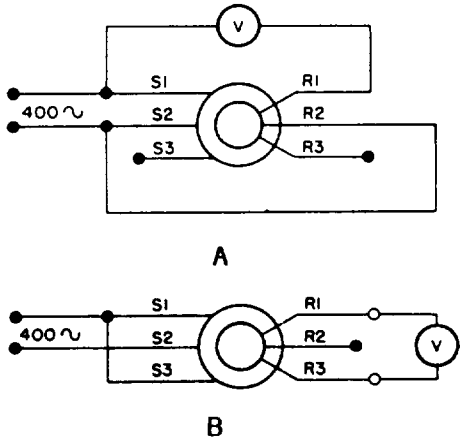


Figure 1-25.-Electrically zeroing a differential transmitter.

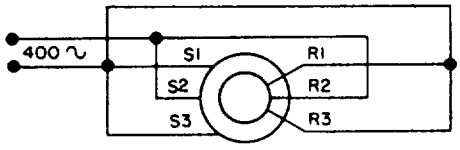


Figure 1-26.-Electrically zeroing a differential synchro receiver.

SYNCHRO ALIGNMENT SET TS-714/U

Learning Objective: *Recall the purpose and use of the synchro alignment set.*

The Synchro Alignment Set TS-714/U (fig. 1-28) is a portable, general-purpose test set used to check the alignment of synchros or resolvers. It can be used to align any 400 Hz synchro or resolver. In addition to its higher sensitivity, the test set has an additional advantage over the methods previously discussed because the test set can also supply excitation voltage for the synchro or resolver being aligned.

The test set (fig. 1-28) basically consists of a bandpass amplifier and power supply, a synchro

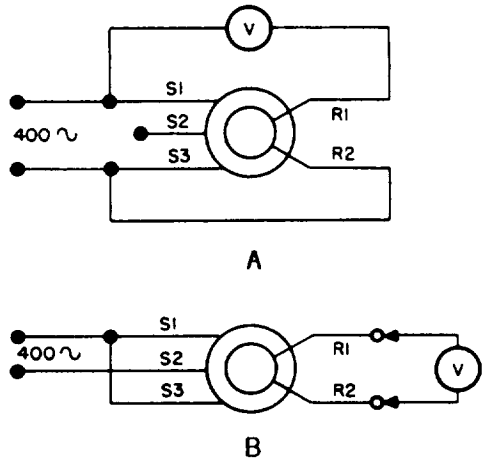
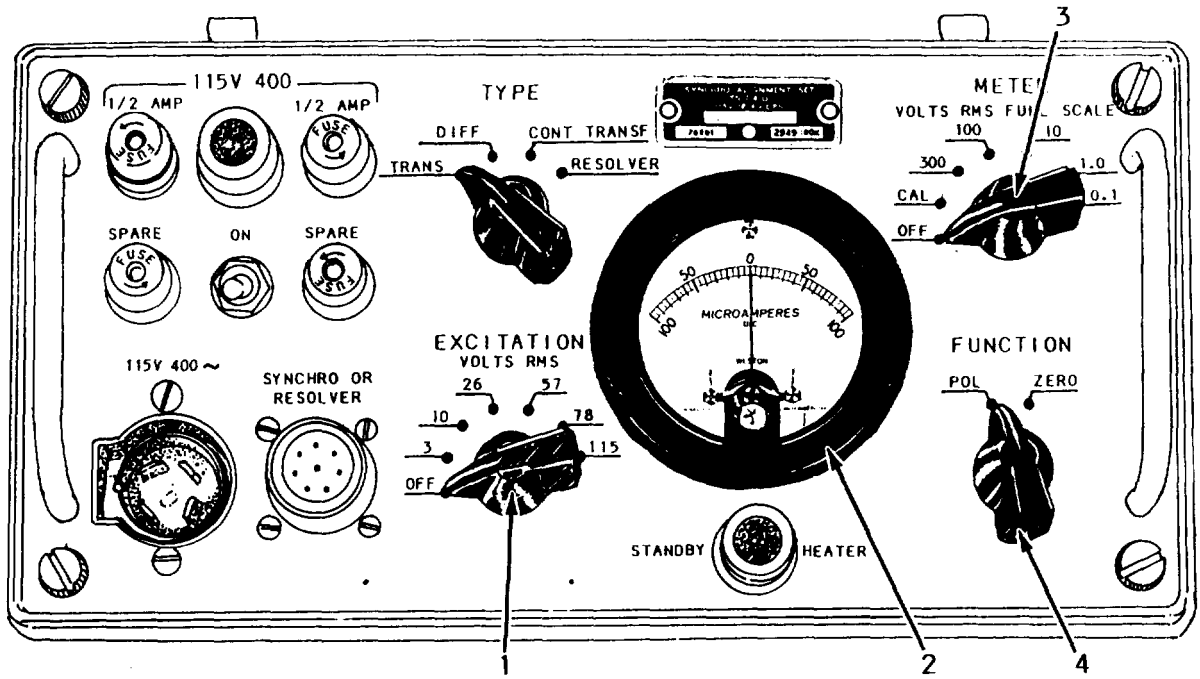


Figure 1-27.-Electrically zeroing a control transformer synchro.



1. **EXCITATION VOLTS RMS switch**
2. **MICROAMMETER**
3. **METER switch**
4. **FUNCTION switch**

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Figure 1-28.-Synchro Alignment Set TS-714/U front panel.

or resolver excitation supply with outputs from 3 to 115 volts rms (1) and switching circuits. The output voltages from the synchros or resolvers are applied to the amplifier, the output of which is fed to a phase sensitive detector circuit. The detector's output is metered by the microammeter (2). A meter switch (3) selects the meter sensitivity from 300 volts full scale to 0.1 volt full scale,

The meter has a ZERO center scale and indicates 0 when the synchro or resolver is adjusted to either of its two nulls. The synchro or resolver is adjusted to a null position with the function switch (4) in the ZERO position. When the null is reached, the function switch is switched to the POL position and a reading is taken from the meter. Then the function switch is returned to the ZERO position and the synchro is rotated 180 degrees to its opposite null. When the opposite null is reached, the function switch is again switched to the POL position and a note made of the reading. The correct null will be the one indicating the lowest reading with the function switch in the POL position. When the synchro is adjusted to this null, it is electrically zeroed with the correct polarity.

For detailed instructions on the use of the TS-714/U test set, consult *Operation and Service Instruction Manual,* NA 11-70-FAG-510.

ANTENNA POSITIONING SERVO SYSTEM

Learning Objective: *Explain the procedures for the application of servomechanisms to include positioning a radar antenna and supplying information to the weapons system.*

In this section, the application of a servomechanism to position a radar antenna and supply target information to the weapons system is discussed. However, before discussing the servo system, consider the scan pattern of a typical aviation fire control radar.

The antenna radiator and reflector form a conical pattern of circular symmetry with beam

dimensions, as shown in figure 1-29. The antenna assembly contains a spinner motor that rotates the beam about the antenna axis to produce a 7-degree conical scan. While the radar is in the search mode of operation, the rotating cone scans both horizontally and vertically, covering an area of 10 degrees vertically by 90 degrees horizontally (fig. 1-30). The search pattern may be positioned vertically from a positive 30 degrees to a negative 30 degrees by the antenna positioning level.

The operator normally observes the targets, identifies each as friend or foe, and determines which target, if any, to pursue. Since the antenna uses its 7-degree conical pattern only during track operation, some means must be provided for positioning the antenna on the selected target to begin the track operation. This is accomplished by bracketing the selected target with strobe lines. When the target has been selected and bracketed, a lock-on switch is depressed, positioning the antenna on the predetermined target, and placing the equipment in the automatic track mode of operation. The antenna is now positioned by the radar receiver output, keeping the target centered in the 7-degree beam.

A block diagram of a typical fire control antenna servo system is shown in figure 1-31. It should be noted that the azimuth channel of the antenna control system has been omitted, as its operation is similar to the elevation channel. Since the antenna servo system uses different components during search and track operation, the system used in each mode of operation is discussed separately.

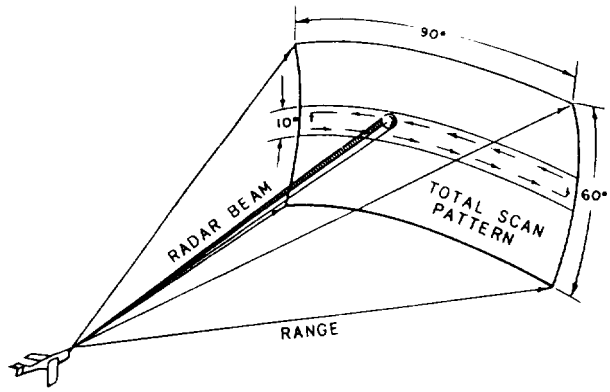


Figure 1-30.-Typical antenna scan pattern.

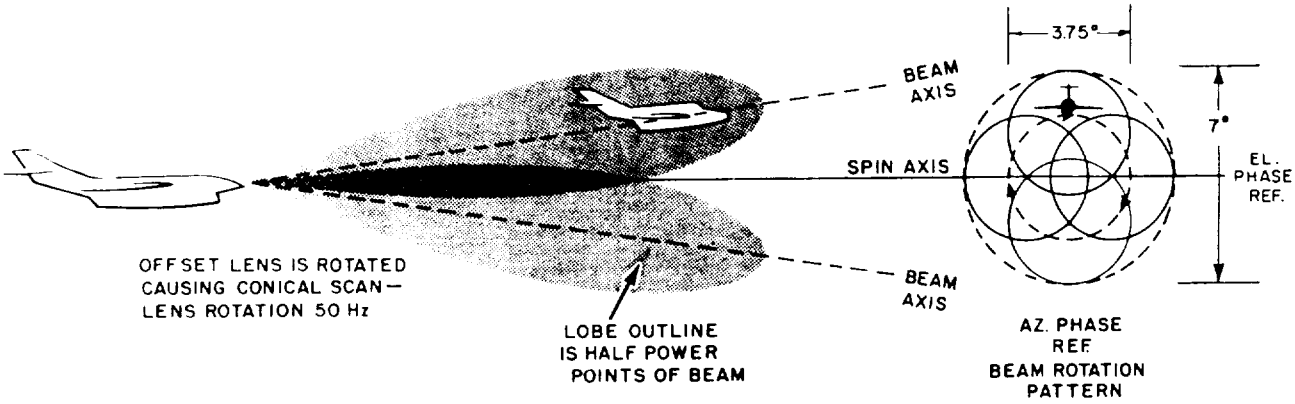
SEARCH OPERATION

The main components of the antenna servo system used during a search operation are as follows:

1. Error detector and its ac voltage source
2. Servo amplifier
3. Servomotor
4. Data transmission system

The ac generator supplies voltage to the input and feedback potentiometers of the balanced potentiometer error detector. However, the voltage fed to the input potentiometer is fed through a gyro space stabilizer and scan generator.

The function of the gyro space stabilizer is to cause the antenna to scan a selected area 90 degrees horizontally and 10 degrees vertically,



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Figure 1-29.-Antenna beam with conical scan.

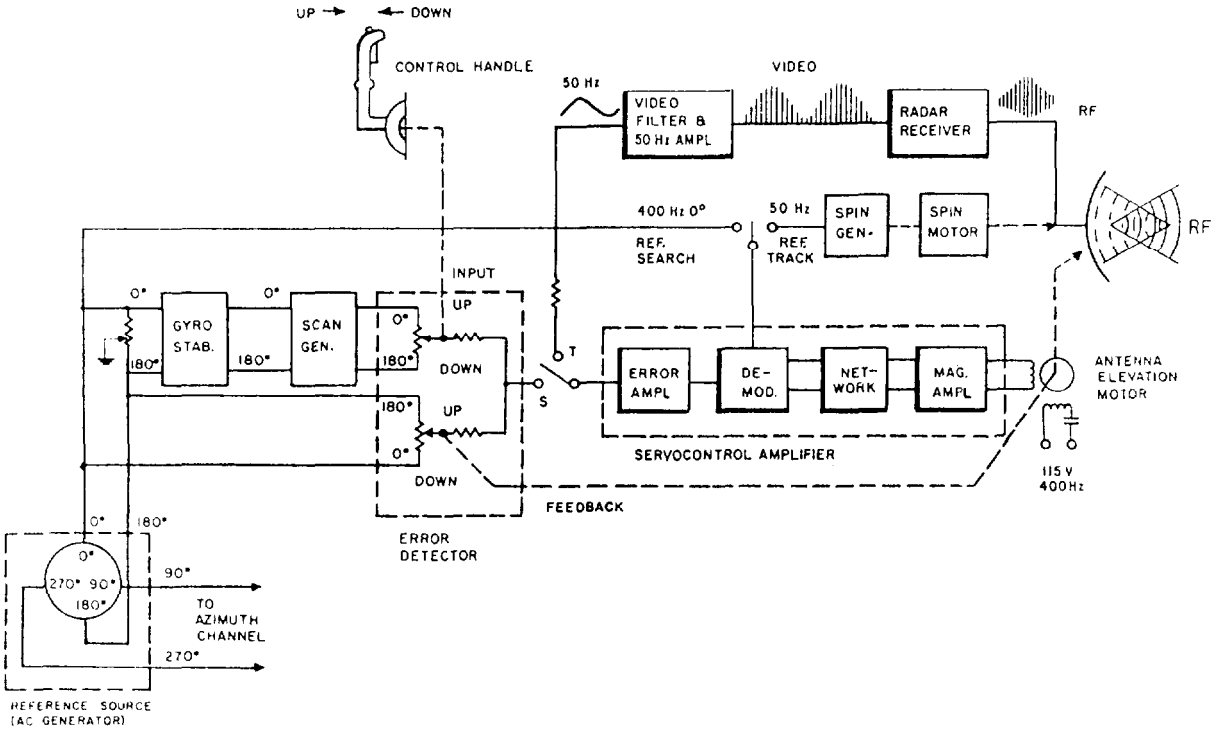


Figure 1-31.-Antennas elevation servo system functional block diagram.

regardless of any roll or pitch of the aircraft. As in all fire control equipment of this type, the amount of correction that can be made by the gyro space stabilizer is limited by the limits of the radar scanner. The output of the gyro space stabilizer is an ac voltage, the amplitude of which is a function of the roll and pitch of the aircraft. The principles of operation of gyros are discussed later in this manual.

The function of the vertical scan generator is to automatically position the antenna in the vertical geometric plane. Refer to figure 1-30. Note that the antenna scans horizontally and vertically. The scan generator provides the necessary voltage change to cause the antenna to change its angle of elevation by 3 degrees when the antenna reaches its azimuth limits.

The error detector has three inputs that are summed and compared against the antenna's position. The gyro space stabilizer and scan generator constitute two inputs by controlling the amplitude of the voltage supplied to the input potentiometer. The third input is the control handle, which positions the wiper contact of the input potentiometer. The output of the error detector is an ac voltage, whose amplitude and phase is determined by the voltages on the wipers of the potentiometers.

The error signal is fed to the servo amplifier, where it is amplified and compared with the phase of the reference voltage. The phase of the output voltage causes the servomotor to rotate in a direction reducing the error voltage.

The data transmission system is the mechanical linkage necessary to drive the wiper of the feedback potentiometer, indicating the actual position of the antenna in the vertical plane at all times.

TRACK OPERATION

The main components of the servo system employed during track operation are as follows:

1, Radar receiver and 50-Hz amplifier
2. Servo amplifier
3. Servomotor
4. 50-Hz spin generator

The radar receiver functions as the error detector, supplying a 50-Hz error voltage. Before

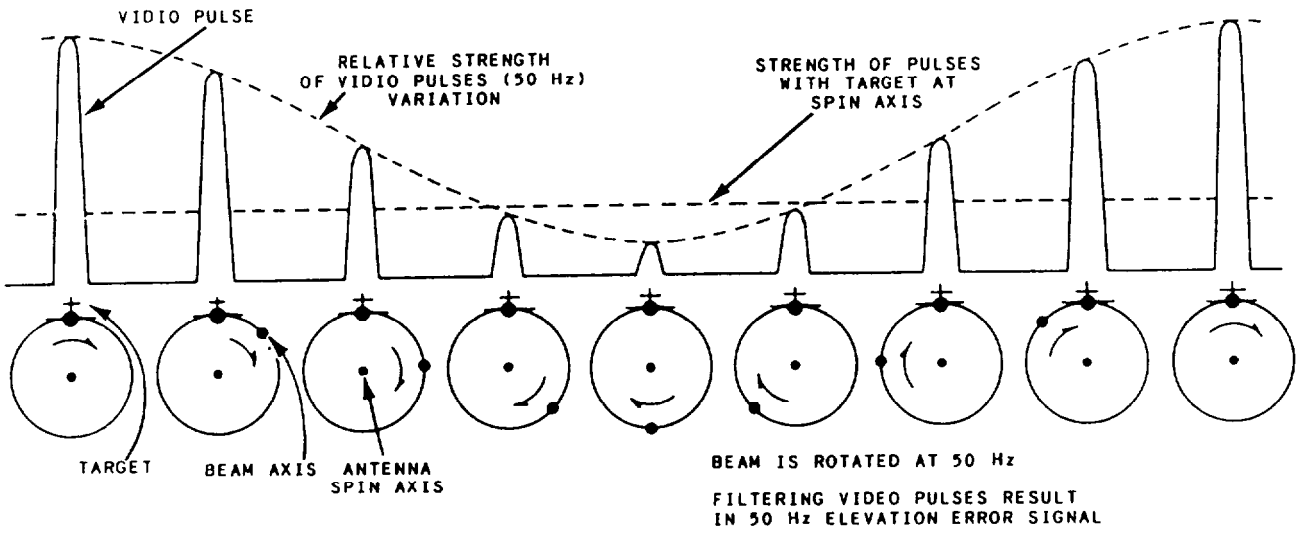
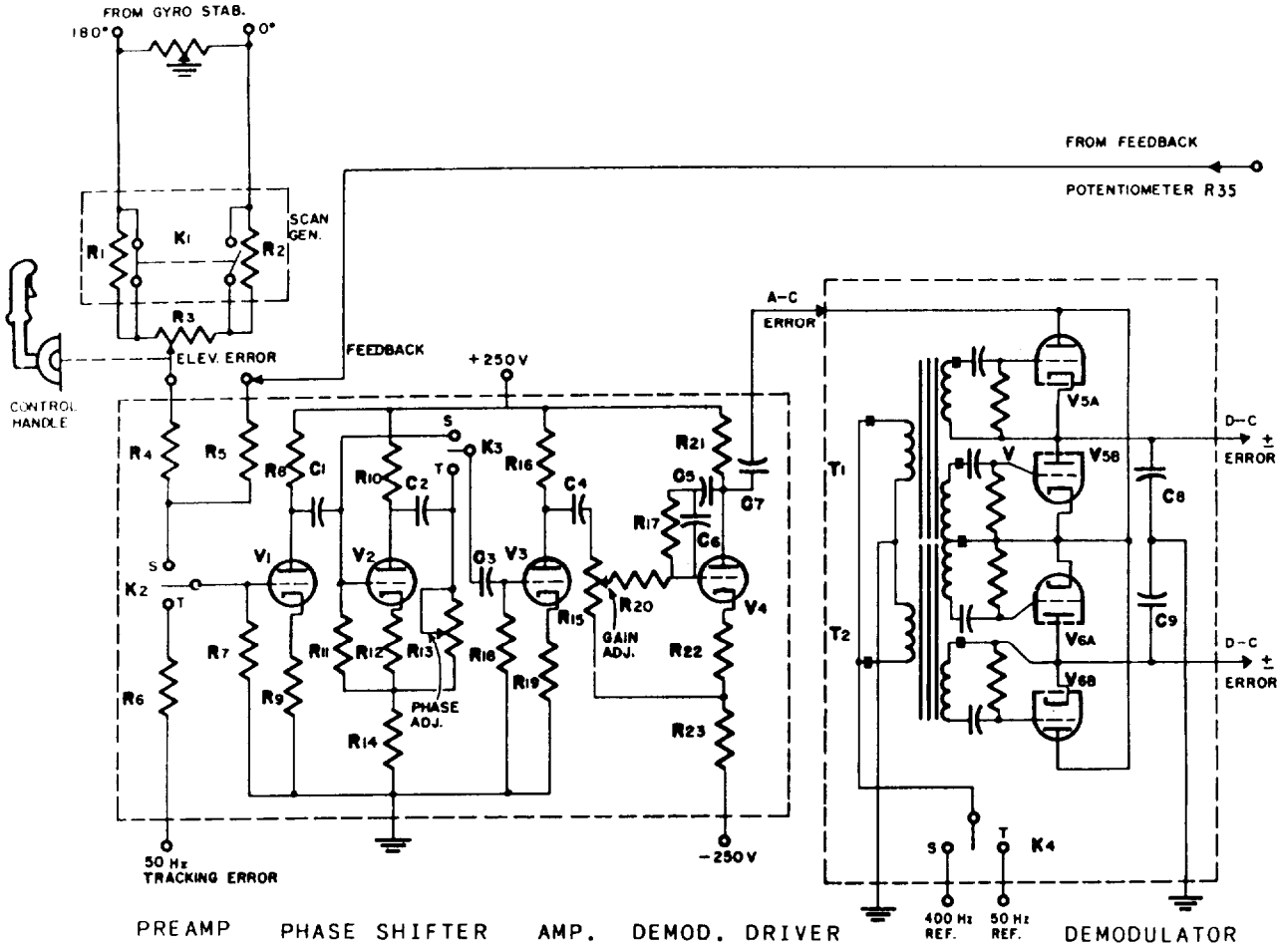


Figure 1-32.-Derivation of elevation error signal.



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Figure 1-33.-(A) Servo system schematic diagram.

discussing the other components of the system, first we must determine how the receiver provides the error signal. As stated previously, the antenna axis is centered approximately on a target prior to going into track operation. The antenna is rotating at 50 revolutions per second while the radar transmitter is transmitting a pulse of energy 450 times per second. When the antenna axis is pointing directly at the target, the target return and receiver video output remain at a constant level. However, if the target were above the antenna axis, as shown in figure 1-32, the amplitude of the video would vary as the antenna rotated about its axis. You should note that the video amplitude is maximum when the beam axis is at its highest elevation and minimum when the beam axis is at its lowest elevation. The video output from the receiver is filtered, leaving only the 50-Hz envelope to be employed as an error voltage.

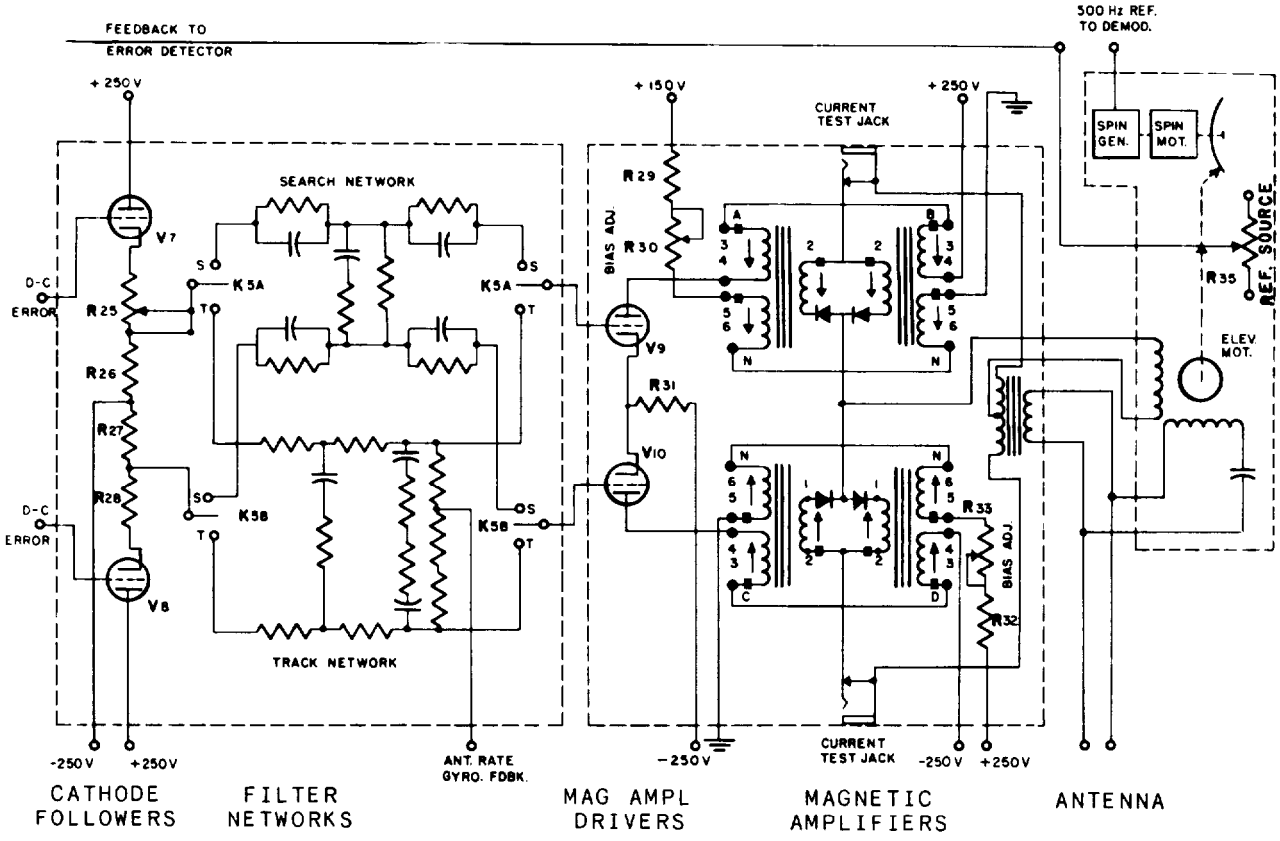
The function of the servo amplifier is to amplify the 50-Hz error voltage and compare its phase with the phase of the 50-Hz reference voltage originating in the 50-Hz spin generator. The phase of the output voltage to the servomotor causes the motor to rotate in the direction that reduces the amplitude of the error signal.

THEORY OF SEARCH OPERATION

The schematic diagram of the antenna servo system described above is shown in figure 1-33. As in the case of the block diagram, the system's search mode of operation is discussed first.

Scan Generator

The elevation scan generator is used during automatic search only. It consists of two resistors and one double-pole relay. Since only one resistor is in the circuit at a time, they serve alternately to unbalance the voltage applied to the error-detector potentiometer R3. The input to the scan generator is an ac voltage with its center point grounded by a resistor network. With both RI and R2 shorted, the center of R3 would also be at



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Figure 1-33.-(B) Servo system schematic diagram—Continued.

ground potential. Inserting R1 in the circuit would cause the center of R3 to be at some potential just as though the wiper of R3 has been moved to the right. Shorting R1 and inserting R2 should have the same effect as moving the wiper of R3 to the left.

The relay is actuated by a cam attached to the azimuth limit mechanism. The cam operates when the antenna reaches either azimuth limit.

Error Detector

The balanced potentiometer error detector consists of potentiometers R3 and R35. Potentiometer R35 is supplied with a 400-Hz reference voltage of approximately 32 volts amplitude while the reference voltage applied to R3 is modified in the manner that we have described. This voltage source is center tapped and grounded, thus reflecting an apparent ground at approximately the center of each potentiometer. A control handle displacement causing a change in the wiper contact of R3 results in an unbalanced voltage condition with an error signal being fed to the search contact of relay K2. With the equipment on search, the error signal is applied to the servo amplifier.

Servo Amplifier

The servo amplifier consists of the following stages:

1. Preamplifier
2. Phase shifter amplifier
3. Amplifier
4. Demodulator driver
5. Demodulator
6. Cathode followers
7. Search/track network
8. Magnetic amplifier drivers
9. Magnetic amplifier

Preamplifier V1 receives the error voltage from relay K2 and amplifies it. The preamplifier output is coupled through C1 to the grid of V2 and to the relay K3.

Phase shifter amplifier V2 is bypassed during search operation and is discussed under track operation.

Amplifier V3 provides an additional stage of amplification of the error signal. Its output is coupled through C4 to the gain control R15. The gain control determines the amplitude of the error signal fed to the demodulator driver. Thus, the gain of the antenna servo system is controlled by R15.

The demodulator driver provides the final amplification of the error signal prior to demodulation. As pointed out above, the amplitude of the signal applied to the grid can be controlled by the gain potentiometer (R15). The gain of the stage is stabilized by degenerative feedback. The feedback is accomplished by two means—an unbypassed cathode resistor R22 and a plate-to-grid feedback loop consisting of C5, C6, and R17. In addition to gain stabilization, the plate-to-grid loop provides the characteristic of an error-rate damper.

The full-wave demodulator employs two dual triodes, V5 and V6. Its operation is somewhat similar to that of a triode demodulator. The input error signal from the demodulator driver is applied to either the plate or cathode of the demodulator triodes. The reference voltage, which is 400 Hz during search operation, is supplied to the grids through either T1 or T2 with the primary-secondary phase relationship, as shown in figure 1-33, view A. (The small black rectangles on the input and output leads of the transformers in figure 1-33, view A, are polarity marks.) Instantaneous voltage polarity at the transformer primary polarity mark corresponds to the same polarity at the secondary polarity mark.)

Figure 1-34 shows a synchrogram of the voltages existing in the demodulator. When the error signal and reference voltage are in phase, V5A conducts on the first half-cycle and V6A conducts on the second half-cycle. Since V5B is cut off during the time V5A is conducting, V5A draws electrons from the top plate of C8, giving it a positive charge. During the second half-cycle,

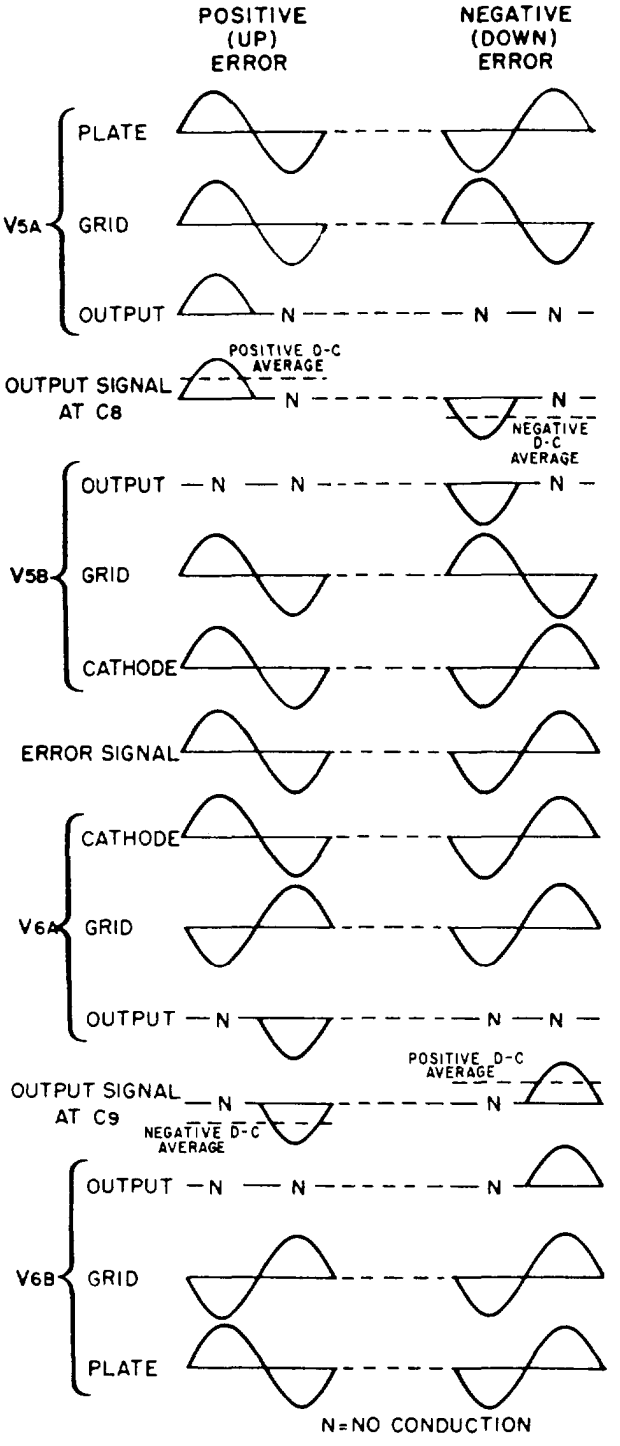


Figure 1-34.-Synchrogram of demodulator waveform.

electrons flowing through V6A are deposited on the lower plate of C9, giving it a negative charge. This action results in a push-pull output being supplied to the cathode followers.

When the error signal is 180 degrees out of phase with the reference voltage, V5B and V6B conduct on alternate half-cycles, charging C8 and C9 to the opposite polarity.

The cathode followers, V7 and V8, isolate the dc output of the demodulator from the low impedance of the search network (fig. 1-33, view B). Potentiometer R25 is provided to balance the outputs of V7 and V8 when no error signal is present.

The search/track network consists of two RC band-pass filter networks that have fairly long time constants. The search filter networks pass a 5-Hz signal while sharply attenuating lower frequencies. This action counteracts the high gain of the magnetic amplifiers, giving a flatter overall response for the servo system.

The magnetic amplifier drivers are dc amplifiers that control the current through the magnetic amplifier's control windings. The dc error signal is applied directly to the control grids of the drivers. The amplified dc error signal is applied to the control windings, controlling the output of the magnetic amplifier.

The magnetic amplifier provides the final stage of amplification of the error signal prior to the servomotor or output member. The amplifier consists of four amplifier sections: A, B, C, and D. Each amplifier section has three windings on its core—control, bias, and load. Referring to figure 1-33, view B, note that the control and bias windings of sections A and B are connected in series. The bias level is determined by the setting of potentiometer R30, and the control current is determined by the output of magnetic amplifier driver V9.

The C and D sections are connected in a similar manner with the bias level determined by potentiometer R33 and control winding current determined by the output of magnetic amplifier driver V10. The load winding of each section has a rectifier connected in series with it, allowing current to flow only in one direction. The polarity of the magnetic field resulting from current in each winding is indicated on the schematic by the direction of the arrows.

A synchrogram of waveforms illustrating the operation of the magnetic amplifier is shown in

figure 1-35. With a zero error signal applied to the grids of V9 and V10, conduction in all sections of the amplifier is equal. Waveforms showing the amount and time of conduction of each amplifier section under zero error signal conditions are shown in column A of figure 1-35. Waveforms (4) and (6) are equal in amplitude in column A but 180-degrees out of phase, resulting in zero output to the servomotor.

Column B of figure 1-35 illustrates the operation of the amplifiers with a positive error (an error signal that would cause the antenna elevation angle to be increased) applied to the magnetic amplifier drivers. The positive voltage applied to the grid of V9 increases the degree of core saturation, reducing the impedance of amplifier sections A and B. The negative voltage applied to the grid of V 10 decreases the degree of core saturation, increasing the impedance of amplifier sections C and D. Since the output of a magnetic amplifier varies inversely with its impedance, the output of sections A and B is increased in amplitude while the output of sections C and D is reduced in amplitude. Waveform (5) shows the algebraic sum of the two waveforms, which is fed to the servomotor.

Column C of figure 1-35 illustrates the operation of the amplifiers with a negative error signal applied. The output amplitudes have been reversed, causing the signal applied to the servomotor to be 180 degrees out of phase with that in column B.

Servomotor

The servomotor is a split phase ac induction motor whose field windings are excited by voltages that are 90 electrical degrees out of phase. The output of the servo amplifier determines whether the controlled winding is leading or lagging the uncontrolled winding. This phase relationship also determines the direction of rotation of the servomotor.

THEORY OF TRACK OPERATION

The purpose of the antenna servo system during track operation is to position the antenna

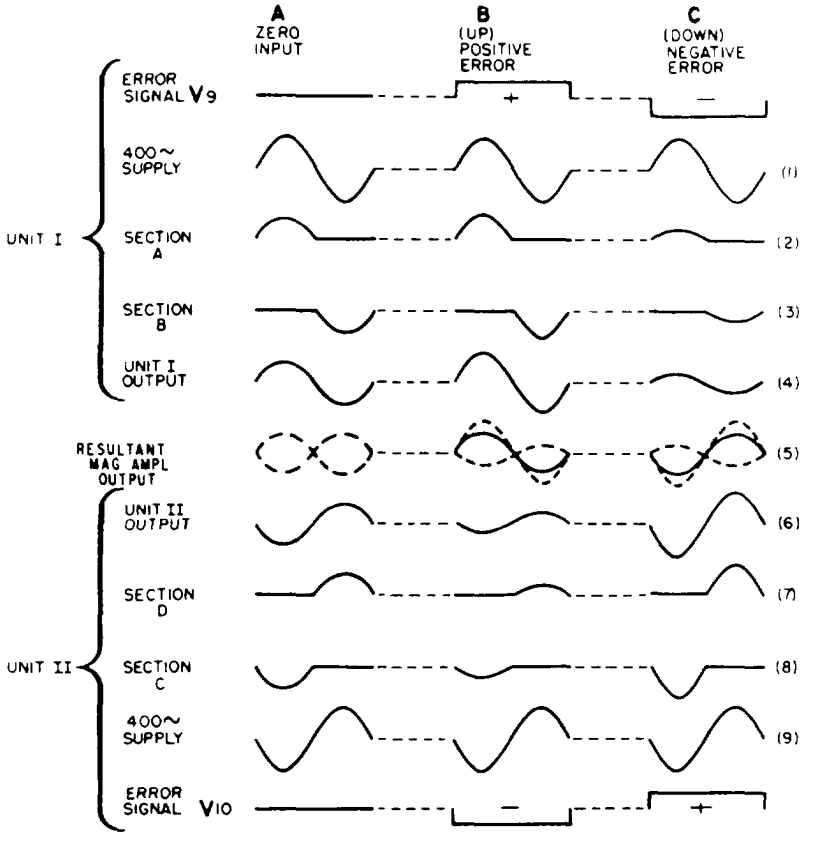


Figure 1-35.-Magnetic amplifier synchrogram.

based on the video output of the radar system. Again referring to figure 1-33, views A and B, note that relay K2 disconnects the error detector and control handle from the servo loop. The 50-HZ envelope of the video output constitutes the error signal and is fed to the servo amplifier.

Servo Amplifier

The 50-Hz error signal is amplified by the preamplifier and fed to the phase shifter. The phase shifter amplifies the error signal and also provides an adjustment, R13, to compensate for any phase shift of the error signal through the servo amplifier (fig, 1-33, view A). The output of the plate is coupled through C2 to R13, with the other end of R13 connected to the junction of R 1 and R12. The plate and cathode voltages are 180 degrees out of phase, and regenerative current flows in R12, resulting in regenerative feedback. The shifting of the error signal's phase is accomplished by varying the resistive-reactive ratio of the plate-to-cathode feedback loop. This varies the phase of the regenerative feedback, controlling the phase of the output error signal.

The operation of the amplifier and demodulator driver stages is identical under both modes of operation. However, the demodulator must now employ a 50-Hz reference voltage. Relay K4 is energized by the track/search switch, disconnecting the 400-Hz reference and connecting the 50-HZ reference supplied by the 50-Hz spin generator.

The outputs of the demodulator are fed through the cathode followers to the track section of the search/track networks. The track section is composed of two identical RC bandpass filters, which pass 2 Hz signals and attenuate all other signals. A signal from the elevation rate gyro is also used to control the error signal amplitude during track operation.

Since the magnetic amplifier drivers and the magnetic amplifier use dc error signals only, their operation is unchanged when switched to track operation.

Spin Generator

The 50-Hz spin generator is a permanent magnet ac generator that is driven by the spin motor. Its only function is to furnish a reference voltage for the demodulator during track operation.

MAINTENANCE AND ADJUSTMENTS

The maintenance of the antenna servo system is normally covered during phased inspections of the entire system. However, this section of the chapter will discuss maintenance pertinent to a typical antenna servo system.

Lubrication

The lubrication of the antenna system should follow the procedure set forth in the maintenance instruction manual (MIM) for the equipment. The lubricants and the time interval between lubrications should also be in accordance with standards established by the maintenance instructions. Instructions are normally issued by the squadron, supplying supplemental maintenance information and establishing schedules to be followed by maintenance personnel.

Alignment

The procedure for alignment of the antenna servo system is also found in the MIM for each piece of equipment. However, for illustration purposes, the alignment procedures applicable to a basic antenna servo system will be discussed here.

The first adjustment to be made is the balance control, R25 (fig. 1-33, view B). Its purpose is to ensure there is no output from the servo amplifier when no error signal is applied. Connect a dc voltmeter between the grid of V9 and the grid of V10. Place a jumper between the grid of V7 and the grid of V8. This shorts out any error signal and allows any imbalance of the cathode followers to be determined. Adjust R25 until there is a zero voltage reading on the voltmeter. Remove the voltmeter from the circuit, but do not disconnect the jumper, as it is required for the next adjustment.

The bias adjustments in the magnetic amplifiers are made by connecting a milliammeter in the load winding of each amplifier and adjusting the bias controls, R30 and R33, to the current specified by the MIM. A current jack is normally incorporated in the equipment so you can use a standard milliammeter. Remove the jumper from the grids of the cathode followers and disconnect the meter.

The gain adjustment is made by inserting a voltage of a specific amplitude and frequency at the input of the preamplifier and measuring the output of the demodulator driver V4, which is the

last ac amplifier stage. The MIM will normally specify the amplitude and frequency of the input signal and the output stage.

Make the phase adjustment when the equipment is in track operation and locked on a strong target. Disable the antenna azimuth channel by removing the demodulator tubes or at some other location specified by the MIM. With the equipment operating as specified above, manually rotate the antenna in azimuth. Any change in the elevation of the antenna indicates an undesirable phase shift in the amplifier. Vary the phase adjustment until any movement of the antenna in azimuth causes no change in its elevation.

Adjustment potentiometers normally have locknuts to prevent vibration from affecting their setting. Loosen the locknut prior to adjusting and be careful when tightening so you do not disturb its setting. The locknut is normally sealed with Glyptal to prevent the locknut from being loosened by vibration.

HYDRAULIC SERVO SYSTEM

Another type of antenna servo system that is in use is a hydraulically driven antenna. This system has the advantages of low response time to a command signal, low weight-to-power ratio, and a high degree of accuracy. Figure 1-36 shows a simplified block diagram of such a system. Only the azimuth channel is shown because the elevation channel functions similarly.

The antenna drive system converts the electrical energy from the magnetic amplifier into hydraulic pressure, which drives the antenna. Antenna position information is mechanically coupled to the induction follow-up potentiometer, which sends a feedback voltage to the magnetic amplifier, nulling out the command signal.

Figure 1-37 is a diagram of a servo valve and actuator used to actuate an antenna. Two such devices are required, one for azimuth and one for elevation.

Drive signals from the magnetic amplifier, acting on the driving coils, produce forces on the flapper at the permanent magnet air gap. Hydraulic fluid, under pressure from the aircraft's hydraulic system, enters the servo valve through fixed orifices (1). After passing through the fixed orifices, the fluid continues through the variable area exit nozzles and then to the low-pressure return (2).

Normally, the flapper is positioned so that an equal amount of hydraulic fluid flows through each nozzle. This balances the pressure forces on the spool valve. The spool valve is positioned so that with zero signal conditions, the pistons on the spool valve close off the hydraulic supply and return lines to the antenna actuator. The spool centering screw (3) is used to adjust the position of the spool valve by balancing the forces developed by the spool valve centering springs.

The flapper moves when a signal passes through the driving coils. The direction and amount of movement depend upon the polarity and magnitude of the drive signal. Movement of the flapper valve to the left increases the pressure in the left control port because the flapper restricts the amount of flow through the left nozzle. Also, flapper movement to the left allows more fluid to flow through the right nozzle, thus reducing pressure in the right control port. The resulting pressure imbalance results in movement of the spool to the right until the counterforce developed

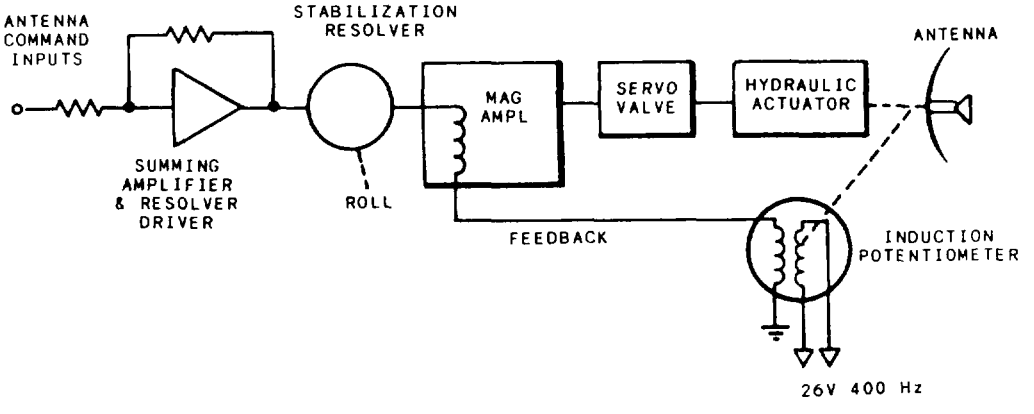
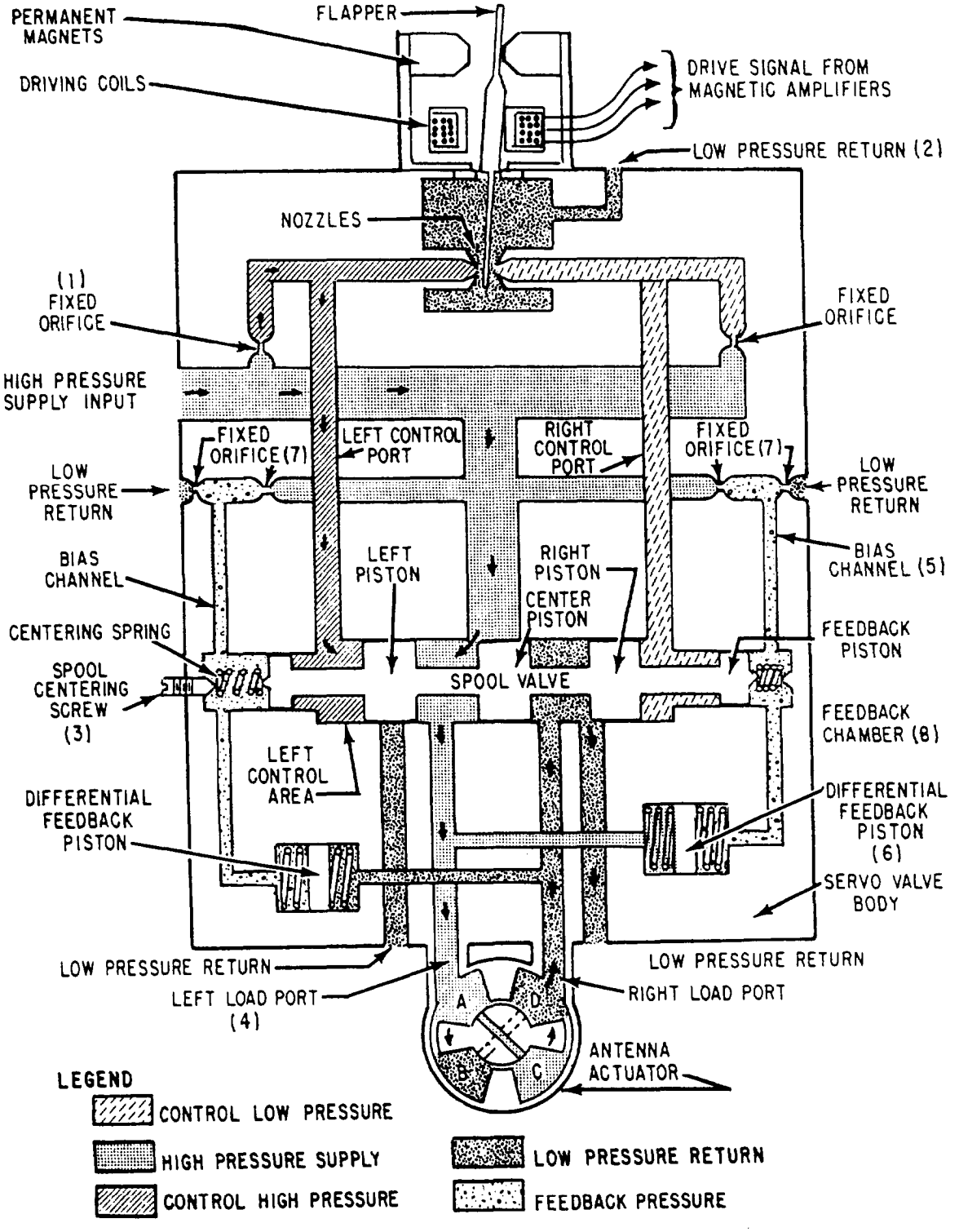


Figure 1-36.-Antenna positioning circuit.



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Figure 1-37.-Servo valve and actuator.

by the compression of the spool valve centering spring equals the force developed by the pressure difference across the spool valve. As the spool valve moves to the right, the center piston allows an amount of hydraulic fluid proportional to the drive signal from the magnetic amplifier to flow from the high-pressure supply through the left load port (4) into the actuator.

Hydraulic fluid applied to the left load port produces an increase in pressure in chamber A. The pressure increase is transmitted through a hole in the rotor to chamber C. At the same time, chambers B and D are connected to the low-pressure return line by the right piston on the spool valve. The pressure difference between the chambers produces a counterclockwise torque on the moving vanes of the rotor. The rotor is connected directly to the antenna gimbal, so that motion of the rotor moves the antenna.

You should be aware of the tendency of servo systems to oscillate. As already discussed, various types of mechanical and electronic devices (dampers) are used to minimize these tendencies. In hydraulic servo systems, oscillation can be damped by the use of hydraulic pressure feedback.

The feedback system is an integral part of the antenna servo valve and consists of a bias channel (5), a feedback channel from each of the antenna actuator drive channels, a differential feedback piston (6), and a feedback piston that is part of the spool valve. Each servo valve has two derivative feedback paths, which together form a push-pull system. Hydraulic fluid from the high-pressure supply flows to the return line through two fixed orifices (7), which are designed so that the bias pressure in the chamber between the orifices is approximately equal to one-half the supply pressure. The bias pressure is applied to one side of the differential feedback piston (6). The piston moves until the restoring force developed by the appropriate centering spring in the piston chamber equals the force developed by the bias pressure.

The other side of the differential feedback piston is connected through the feedback channel to the load port of the opposite channel. The cross-coupling makes the feedback 180 degrees out of phase with the driving signal, providing negative feedback. The steady-state pressures in the two feedback chambers (8) are equal. Because of the absence of differential feedback piston motion, the feedback chamber pressures are determined by the supply and return pressures at the fixed orifices. Any change in the input drive signal changes the pressure in the appropriate load port. When coupled through the feedback channels, the pressure change causes the differential feedback pistons to move so that the pressures in the feedback channels and the bias channels tend to be equalized. The change in bias pressure (caused by the movement of the differential feedback piston) bleeds off into the low-pressure return line at a rate determined by the size of the fixed orifices (7). This returns the forces on the feedback pistons to their balanced condition.

While the differential feedback piston is in motion, the pressure it develops within its chamber opposes the motion of the spool valve caused by the input drive signal. The force developed in the feedback chamber is proportional to the rate of change of the drive signal to the torque motor. The greater the rate of change of the drive signal, the greater the instantaneous feedback pressure.

Hydraulic servo valves and actuators are precision devices. The parts are delicately balanced and can easily be damaged by rough treatment and contamination. The torque motor, for instance, can be ruined by tapping it with a screwdriver handle. The components must be handled with care. Avoid dropping or jarring, and prevent wrenches and other tools from striking them. When replacing hydraulic components, extreme care must be exercised to prevent contamination of the system. The necessity for cleanliness cannot be overemphasized. The spool valve and the small filters within can be jammed or damaged by dirt. The valve can become prematurely loaded by contamination introduced through careless handling. Even the tiniest particle from a person's finger can completely block the servo valve and precision orifices. These orifices are as small as a human hair, and they do not have to be completely blocked to destroy their operation.

CHAPTER 2

LOGIC DEVICES

Chapter Objective: *Upon completion of this chapter, you should have a working knowledge of logic devices to include semiconductor and integrated circuits.*

The purpose of this chapter is to help you understand how integrated circuits (ICs) work and how they are used. Both analog and digital circuit are used. First, semiconductor theory is covered to give you an understanding of how semiconductors work. Then fabrication of transistors and integrated circuits are discussed to give you an insight into the manufacturing methods. The function of some ICs are discussed in detail to give you a working knowledge of ICs and their uses.

SEMICONDUCTOR

Learning Objective: *Recall the basic concepts of semiconductor transistor theory and the operation of transistors.*

Semiconductor refers to each of the unique components produced by solid-state technology. These components are called semiconductors because a device so designated is capable of functioning as a conductor and as a nonconductor. Its operation depends on the amount of applied voltage and its polarity and/or its specific connection in an electronics circuit. In this section, we will discuss the diode, transistor, and thyristor semiconductors.

DIODE

Semiconductor diodes are employed for rectification and detection. In addition, they have special properties that make them particularly useful for bias and voltage stabilization.

Since junction diodes can be made of the same material as the transistor and have the same temperature coefficient and resistance, they will track better over the same temperature range, providing nearly ideal thermal compensation. Likewise, application of the avalanche breakdown phenomena provides a special voltage-stabilizing (Zener) diode.

The junction diode has four important ratings that must be taken into consideration when used in a power supply. They are the maximum

1. average forward current,
2. repetitive reverse voltage,
3. surge current, and
4. repetitive forward current.

These ratings are important to you, as a technician, when it becomes necessary to troubleshoot a power supply or when selecting junction diodes for replacement when the desired one is not readily available.

The maximum average forward circuit is the maximum amount of average current that can be permitted to flow in the forward direction. This rating is usually given for a specified ambient temperature and should not be exceeded for any length of time, as damage to the diode will occur.

The maximum repetitive reverse voltage is that value of reverse bias voltage that can be applied to the diode without causing it to break down.

The maximum surge current is that amount of current allowed to flow in the forward direction in nonrepetitive pulses.

The repetitive forward current is that value of forward bias voltage that can be applied to the diode without causing it to breakdown.

All of the ratings mentioned above are subject to change with temperature variations. If the temperature increases, the ratings given on the specification sheet should all be lowered to prevent damage to the diode.

TRANSISTORS

A transistor is an active component of an electronic circuit that may be used as an amplifier, detector, or switch. A semiconductor transistor consists of a small block of semiconducting material to which at least three

electrical contacts are made. Transistors are of two general types-bipolar and field-effect. The bipolar type involves excess minority current carrier injection. The field-effect type involves only majority current carriers. Historically, the bipolar type was developed before the field-effect type; both are widely used today. The unmodified term *transistor* usually refers to the bipolar type.

Bipolar Transistor

In a bipolar transistor, at least one contact is ohmic (nonrectifying), and at least one contact is rectifying. Usually there are two closely spaced rectifying contacts and one ohmic contact.

The operation of a simple transistor consists of the control of the current flowing in the high-resistance direction through one rectifying contact (called the collector) by the current flowing in the low-resistance direction in the other rectifying contact (called the emitter). The third contact, which is ohmic, is called the base contact.

These contacts usually consist of two or more regions. The regions in which the actual rectification processes take place are called the emitter barrier and collector barrier. The region between these two barriers is called the base region, or simply the base. The regions outside of these barriers are called the emitter and collector regions.

Transistors are used in radio receivers, in electronic computers, in electronic instrumentation and control equipment, and in almost any electronic circuit where vacuum tubes are useful and the required voltages are not too high. Transistors have the advantages over their vacuum-tube counterparts of being much smaller, consuming less power, and having no filament to burn out. A disadvantage in using transistors is that they do not yet operate at high voltages as some vacuum tubes do, and their operation is degraded at high temperatures.

CLASSIFICATION OF TRANSISTORS.— Transistors are classified chiefly by four criteria: (1) by the type and number of structural regions of the semiconductor crystal; (2) by the technology used in fabrication; (3) by the semiconductor material used; and (4) by the intended use of the device. A typical designation following this scheme would be *npn* double-diffused silicon switching transistor. It is not necessary to include all of the above criteria in a single designation nor to rigidly follow this order.

A modem transistor type is the *npn* double-diffused silicon planar passivated transistor (fig. 2-1). The term *double-diffused* refers to the fabrication technique in which the base region is formed by diffusion through a mask into the body of the silicon wafer, which forms the collector region.

In turn, the emitter region is formed by diffusion through a second mask into the previously formed base region. The term *planar* refers to the fact that all three electrical connections are found on a single surface of the device. The term *passivated* means that the surface to which all junctions return is protected by a layer of naturally grown silicon oxide, which, together with an overcoating of glass or other inert material, passivates the surface, electrically minimizing leakage currents. The double-diffusion process allows very close control of narrow base widths. The base diffusion provides a resistivity gradient in the base region, which has an associated electric field. In this field charge transport is by drift. Such transistors have been called drift transistors to distinguish them from most other transistors in which the charge transport is by a diffusion process. Silicon planar transistors have power ratings in the 100 mW to 50 W range with characteristic frequencies between 50 and 2000 MHz, usually of the *npn* type. The designation *npn* stands for the conductivity type of the emitter, base, and collector regions, respectively. The *n* stands for negative, since the charge on an electron is negative, and electrons carry most of the current in a region of *n*- type conductivity. In a region of *p*- type conductivity, most of the current is carried by electron vacancies called holes, which behave as if they were positively charged.

A historically important type was the *pnp* alloy-junction germanium transistor. This type was widely used in the first decade of the solid-state electronics era. The term *alloy-junction* in this transistor designation refers to the fabrication method. The emitter and collector regions were produced by recrystallization from an alloy of some suitable metal doped with a *p*-type impurity. The alloy had previously been fused in contact with the opposite surfaces of the original *n*-type semiconductor body, and it had dissolved some of the semiconductor material. Fused-junction is equivalent terminology. This type of transistor was made in power ratings from 50 mW to 200 W, and in frequency ranges up to 20 MHz.

TRANSISTOR ACTION.— To explain transistor action in more detail, some of the basic properties of a semiconductor material are presented first. An *n*-type semiconductor contains electrons, and a *p*-type

semiconductor contains holes. These are called the majority carriers of the two types. Actually, there are always present a small number of holes in an *n*-type semiconductor and a small number of electrons in a *p*-type semiconductor. These are called the minority carriers of the two types. At a given temperature with a given material, the product of the densities of the majority and minority carriers is a constant. This means that if there is a very high density of majority carriers (low-resistivity material) present, there will be a correspondingly low density of minority carriers.

The emitter current controls the collector current in a simple transistor. To understand this, first consider the magnitude of the collector current in the absence of emitter current. In normal operation the collector barrier is biased in the high-resistance (reverse) direction. Under this condition of bias, the majority carriers are stopped by the barrier, and only the minority carriers are free to flow. If the collector barrier is a silicon *pn* junction, the minority-carrier diffusion current is negligible, and the reverse-bias leakage current will consist of thermally generated carriers and be in the nanoampere range. If emitter current is present, the portion consisting of carriers entering the base will continue across the collector barrier, and thus control the collector current.

INJECTION.— The emitter controls the density of minority carriers by injecting extra minority carriers into the base region when the emitter is biased in the low-resistance (forward) direction. This is the fundamental process of simple transistor action. Whenever a rectifying barrier is forward-biased, extra minority carriers are added to the semiconductor near the barrier. Since the source of these minority carriers is the majority-carrier density on the other side of the barrier, it is clear that the largest part of the forward

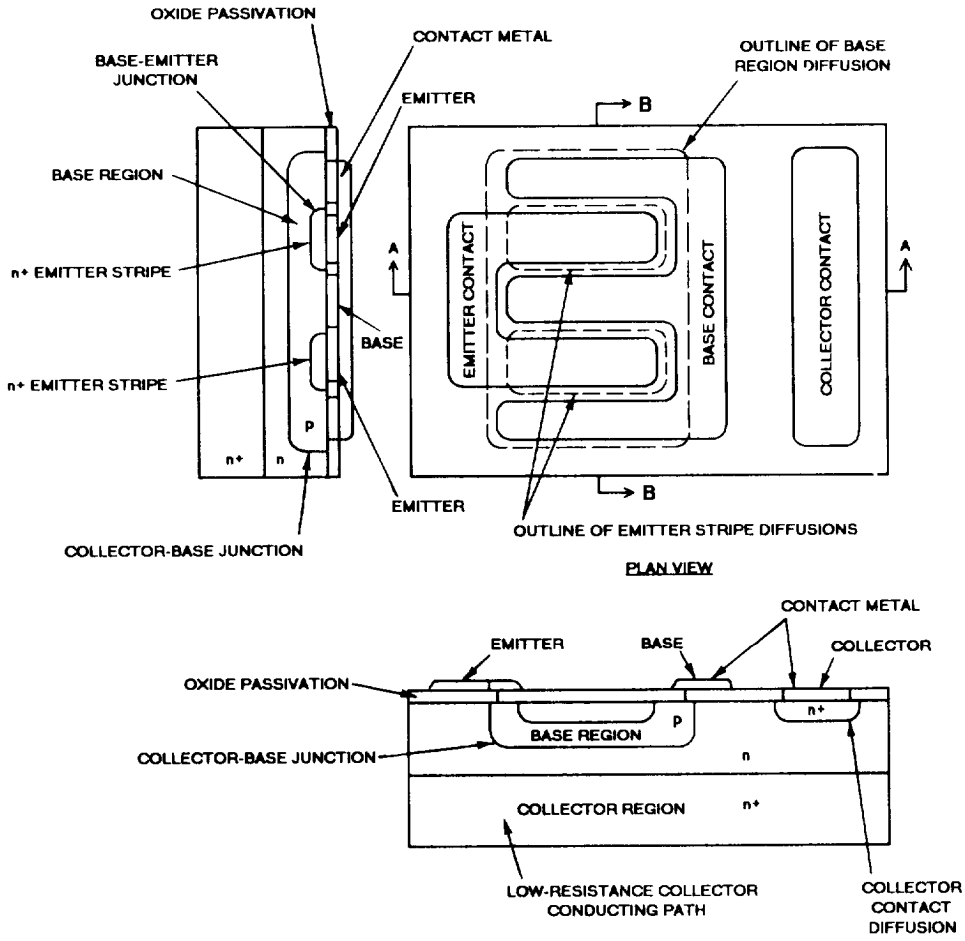


Figure 2-1.-Sections of a planer npn double-diffused silicon transistor.

current will be carried by those carriers that come from the largest majority density. A *pn* junction will have a high injection efficiency for electrons if then region has a much larger density of carriers (lower resistivity) than the *p* region. Therefore, in the *npn* transistor, the emitter *n* region should have a low resistivity compared to the p-type base region. Also, the phenomenon of minority-carrier injection is observed in rectifying metal semiconductor contacts, and such contacts may be used as emitters as well as *pn* junctions.

CURRENT GAIN.— The current gain of a simple transistor may be expressed as the product of three factors: (1) the fraction γ of the emitter current earned by the injected carriers, and (2) fraction β of the injected carriers that arrive at the collector barrier, and (3) the current multiplication factor α^* of the collector. For a double-diffused transistor, typical values of these factors are $\gamma = 0.985, \beta = 0.999, \text{and } \alpha^* = 1.000$, giving $\alpha = 0.984$. From this you can see that most of the current that flows into the emitter flows right on through the base region and out the collector, while only a small fraction (here 0.016) flows out the base connection.

For a fixed value of emitter current I_e there is a fixed value of collector current αI_e added to the collector-barrier leakage current I_{co}, giving a total collector current, $I_c = I_{co} + \alpha I_e$. This means that the slope of the dc characteristics should be the same as the slope of the collector-barrier leakage current curve for $I_e = 0$. The typical characteristics shown in figure 2-2 illustrate this. The slope of the collector leakage curve is very low since the collector voltage does not influence the relatively fixed number of minority carriers carrying the current.

HIGH-FREQUENCY EFFECTS.— These originate in three distinct properties of transistors: the transit time of injected carriers across the base region, the charging time of the collector- or emitter-barrier capacitance through the base region and collector-region resistance in series, and the time required to build up the proper density of injected carriers in the base region (called storage-capacity effect). In alloy-junction transistors with a base region of uniform resistivity, the transport of injected carriers across the base is usually the limiting factor. Of course, base transit time alone introduces only a phase shift between the emitter and collector signals, but this time also gives a chance for injected carriers, bunched by the emitter signal, to diffuse apart, and therefore degrade the signal (fig. 2-3).

In double-diffused (drift) transistors, the base transit time is usually negligible compared to the charging time of the collector or emitter capacitance. In some units, the storage capacity (often called diffusion capacity) seems to be an appreciable limitation.

Storage capacity also shows up in another way in transistors used as switches. Here it introduces a time delay both in turning on and in turning off the transistor. The turn-off delay is usually longer than the turn-on delay, because the density of injected carriers in the base region has had time to build up to large values during the time the transistor was on, and therefore takes a long time to subside to the level where the transistor can turn off. These delays are only slightly related to the actual time of rise or fall of the collector level, which is determined primarily by the collector-capacitance-base-resistance time constant.

A fabrication technique, called "epitaxial growth," is used to minimize the storage capacity effects in high-speed transistors. In this process a transistor structure is formed entirely in a very thin skin of good

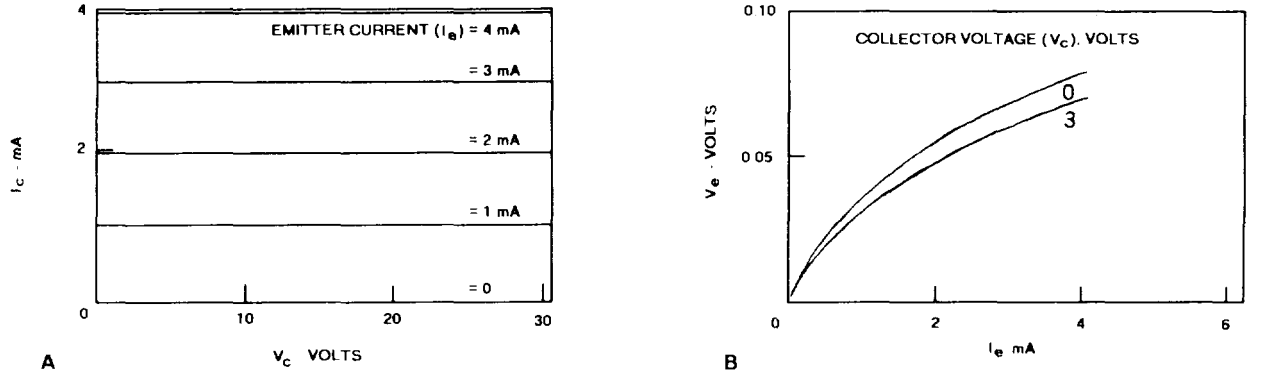


Figure 2-2.-Transistor dc characteristics-(a) collector, and (b) emitter.

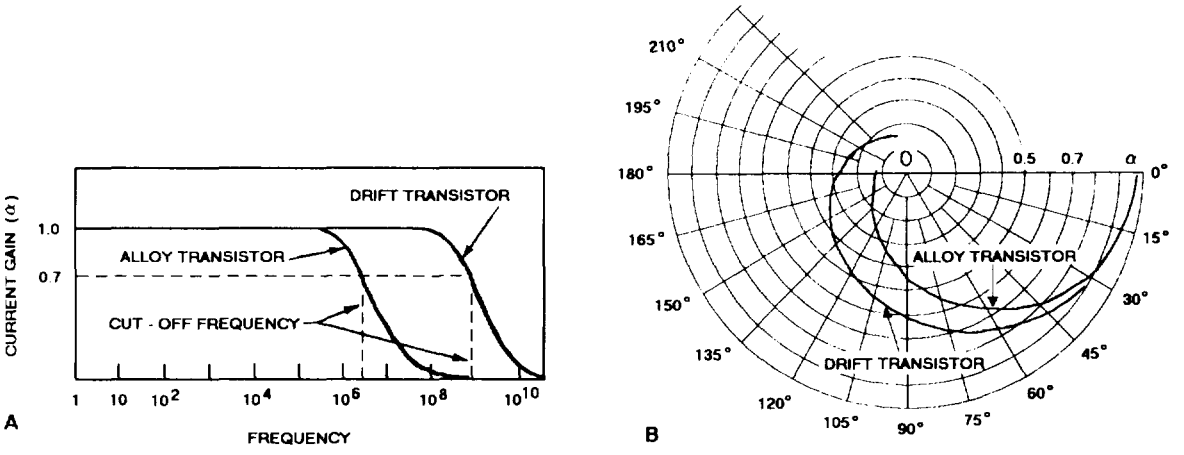


Figure 2-3.-Transistor frequency characteristic-(a) frequency dependence of α, and (b) phase of collector current I_c versus emitter current I_e.

semiconductor material grown upon the surface of a wafer of heavily doped material. The heavily doped material has very low lifetime for excess carriers and, therefore, a very low storage effect, as well as a low series resistance. The collector junction of such a transistor is close to this low-lifetime material, but is formed in the high-quality, epitaxially grown skin so that its properties are not degraded by the heavily doped material. Such transistors are called "epitaxial transistors."

Close control of the injection ratio γ, defined above, is afforded by the fabrication technique of ion implantation. In this technique a beam of ions composed of the desired dopant material is accelerated to a specific kinetic energy and caused to strike the surface of the region to be doped. The ions penetrate the surface and remain embedded in the semiconductor material. By controlling the ion-beam current and the time of bombardment, a very accurate control of the total number of dopant ions in the region is achieved. After heating the semiconductor to diffusion temperature, the ions move on into the material, creating the emitter and base regions of the double-diffused structure. These regions now have precisely controlled doping, and hence show a γ-factor within ±1 percent of the design value.

TRANSISTOR NOISE.— Noise is quite low if a low source impedance is used. With source impedances of about 1000 ohms, a good junction transistor will have a noise factor of about 4 dB. The noise factor is independent of the connection, but rises with source impedances above 10,000 ohms and with frequencies below 1000 Hz.

TEMPERATURE EFFECTS.— These are most marked in connection with the collector-barrier leakage current with no emitter current flowing I_{co}. This current increases exponentially with temperature and leads to a phenomenon called "thermal runaway." If a transistor is operated at a given ambient temperature and a given initial power dissipation, this power will soon raise the temperature of the collector barrier, which then draws more current, and, in turn, increases the dissipation. The process is cumulative, and precautions must be taken to stabilize against it. Current gain increases slightly with increased temperature in most ***npn*** transistors, but this is a small effect unless the current gain is unusually close to unity.

POWER SWITCHING.— There are several transistor structures that are used for power switching and that make use of current gains greater than unity to achieve a thyraton-like characteristic. These devices are often called "four-layer" devices because they usually contain four regions of alternating *n*- and *p*-type semiconductor material. Connections are made to the end regions and to one of the interior regions. The end regions are oppositely biased so that the center junction is reverse-biased. The connection to the interior region is then the control, and is usually called the "gate." When the gate is biased to cause injection of excess carriers across the junction between it and the nearest end connection, the device is triggered on, and a saturation current is drawn between the two end connections, normally called anode and cathode. Such devices are normally classified as rectifiers, but, in reality, they are a form of transistor.

Field-Effect Transistor (FET)

There are two major types of field-effect transistors–the junction-gate FET (JFET) and the insulated-gate FET (IGFET). The IGFET is commonly called MOSFET or MOS transistor. The acronym MOS stands for metal-oxide-semiconductor, which describes, in order, the structure of the device from the gate toward the channel. The JFET was developed first, since it involved no technology beyond that of the planar bipolar silicon transistor. The development of the MOSFET was delayed while the technology was extended to stable control of silicon surface potential. The MOSFET is widely applied in large-scale integration, particularly in implementing large random-access, high-speed memories for computers.

JFET.— Figure 2-4, view A, shows a section of a JFET. The channel consists of relatively low-conductivity semiconductor material sandwiched between two regions of high-conductivity material of opposite type. When these junctions are reverse-biased, the junction depletion regions encroach upon the channel, and finally, at a high reverse bias, pinch it off entirely. The thickness of the channel, and hence its conductivity, is controlled by the voltage on the two gates. Therefore, this device is normally on and maybe switched off. It is called a "depletion-mode" FET. In practice, this FET has an input impedance several orders of magnitude greater than that of a silicon bipolar transistor. JFET's are made in both *n*-channel and p-channel types. They are used in amplifiers, oscillators, mixers, and switches. The general performance limits are about 500 MHz, 1 W, 100 V, and 100 mA (saturation drain current). They also find application in integrated circuits employing bipolar transistors because their technology is compatible.

MOSFET.— Figure 2-4, view B, shows a section of a MOSFET. Here the source and drain regions consist of *n* diffusion in a *p*-type substrate. The gate is a metal film evaporated on a thin silicon dioxide (SiO_2) insulator spanning the separation between the source and drain. With no voltage on the gate, the source and drain are insulated from each other by their surrounding junctions. When a positive voltage is applied to the gate, electrons are induced to move to the surface of the p-type substrate immediately beneath the gate, producing a thin surface of induced n-type material, which now forms a channel connecting the source and drain. Such a surface layer is called an inversion layer because it is of opposite conductivity type to the substrate. The number of induced electrons is directly proportional to the gate voltage, so that the conductivity of the channel increases with gate voltage. This device is called an "*n*-channel enhancement-mode" MOSFET. It is normally off at zero gate voltage.

Because of the quality of the silicon dioxide gate insulator, the input impedance of a MOSFET is several orders of magnitude greater than that of a JFET. Typical MOSFET dc characteristics are shown in figure 2-5. The low-drain voltage channel resistance is inversely proportional to $(V_{gs} - V_{th})$, where V_{gs} is the gate source voltage and V_{th} is the threshold voltage, and the saturation drain current is proportional to $(V_{gs} - V_{th})^2$.

MOSFET devices are fabricated in both p-channel and n-channel types, as well as for both depletion (normally on) and enhancement (normally off) modes of operation. In a MOSFET the mode of operation is determined by a threshold voltage of the gate at which the device changes from off to on, or vice versa. In modem technology this threshold voltage can be set for a wide range of values by the use of ion implantation through the gate oxide.

MOSFET discrete devices are used for ultrahigh-input impedance amplifiers such as electrometers where the input leakage current is less than 10^{-14} A. Dual-gate depletion types can be used as mixers up to 1000 MHz, and power-switching types (the

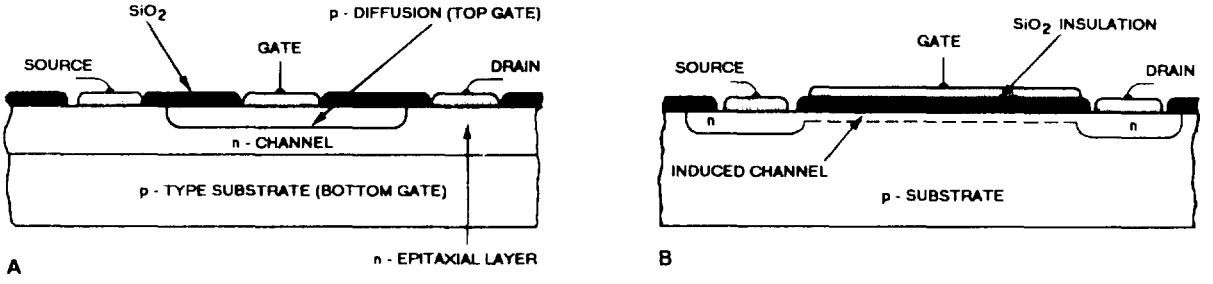


Figure 2-4.-Field-effect transistor—(A) junction-gate FET (JFET); (B) insulated-gate FET (IGFET or MOSFET).

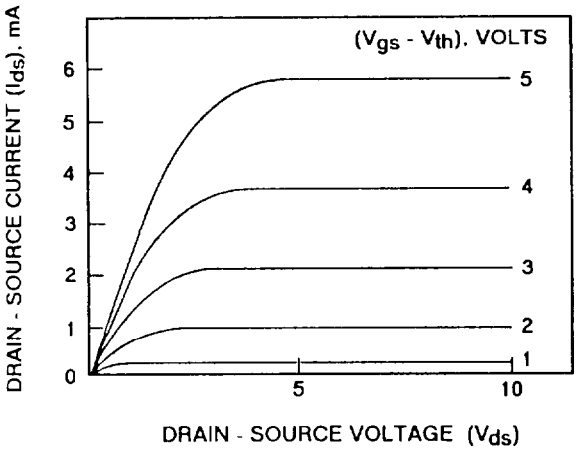


Figure 2-5.-MOSFET drain characteristics.

VMOS discussed below) are good to 25 W, 2 A, or 100 V. Most integrated circuits using MOSFETs are called CMOS integrated circuits, where the C stands for complementary. These circuits use *n*-channel and p-channel types together to achieve digital logic. Typical propagation delay times through small-scale integrated building-block circuits such as three-input NAND or NOR gates is about 20 nanoseconds (ns) for a 20-picofarad load. At a 10-MHZ clock rate, the power dissipation for such a gate is about 10 mW. For large-scale integration, a typical 16 kilobit random-access memory has an access time of 200 ns, an active power of 500 mW, and a standby power of 20 mW.

There are a number of variations of the MOS technology. Two of particular interest are VMOS (V for vertical) and SOS (silicon on sapphire). The VMOS device is fabricated by etching a notch down through a planar double-diffused structure similar to that of an *npn* bipolar transistor. The surface of the notch is first oxidized and then covered with the gate metallization. The source contact bridges the *n+-p* junction near the surface, and the drain connection corresponds to the collector contact of the bipolar structure. The channel length is now determined by the thickness of the *p* region. This allows controlled short channels, and gives both high-current and high-voltage capability.

The SOS device is fabricated in a very small silicon body grown epitaxially on a sapphire substrate. An experimental MOS/SOS 1000-bit memory has shown a standby power of only 1 microwatt.

Transistor Manufacture

The manufacture of transistors has required a whole new field of exacting technology. Good semiconductor material requires the maintenance of chemical purities far beyond the spectroscopic range. A purity of 1 part in 10^8 is not unusual. Most devices must be made from oriented single crystals of semiconductor material, which can have only very low densities of structural defects.

Physical tolerances of the high-frequency transistor structures are microscopic; the separation of emitter and collector junctions must be of the order of a few micrometers in these units.

To solve these problems new techniques have appeared. Purity is achieved by melting a small zone of a bar, or ingot, and gradually passing this molten zone from one end of the bar to the other. Impurities in the material remain in the liquid phase and are earned along with the molten zone, leaving high-purity material behind.

Tolerances are achieved by a collection of new techniques, such as epitaxial growth, solid-state diffusion, ion implantation, and the photolithographic delineation of diffusion masks.

Transistor Connections

It is important for you to understand the method of connecting a transistor into a circuit. Bipolar transistor connections and field-effect transistor (FET) connections are discussed in the following text.

BIPOLAR TRANSISTOR CONNECTIONS.— The common-emitter, common-base, and common-collector connections are the most frequently used connections of bipolar transistors, and of these the common-emitter connection is by far the most popular for transistors with current gain $\alpha < 1.0$. To compare these connections, you should examine the small-signal current gain, voltage gain, input resistance with shorted output, and output resistance with open-circuited input. Shorted output means an ac short; the dc bias voltage is still present. Open-circuited input means a constant dc current bias. These quantities are easily measured, and they are useful in calculating the performance of a transistor in a circuit connection.

Common-Emitter Connection.— This connection (fig. 2-6) has a base-to-emitter input and a collector-to-emitter output connection. The slope of the illustrated I_c versus V_c characteristics gives a commonly used conductance h_{cc}, which is the reciprocal of the output resistance with open-circuited (I_b = constant) input.

The current gain h_{cb} (equal to $\partial I_c/\partial I_b$, with V_c = constant) is related to the current gain α by $h_{cb} = \alpha/(1-\alpha)$. The input resistance h_{bb}, with a short-circuited (Vc = constant) output, is given by the slope of the V_b versus I_b characteristics. Finally, the voltage gain $1/h_{bc}$ (equal to $\partial V_c/\partial V_b$, with I_b = constant) can be obtained from the separation of these curves. Typical values of these small signal parameters arc h_{bb} = 1800 ohms; $1/h_{bc}$ = 1600; h_{cb} = 50; and $1/h_{cc}$ = 0.05 x 10^6 ohms.

Common-Base Connection.— This has an emitter-to-base input and a collector-to-base output connection. As with the common-emitter connection, the output resistance $1/h_{cc}$ is found from the slope h_{cc} of the I_c versus V_c characteristics. The current gain h_{ce} equals $\partial I_c / \partial I_e (h_{ce}) = -\alpha$ because of the assigned polarity of the currents (fig. 2-7). The input resistance h_{ee} is found from the slope of the V_e versus I_e characteristics. The voltage gain $1/h_{ec}$ equals $\partial V_e/\partial Vc$. Typical values of the common-base parameters arc h_{ee} = 36 ohms; $1/h_{ec}$ = 1500; h_{ce} = –0.98; and $1/h_{cc}$ = 2.5 x 10^6 ohms.

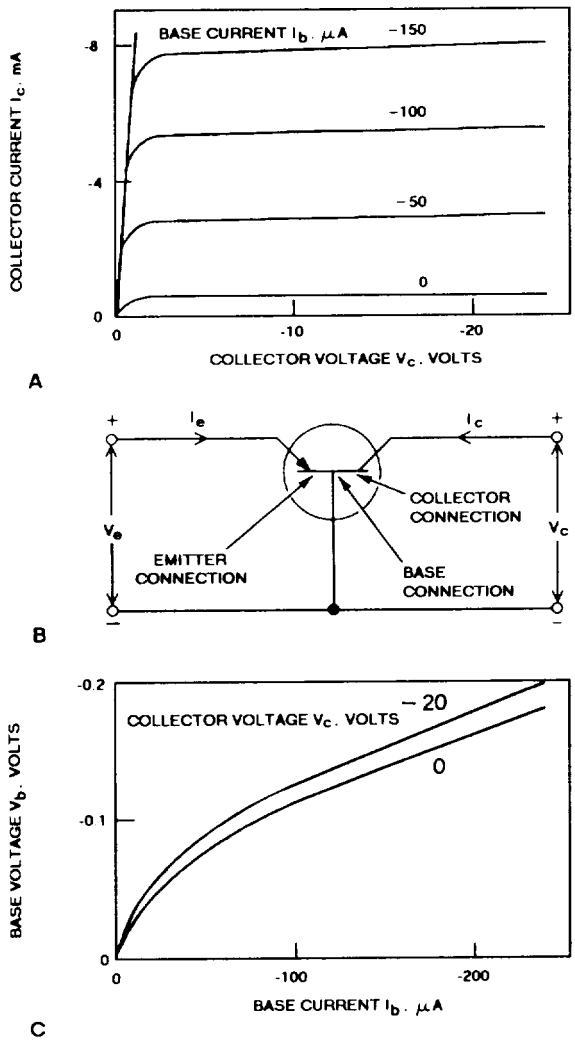


Figure 2-6.-Common-emitter connection of an alloy-junction transistor; (a) collector characteristics; (b) schematic diagram; (c) base characteristics.

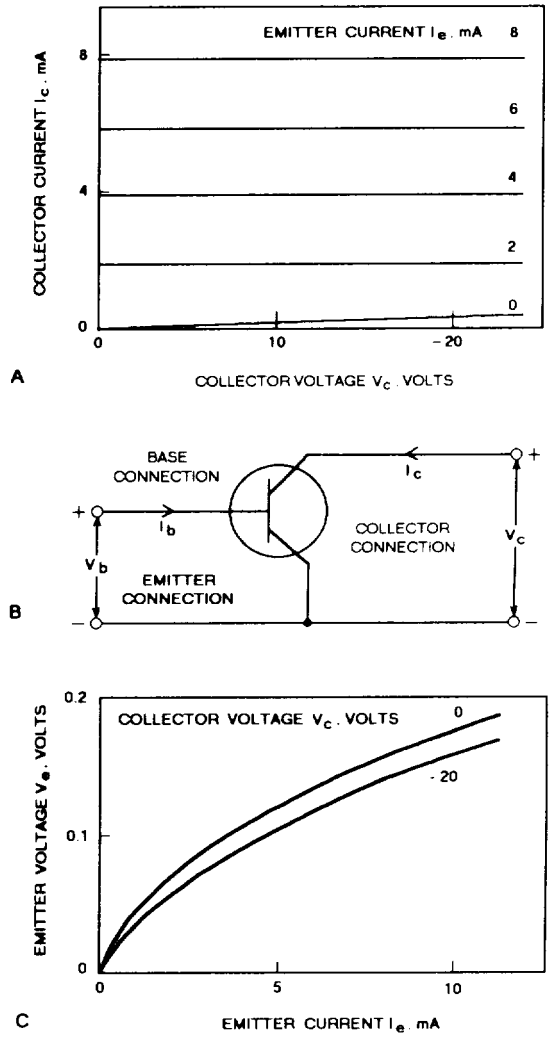


Figure 2-7.-Common-base connection of an alloy-function transistor; (a) collector characteristics; (b) schematic diagram; (c) emitter characteristics.

Common-Collector Connection.— This connection, as shown in figure 2-8, has a base-to-collector input and an emitter-to-collector output connection. Again, the output resistance $1/h_{ee}$ is found from the slope h_{ee} of the I_e versus V_e characteristics. The current gain h_{eb} equals $\alpha Ie/\alpha I_b$. The *input resistance* h_{bb} is found from the slope of the V_b versus I_b characteristics. The voltage gain $1/h_{bc}$ equals $\alpha V_c/\alpha V_b$. *Typical values of the common-collector parameters are* h_{bb} = 1800 ohms; $1/h_{be}$ = 1; h_{eb} = –50; and $1/h_{ee}$ = 0.05 x 10^6 ohms.

Selection of Bipolar Transistor Connections.— From the foregoing definitions, it is possible to compare these connections for use in amplifiers. The power gain of the common-emitter connection (current gain times voltage gain) is the highest. This is the reason that this connection is used most frequently. The common-base connection shows the lowest input resistance and the highest output resistance. Its most useful characteristic is its linearity, which gives low distortion when driven by a current source. Accordingly, it is often used with a driver transformer. The common-collector connection shows the highest input resistance and the lowest output resistance. It is somewhat analogous to a cathode follows.

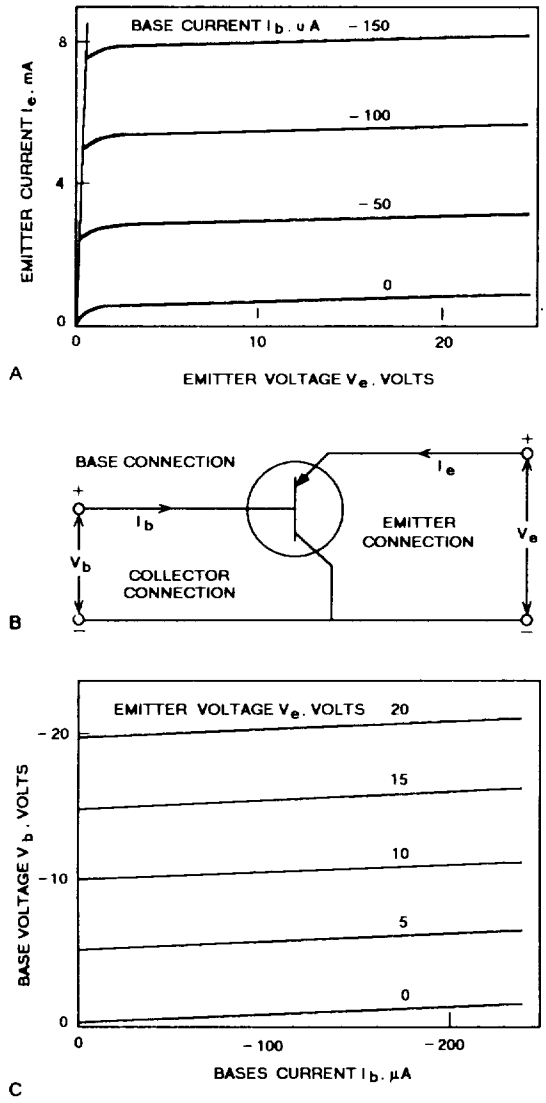


Figure 2-8.-Common-collector connection of an alloy-junction transistor; (a) emitter characteristics; (b) schematic diagram; (c) base characteristics.

Switching circuits or pulse circuits can be monostable, bistable, or astable. The input characteristic must have a negative-resistance region to be useful in such circuits. Such a characteristic may be achieved by a single transistor if it has a current gain a) 1.0. Four-region transistors have such an α and show a negative-resistance region in the base I_b versus V_b characteristic directly, or in the emitter I_e versus V_e characteristic if there is a sufficiently high resistance to the base circuit. It is customary to use more than one transistor to achieve a negative-resistance characteristic, with the exception of the controlled rectifier device used in power switching. The operation of the circuit is *monostable* if the load line intersects this characteristic in only one point on one of the positive-resistance branches of the curve, *bistable* if the load line intersects the curve in two such points, and *astable* if it intersects the curve in only one point in the negative-resistance region. The bistable circuit finds wide application in counters and computers. If transistors are used which have α < 1.0, two transistors are required for each bistable switching circuit.

Complementary symmetry is the use of both *pnp* and *npn* transistors together to take full advantage of their opposite bias and signal polarities. For example, an emitter-follower circuit with base input and grounded collector provides a positive drive to a load for a negative base signal in the case of a *pnp* transistor, and for a positive base signal in the case of an *npn* transistor. If the two are connected in parallel, they give a positive drive to a load with either polarity of input signal.

FET TRANSISTOR CONNECTIONS.— By far the largest use of field-effect transistors (FET.) today is in the large-scale integration of computer memory and logic circuits. In particular, the n-channel MOSFET technology (NMOS) prevails. The basic circuit most used in this technology is the inverter circuit, shown in

figure 2-9, view A. In this circuit a depletion-mode MOSFET is used to load an enhancement-mode MOSFET switching device. The switching device is designated Q, and is shown at the bottom of figure 2-9, view A. The load device is designated *L*, and is shown at the top of figure 2-9, view A. The grounded arrows indicate that the substrates of both devices are grounded The two states of the inverter are given in figure 2-9, view B, and are shown in the loaded drain characteristic diagram shown in figure 2-9, view C.

The nature of an inverter circuit is that if the input voltage goes up the output voltage goes down, and vice versa. Figure 2-9, view B, illustrates this. Considering the circuit of the inverter (fig. 2-9, view B), it can be seen that Q and *L* are in series between the supply voltage V_{DD} and ground. The load device is always conducting because it is a depletion-mode device and because its gate is permanently connected to its source. The switching device may be either conducting or nonconducting, depending on the input signal *V;* on its gate terminal. When V_i is positive, electrons are collected in the channel of *Q,* and it is conducting. When conducting, the channel resistance of *Q* is much lower than that of *L*. This means that the output voltage V_o is held just above ground. When V_i *is* nearly zero, the switching device is not conducting and the conducting channel of *L* holds V_o just below the positive supply voltage V_{DD}. The circuit thus fulfills the criterion for inverter action. This behavior is illustrated in figure 2-9, view C. Here the drain characteristic of the load device is drawn as a nonlinear load line on the drain characteristic curves of the switching device. This load line is marked *L* device. The intersection of the load line with the operating characteristic of the Q device determines the quiescent point of the circuit.

For a switching inverter there are two quiescent points. One, where the Q device is not conducting is designated A in figure 2-9, view C, and is called the off-state. The second, where the Q device is conducting is designated B in figure 2-9, view C, and is called the on-state. There is a single load line while there are several curves in the family of the switching characteristics. The reason for this is that the gate of the load device is connected to its source and cannot change its voltage relative to the source, whereas the gate of the switching device can take on any value of input (gate-source) voltage. In the circuit shown, however, the Q device gate voltage moves between the limits of zero and $+V_{DD}$. Some intermediate gate voltage curves are shown as a reminder that there area multiplicity of states of the inverter between the off-state and the on-state, and that considerable power may be dissipated during the switching process. The off-state (A) has negligible standby power drain. The on-state (B) dissipates typically about 0.1 mW. The switching time ratio (pull-up time to pull-down time) is about 4 to 1, and the total switching delay time of a pair of inverters is approximately 20 nanoseconds.

In small-scale integrated circuit chip components, it is customary to use complementary MOSFET devices (CMOS). In such circuits both n-channel and p-channel devices are used together, one as the load of the other. The use of complementary devices this way greatly reduces standby power to about 10 nanowatts.

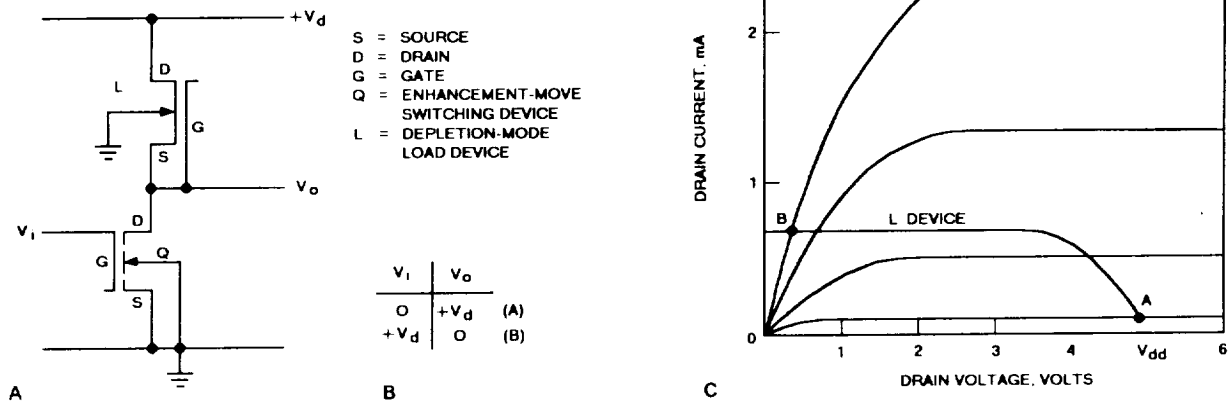


Figure 2-9.-Typical n-channel MOSFET inverter; (a) circuits; (b) truth table for the circuit; (c) loaded drain characteristics.

THYRISTORS

A semiconductor thyristor is a three-terminal semiconductor switching device with separate input (control) and output (load) circuits. Relatively low control current causes the output section of the thyristor to be turned ON, allowing high current to flow in the load. Once the device is turned ON, the input section no longer has control of the device. Turn-off is controlled only by the output circuit supply voltage. The two major components in the thyristors family are the *silicon-controlled rectifier* (SCR) and the *bidirectional triode ac switch* (TRIAC). Thyristors are generally called ac *switches* and are used in a variety of power applications.

Silicon-Controlled Rectifier (SCR)

The SCR is a triode reverse-blocking thyristor. Whereas diodes use two alternate layers of *pn*- type semiconductor material and transistors use three such layers, thyristor devices use four layers, forming three or more junctions within a slice of silicon semiconductor material. Thyristor devices exhibit regenerative, or latching-type, switching action in one or two quadrants of their volt-ampere characteristic. They can be switched into the ON state (conducting condition), but must usually be restored to their OFF state (voltage-blocking condition) by circuit action. Figure 2-10 shows the different types of thyristor devices.

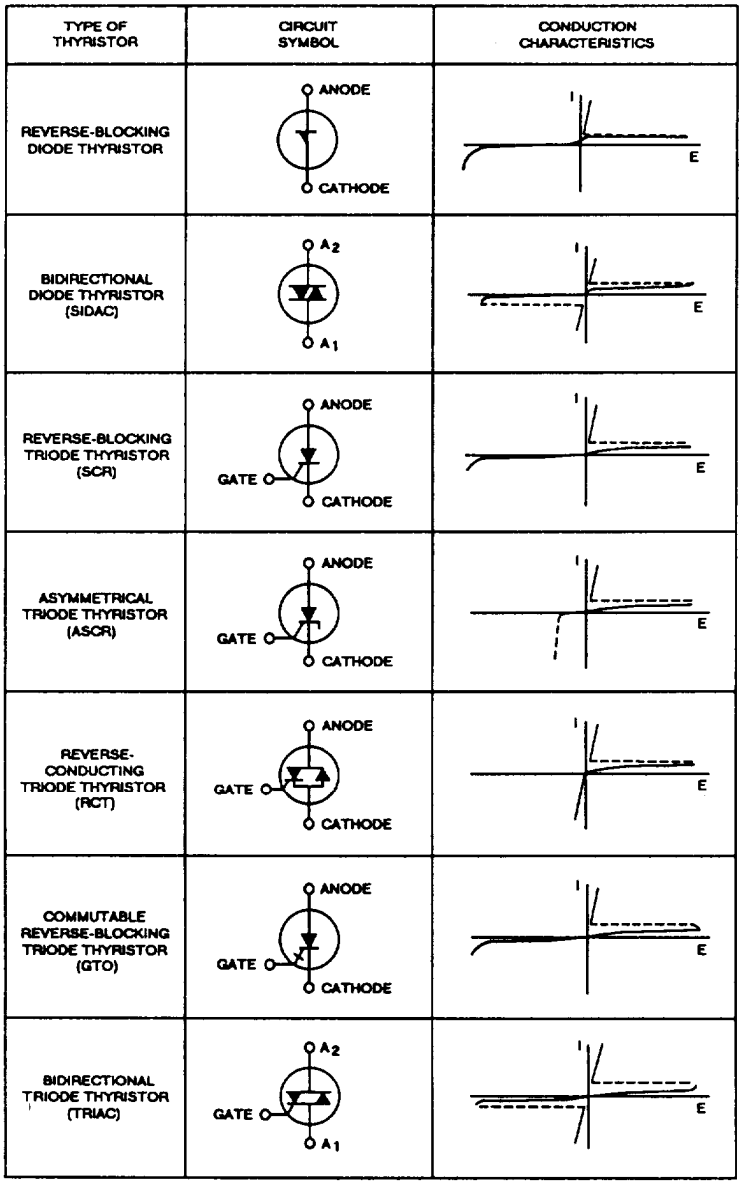


Figure 2-10.-Types of thyristor devices.

TIM most widely used of all thyristor devices for power control is the SCR. For specialized ac-switching power control, such as in lamp dimmers and heating controls, the bidirectional triode thyristor, popularly called the TRIAC, has also come into widespread usage.

Figure 2-11 shows a typical arrangement of alternate *p* and *n* layers in an SCR structure. The thickness is exaggerated in proportion for clarity of illustration. With positive voltage on its cathode with respect to its anode, the SCR blocks the flow of reverse current in a manner similar to that of a conventional silicon rectifier diode. When the voltage is reversed, the SCR blocks forward current flow until a low-power trigger signal is applied between the gate terminal and the cathode. The SCR switches into a highly conductive state with a voltage drop of approximately 1 V between anode and cathode similar to that of the rectifier diode. Once in conduction, the SCR continues to conduct even after the gate signal is removed, provided the anode (or load) current remains above the holding current level, typically in the order of milliamperes. If anode current momentarily drops below the holding current level or if the anode voltage is momentarily reversed, the SCR reverts to its blocking state, and the gate terminal regains control. Typical SCRs turn on in about 1 to 5 microseconds, and require 10 to 100 microseconds of momentary reverse voltage on the anode to regain their forward-blocking ability. The details of the ON and OFF switching characteristics of SCRs vary with different types made for varying applications.

Anode voltage applied to the SCR significantly in excess of the voltage rating of the SCR can trigger the device into conduction even in the absence of a gate signal. Excess reverse voltage, however, can permanently damage the SCR, such as in the case of the silicon rectifier diode. SCRs, like the silicon diode and all power semiconductors, have a failure mechanism called "thermal fatigue." Thermal fatigue failure is due to the thermal stresses induced during repetitive temperature changes occurring in the normal operation of the device. These stresses are inherent in all devices undergoing substantial temperature changes that contain dissimilar metals. When power semiconductors, such as rectifier diodes and SCRs, are properly applied to take into account their thermal fatigue limitations, they can be expected to perform their function faultlessly for the life of the equipment in which they are used.

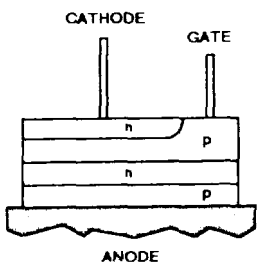


Figure 2-11.-Typical SCR structure.

Current ratings of SCRs range from under 1 to 5000 A. Blocking voltage capability of commercially available devices extends to 4400 V for the higher-power types, with voltages up to 6 kV having been demonstrated in the laboratory.

Like most semiconductor devices, SCRs are dependent on temperature in some of their characteristics. Usual operating junction temperatures are 125°C (257°F), and some devices are available up to 150°C (302°F).

The mounting considerations for SCRs are similar to those for diodes. Small devices are lead-mounted, and above 2 and 4 A SCRs are generally mounted to radiating fins or some type of heat sink for adequate cooling of the semiconductor junction.

Processing techniques used in SCRs are similar to and extensions of the processing used for silicon diodes. In addition to alloy and diffusion processing technology, epitaxial processing is sometimes used. In small devices, a planar structure such as that developed for signal transistors and monolithic integrated circuits is used. Higher-power SCR structures are of a mesa type of construction, with the edges of the pellet often shaped in a manner to reduce the surface field across the blocking junction for higher voltage-blocking capability.

SCR applications fall into two general categories. In one category the devices are used from an ac supply, much as the silicon rectifier diode is used. However, unlike the rectifier diode, which conducts load current as soon as the anode voltage assumes a positive value, the SCR will not conduct load current until it is triggered into conduction. If, when applied in rectifier circuits, conduction through the SCR is delayed from the point of the natural zero crossing of the forward anode-to-cathode voltage, the power delivered to the load can be varied. This mode of control is referred to as ac phase control. It is used extensively in ac to variable-voltage dc output types of applications. The circuit shown in figure 2-12 is the parallel-inverse, or at-switch, circuit that can supply variable voltage to ac loads. It is used extensively in lighting and heating control.

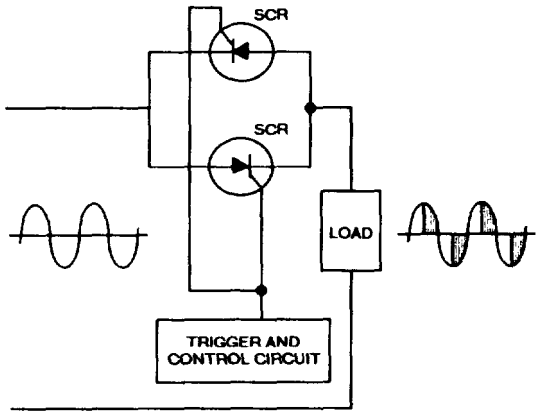


Figure 2-12.-Phase-controlled SCR ac switch circuit.

Turnoff of the SCR in ac circuits is accomplished by the reversal in voltage polarity of the ac supply line. Since the SCR, like the diode, blocks the flow of reverse current, there is no flow of load current during the half-cycle of applied line voltage, which places reverse bias across the SCR.

The other basic category of application for SCRs from a circuit point of view is operation from a dc supply. The major distinguishing feature of dc operation from ac operation is that there is no reversal of supply voltage polarity for turnoff of the SCR, which would allow the gate electrode to regain control of the device. Therefore, auxiliary circuit means must be employed to effect turnoff of the SCR. One common way to accomplish this is to switch a previously charged capacitor across the load-carrying SCR in such a reamer that the voltage on the capacitor reverse-biases the SCR sufficiently to reduce the load current through the SCR to zero, and then to allow the device a short time (about 10 to 50 μs) before reapplying forward blocking voltage to the anode of the SCR.

ASCRs, RCTs, and GTOs

These devices are all in the thyristor family and are mainly used in place of SCRs in power circuits requiring operation from a dc source. The ASCR, asymmetrical silicon-controlled rectifier, and RCT, reverse-conducting thyristor, have the advantage of faster turnoff time than the SCR; that is, 5 to 25 μs, and thus require a less costly auxiliary circuit to effect turnoff. The RCT has an added circuit advantage, as it has a built-in reverse rectifier diode in parallel with the device. (The RCT is the integrated equivalent of a discrete ASCR in parallel with a discrete rectifier diode.) Along with faster turnoff times, the ASCR and RCT devices have lower forward-voltage drops for comparable forward-blocking voltage ratings and silicon area, thus increasing the device's efficiency.

The GTO, gate turnoff device, is also a thyristor. Like the SCR, it is a symmetrical reverse-blocking triode thyristor (unlike the ASCR and RCT, which cannot block reverse voltages), but it has the added advantage of being able to turn off current when a negative signal is applied to the gate. (Voltage and current characteristics are given in figure 2-10.) Thus, the GTO does not require an auxiliary circuit to communicate it off as do the SCR, ASCR, and RCT devices. The added complexity of gate turnoff makes the GTO higher priced than similarly rated SCRs.

The GTO), ASCR, and RCT all came into commercial usage during the latter half of the 1970s, primarily at voltage levels greater than 800 V and currents exceeding 25 A. It is expected that they will soon be manufactured in all current and voltage ratings that are presently available in SCRs above the lower values stated. It is projected that the greatest utilization will be of voltage ratings exceeding 1400 V and currents exceeding 100 A. Figure 2-13 shows a rating comparison of silicon rectifier diodes, thyristors, and Schottky barrier diodes.

It is necessary to operate thyristors from a dc supply in order to achieve power conversion from the dc (battery or rectified ac line) supply to a load requiring an alternating supply (dc to ac inversion) or to a load requiring a variable-voltage dc supply (dc to dc conversion). Since the rate of switching the thyristors in dc circuits can be varied by the control circuit, a thyristor inverter circuit can supply ac load with a variable

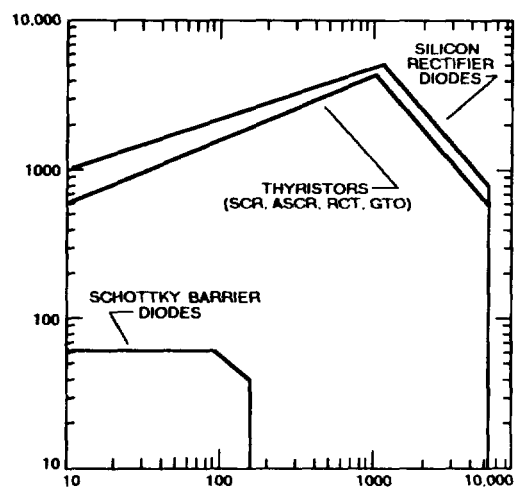


Figure 2-13.-Rating comparison of silicon rectifier diodes, thyristors, and Schottky barrier diodes.

frequency. An important application of this mode of operation is for adjustable speed operation of ac synchronous and induction motors.

A battery source can be converted to a variable-voltage dc source for a dc motor by "chopping" the dc source voltage either at a variable rate at constant pulse width (frequency power modulation) or by operating the chopper circuit at a constant frequency and varying the pulse width (pulse-width power modulation).

INTEGRATED CIRCUITS

Learning Objective: *Identify integrated circuits, their construction, and the types of circuits they produce.*

Miniature electronic circuits are produced within and upon a single semiconductor crystal, usually silicon. Integrated circuits range in complexity from simple logic circuits and amplifiers, about 1/20 inch (1.3 mm) square to large-scale integrated circuits up to about 1/3 inch (8 mm) square. They contain hundreds of thousands of transistors and other components that provide computer memory circuits and complex logic subsystems such as microcomputer central processor units.

Since the mid-1960s, integrated circuits have become the primary components of most electronic systems. Their low cost, high reliability, and speed have been essential in furthering the wide use of digital computers. Microcomputers have spread the use of computer technology to instruments, business machines, automobiles, and other equipment. For analog signal processing, integrated subsystems such as FM stereo demodulators and switched-capacitor filters are made.

Integrated circuits consist of the combination of active electronic devices, such as transistors and diodes, with passive components, such as resistors and capacitors, within and upon a single semiconductor crystal. The construction of these elements within the semiconductor is achieved through the introduction of electrically active impurities into well-defined regions of the semiconductor. The fabrication of integrated circuits thus involves such processes as vapor-phase deposition of semiconductors and insulators, oxidation, solid-stage diffusion, ion implantation, vacuum deposition, and sputtering.

Generally, integrated circuits are not straight-forward replacements of electronic circuits assembled from discrete components. They represent an extension of the technology by which silicon planar transistors are made. Because of this, transistors or modifications of transistor structures are the primary devices of integrated circuits. Methods of fabricating good quality resistors and capacitors have been devised, but the third major type of passive component, inductors, must be simulated with complex circuitry or added to the integrated circuit as discrete components.

Simple logic circuits were the easiest to adapt to these design changes. The first of these, such as inverters and gates, were produced in the early 1960s primarily for miniaturization of missile guidance computers and other aerospace systems. Analog circuits, called linear integrated circuits, did not become commercially practical until several years later because of their heavy dependence on passive components such as resistors and capacitors. The first good quality operational amplifiers for analog computers and instruments were produced in 1966.

TYPES OF INTEGRATED SILICON CIRCUITS

Integrated silicon circuits can be classified into two groups on the basis of the type of transistors they employ. They are the bipolar integrated circuits, in which the principal element is the bipolar junction transistor; and metal oxide semiconductor (MOS) integrated circuits, in which the principal element is the MOS transistor. Both depend upon the construction of a desired pattern of electrically active impurities within the semiconductor body, and upon the formation of an interconnection pattern of metal films on the surface of the semiconductor.

Bipolar circuits are generally used where highest logic speed is desired, and MOS for largest-scale integration or lowest-power dissipation. Linear circuits are mostly bipolar, but MOS devices are used extensively in switched-capacitor filters.

Bipolar Circuits

A simple bipolar inverter circuit using a diffused resistor and an *npn* transistor is shown in figure 2-14. The input voltage V_{in} is applied to the base of the transistor. When V_{in} is zero or negative with respect to the emitter, no current flows. As a result, no voltage drop exists across the resistor, and the output voltage V_{out} will be the same as the externally applied biasing voltage, +5 volts in this example. When a positive input voltage is applied, the transistor becomes conducting. Current now flows through the transistor, hence through the

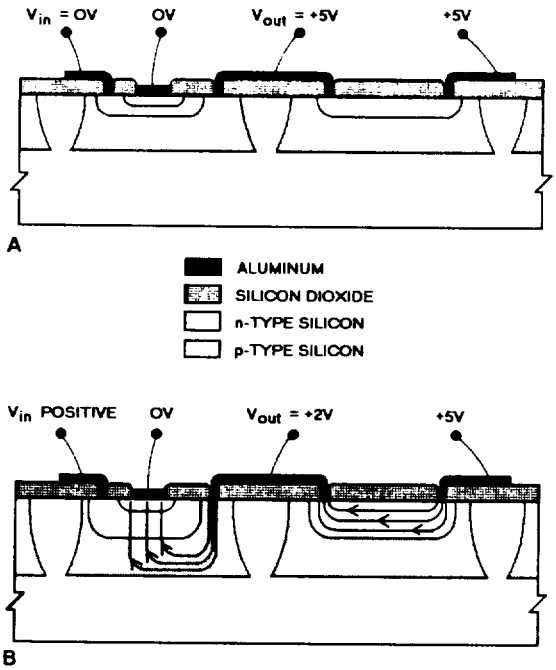


Figure 2-14.-Operation of bipolar Inverter circuit.

resistor; as a result, the output voltage decreases. Thus, the change in input voltage appears inverted at the output.

The circuit symbol and the changes in input and output voltages during the switching process just described are illustrated in figure 2-15. The change in the output voltage occurs slightly later than the change in the input voltage. This time difference, called propagation delay, is an important characteristic of all integrated circuits. Much effort has been spent on reducing it, and values less than one-billionth of a second have been achieved.

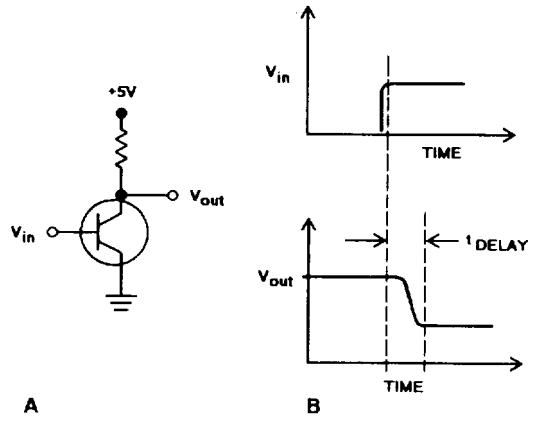


Figure 2-15.-Inverter circuit, (a) symbol, (b) switching waveforms.

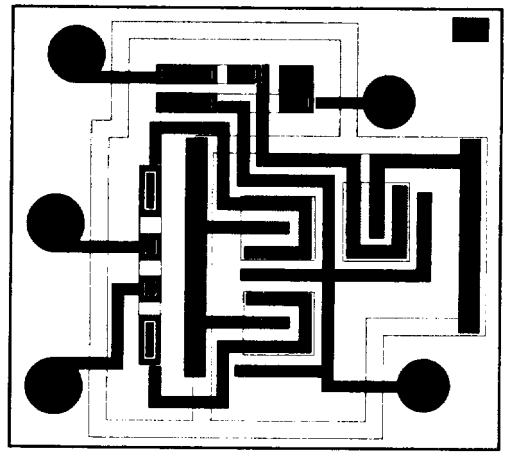


Figure 2-16.-Bipotar logic gate.

Most simple digital circuits can be fabricated, much as the inverter circuit described above. As an example, a photomicrograph of an early logic gate circuit is shown in figure 2-16. This circuit is one of the earliest digital integrated circuits. For comparison, a 16-bit microcomputer digital integrated circuit is shown in figure 2-17. This circuit contains more than 3,000 times as much circuitry, illustrating the tremendous increase in density that has occurred.

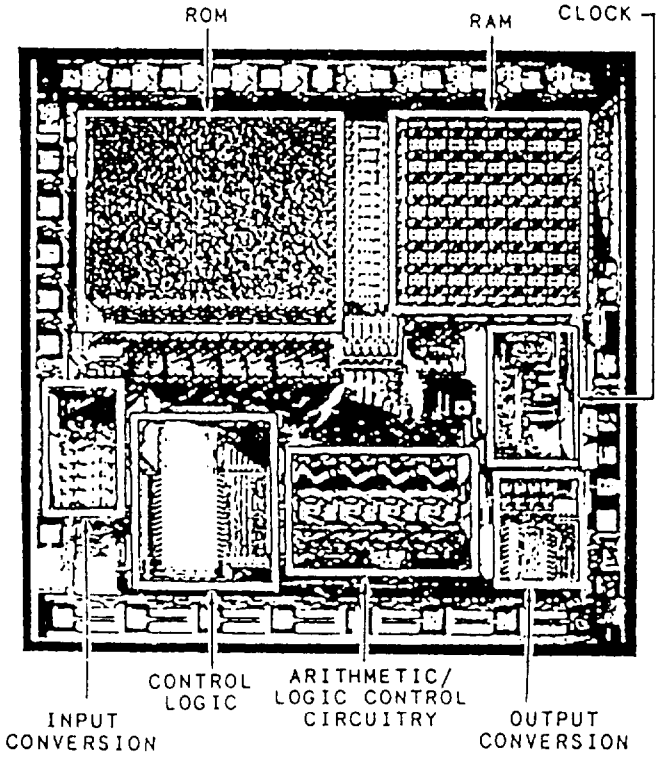


Figure 2-17.-A 16-bit bipolar microprocessor.

This tendency toward increased complexity is dictated by the economics of integrated circuit manufacturing. Because of the nature of this manufacturing process, all circuits on a slice are fabricated together. Consequently, the more circuitry accommodated on a slice, the cheaper the circuitry becomes. Because testing and packaging costs depend on the number of chips, it is desirable, to keep costs down, to crowd more circuitry onto a given chip rather than to increase the number of chips on a wafer.

Linear Circuits

Integrated circuits based on amplifiers are called "linear" because amplifiers usually exhibit a linearly proportional response to input signal variations. However, the category includes memory sense amplifiers, combinations of analog and digital processing functions, and other circuits with nonlinear characteristics. Some digital and analog combinations include analog-to-digital converters, timing controls, and modems (data communications modulator-demodulator units).

Along-standing drawback in these circuits was the lack of inductors for tuning and filtering. That was overcome by the use of resistor-capacitor networks and additional circuitry. For low-frequency circuits, the resistor in these networks is being replaced by the switched capacitor. At the higher frequencies, an oscillator-based circuit known as the phase-locked loop provides a general-purpose replacement for inductors in applications such as radio transmission demodulation.

At first, the development of linear circuits was slow because of the difficulty of integrating passive components, and also because of undesirable interactions between the semiconductor substrate and the operating components. Thus, much greater ingenuity was required to design and use the early linear circuits.

In addition, manufacturing economics favors digital circuits. A computer can be built by repetitious use of simple inverters and gates, while analog signal processing requires a variety of specialized linear circuits.

MOS Circuits

The other major class of integrated circuits is called MOS because its principal device is a metal oxide semiconductor field-effect transistor (MOSFET). It is more suitable for very large-scale integration (VLSI) than bipolar circuits because MOS transistors are self-isolating and can have an average size of less than a millionth of a square inch (5×10^{-5}). This has made it practical to use over one million transistors per circuit. Because of this high density capability, MOS transistors are used for high density random-access memories (RAMs) (fig. 2-18), read-only memories (RAMs), and microprocessors.

Several major types of MOS device fabrication technologies have been developed since the mid- 1960s. They are (1) metal-gate *p*- channel MOS (PMOS), which uses aluminum for electrodes and interconnections; (2) silicon-gate p-channel MOS, employing polycrystalline silicon for gate electrodes and the first interconnection layer; (3) n-channel MOS (NMOS), which is usually silicon gate; and (4) complementary MOS (CMOS), which employs both *p*- channel and *n*- channel devices. In 1984 the silicon gate NMOS and CMOS were the

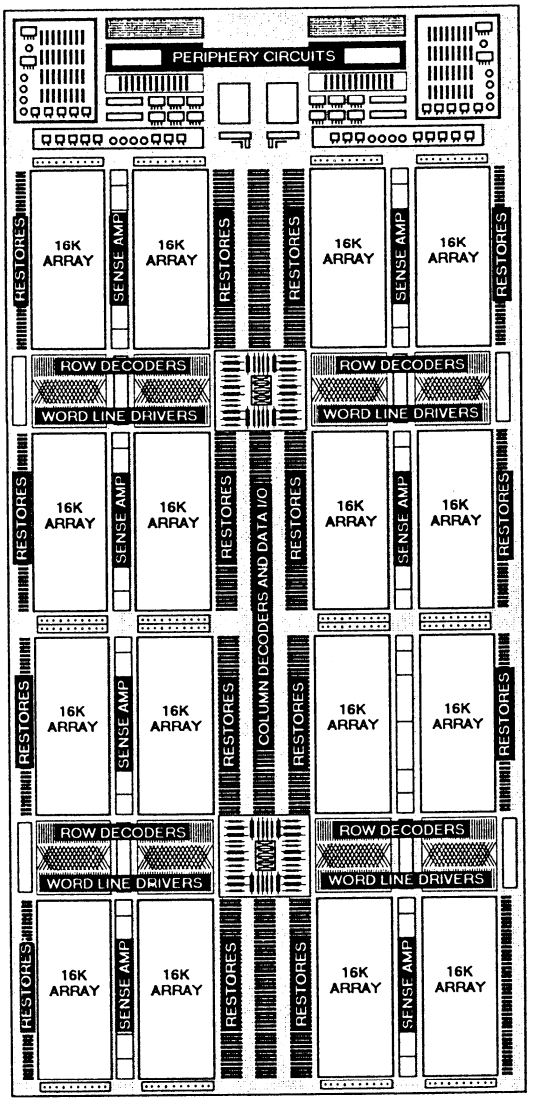


Figure 2-18.—MOS LSI ram circuit.

dominant technologies, with CMOS using silicon gates and becoming the most attractive for new designs.

Both conceptually and structurally the MOS transistor is a much simpler device than the bipolar transistor. In fact, its principle of operation has been known since the late 1930s, and the research effort that led to the discovery of the bipolar transistor was originally aimed at developing the MOS transistor. This simple device was kept from commercial use until 1964 because it depends on the properties of the semiconductor surface for its operation, while the bipolar transistor depends principally on the bulk properties of the semiconductor crystal. Hence, MOS transistors became practical only when understanding and control of the properties of the oxidized silicon surface had been perfected to a very great degree.

CMOS Circuits

A simple CMOS inverter circuit is shown in figure 2-19, and a circuit schematic is shown in

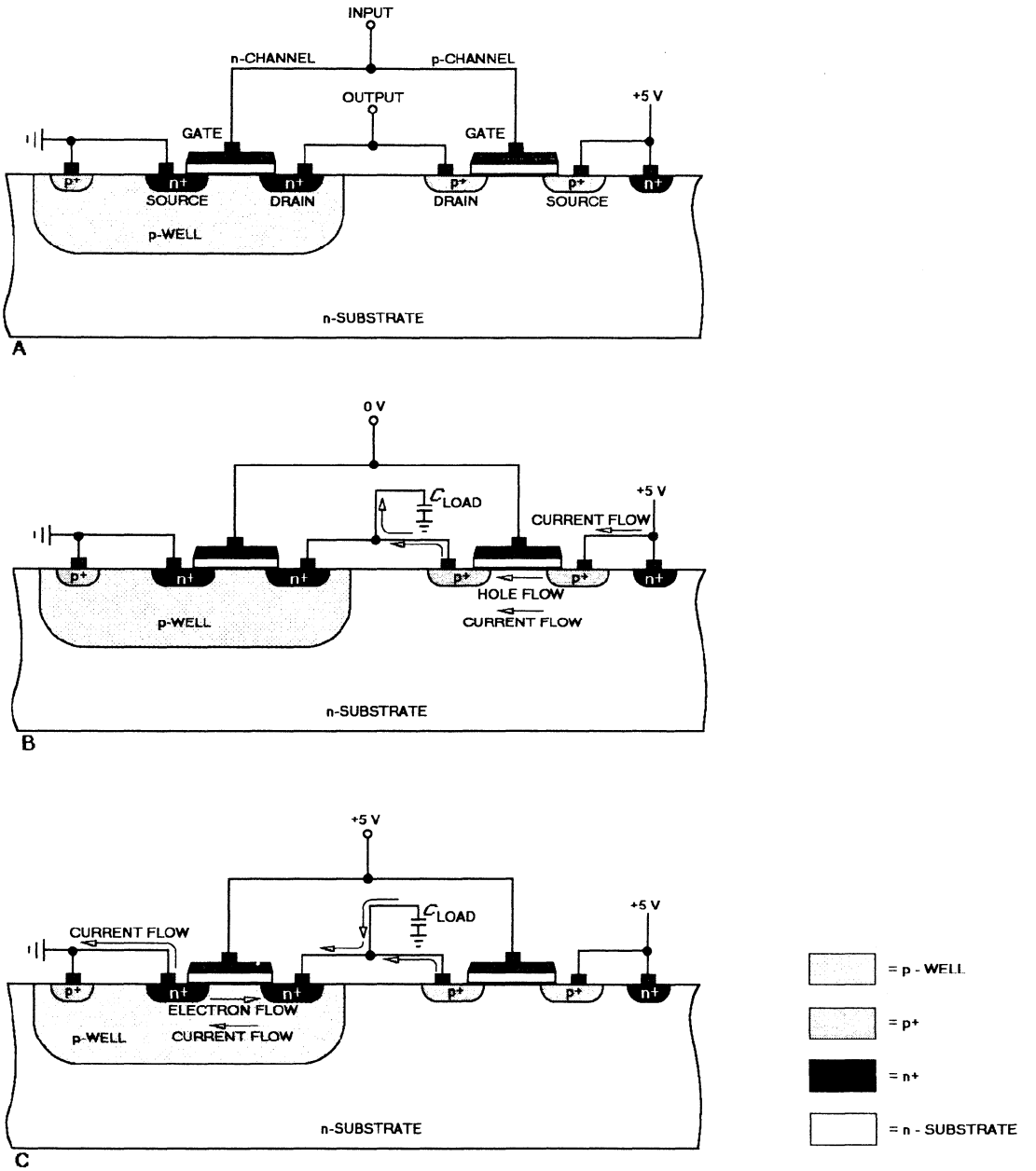


Figure 2-19.—CMOS inverter circuit; (a) schematic cross section; (b) current flow when input is "low" at OV; (c) current flow when input is "high" at 5 V.

figure 2-20. The gates of the n-channel and p-channel transistors are connected together as are the drains. The common gate connection is the input node while the common drain connection is the output node. A capacitor is added to the output node to model the loading expected from the subsequent stages on typical circuits.

When the input node is in the "low state," at 0 V, the n-channel gate to source voltage is 0 V while the p-channel gate to source voltage is -5 V. The *n*- channel transistor requires a positive gate-to-source voltage, which is greater than the transistor threshold voltage (typically 0.5-1 V), before it will start conducting current between the drain and source. Thus, with a 0-V gate-to-source voltage, it will be off and no current will flow through the drain and source regions. The p-channel transistor, however, requires a negative voltage between the gate and source which is less than its threshold voltage (typically -0.5 to -1.5 V). The -5 volt gate-to-source potential is clearly less than the threshold voltage, and the *p*- channel will be turned on, conducting current from the source to the drain, and thereby charging up the loading capacitor. Once the capacitor is charged to the "high state" at 5 V, the transistor will no longer conduct because there will no longer be a potential difference between the source and drain regions.

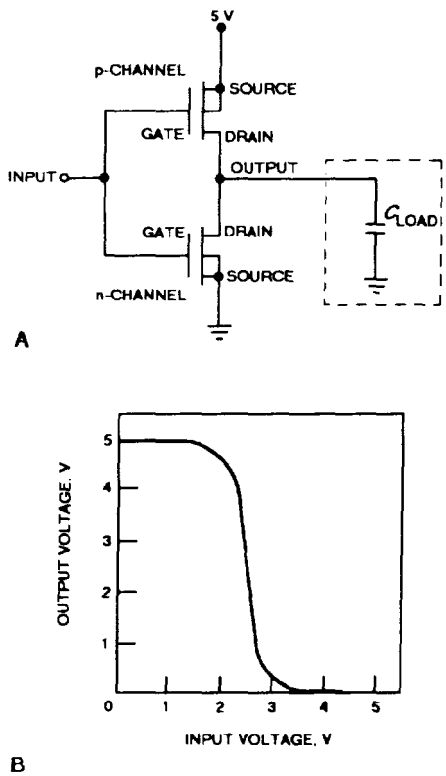


Figure 2-20.-MOS inverter circuit (a) circuit diagram; (b) transfer characteristics.

When the input is now put to the "high state" at 5 V, just the opposite occurs. The n-channel transistor will be turned on while the p-channel will be off. This will allow the load capacitor to discharge through the n-channel transistor, resulting in the output voltage dropping from a "high state" at 5 V to a "low state" at 0 V. Again, once there is no potential difference between the drain and source (capacitor discharged to 0 V), the current flow will stop, and the circuit will be stable.

This simple circuit illustrates a very important feature of CMOS circuits. Once the loading capacitor has been either charged to 5 V or discharged back to 0 V, there is no current flow, and the standby power is very low. This is the reason for the high popularity of CMOS for battery-based systems. None of the other MOS technologies offers this feature without complex circuit techniques, and even then will typically not match the low standby power of CMOS. The bipolar circuits discussed above require even more power than these other MOS technologies. The price for CMOS's lower power are the additional fabrication steps required (10 to 20 percent more) when compared to NMOS.

Sampled-Data Device Circuits

In addition to the digital logic applications discussed above with the simple CMOS inverter circuit, MOS devices also offer unique features for some analog circuit applications. These include signal-processing applications that are based on sampled-data techniques. Two classes of devices, the charge-coupled devices (CCD) and switched-capacitor networks, play the major role in these applications.

In CCDs the stored charge at the semiconductor surface can also be made to propagate along the surface via potential wells created by a series of these MOS structures. The storage cell in the RAM circuit shown in figure 2-18 can be viewed as using a single CCD element for each bit.

The capacitance C from the MOS structure can be integrated with MOS transistors, which are used as switches, to form a switched-capacitor circuit (fig. 2-21). One of the switches 1 is closed when the other switch 2 is open, and vice versa. The circuit is equivalent to a resistor with a value of $R = T/C$, where T is the sampling interval. The advantage of the technique is the high quality and the high value of resistors that can be put on an integrated circuit. These resistors combined

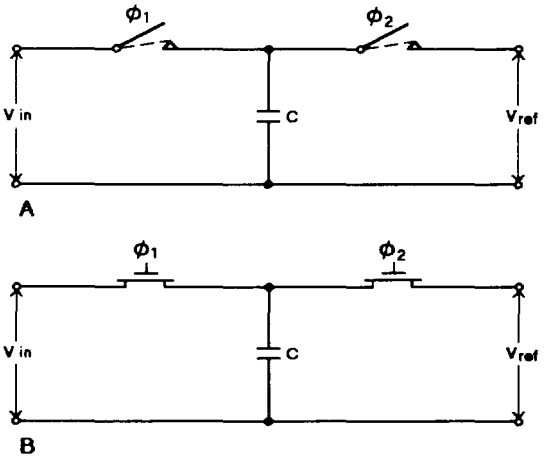


Figure 2-21.-Switched-capacitor circuit; (a) schematic of RC section; (b) implementation of the switches with MOS transistors with clocks applied to their gates.

with capacitors and operational amplifiers (all made with MOS technology) can be used to make active filters. Since many stages of these filters can be integrated into one chip, precise filters are now possible with integrated circuit technology. These filters are having a tremendous impact on low-frequency filters (frequencies less than 100 kHz), and will be used extensively in telecommunication equipment.

Transistor-Transistor Logic Devices (TTL)

Perhaps the best-known and most widely used implementation of logic switches is the bipolar transistor-transistor logic, or TTL. Shown schematically in figure 2-22, the basic TTL NAND gate is formed by a multiemitter transistor (turned on only if every input is high), followed by an output transistor that acts as a pullup/buffer. Thus, the first transistor performs an AND operation on the inputs, and the second transistor completes the NAND by performing an inversion.

TTL transistors are operated in the saturation mode; in other words, the transistors are driven hard to either the cutoff or the saturation limits. This overdriving introduces a time delay that does not exist if the transistors are operated in the nonsaturated mode. Such nonsaturating logic, while inherently faster, is more susceptible to noise since it is biased in the linear region.

The bulk of the available TTL is in the 5400-7400 series. The 7400 line is lower in cost and is useful over 0 to 70 degrees Celsius. The 5400 series is a military line good from -55 to +125 degrees Celsius.

There are several different varieties of TTL that have specific and special uses. They are as follows:

- Regular TTL,
- Low-Power TTL,
- High-Power TTL,
- Schottky TTL, and
- Low-Power Schottky TTL.

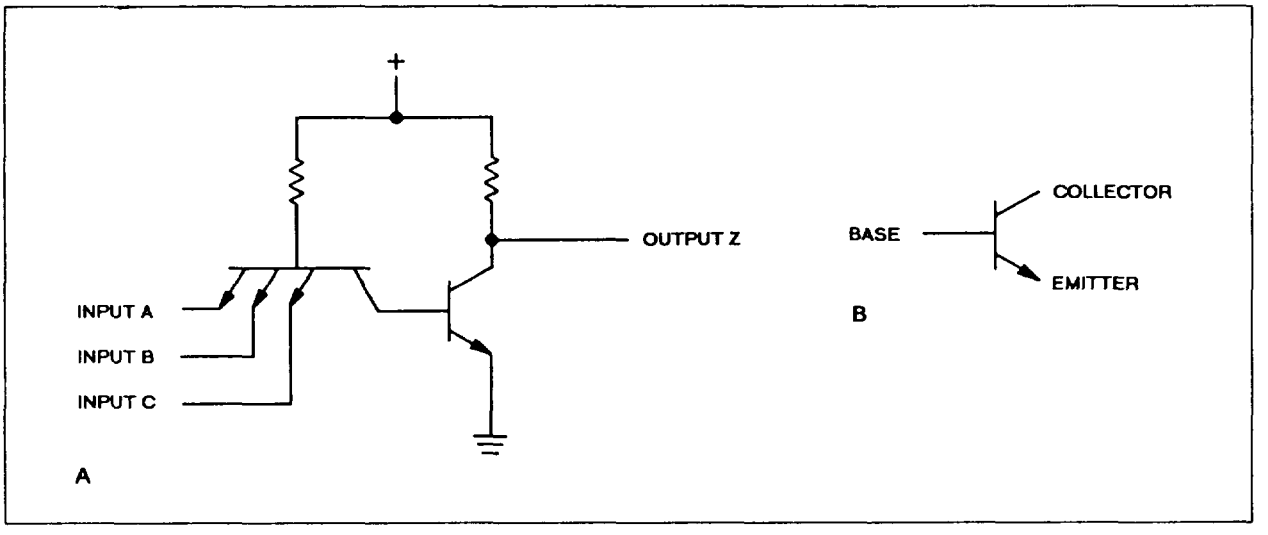


Figure 2-22.-Basic TTL NAND gate formed by multiemittter transistor.

Table 2-1.-Characteristics of Various TTL Types

Type	Gate Propagation Time	Power per Gate	Maximum Counter Frequency
Regular TTL	10 nanoseconds	10 milliwatts	35 megahertz
High-Power TTL	6 nanoseconds	22 milliwatts	50 megahertz
Low-Power TTL	33 nanoseconds	1 milliwatts	3 megahertz
Schottky TTL	3 nanoseconds	19 milliwatt	125 megahertz
Low-Power Schottky TTL	10 nanoseconds	2 milliwatts	45 megahertz

Table 2-1 compares the typical characteristics of each type, while table 2-2 shows the typical numbering system to identify a given temperature range.

REGULAR TTL.– Regular TTL is normally the widest available and the lowest-priced type of TTL, and it has far and away the greatest variety and second-sourcing. A typical gate-propagation time is 10 nanoseconds; this is the time it takes for a logic change at a gate input to appear as a logic change on the output. Around 10 milliwatts per gate is needed, and counting flip-flops go as high as 35 MHz.

LOW-POWER TTL.– Low-power TTL exchanges power consumption for speed and is identified by an *L* in the part number. For instance, a 74L00 is a low-power, commercial version of the 7400 regular TTL NAND gate. There is roughly a 10:1 tradeoff in the low-power version-one-tenth the speed to counters atone-tenth the power, although the simpler gates run one-fourth the speed on one-tenth the power. Flip-flops and counters have a maximum toggle frequency of 3 MHz or so. Within the low-power subfamily, the output-drive remains 10, but a low-power TTL gate can drive only one regular TTL gate. While the 54L00 and 74L00 series TTL do offer low-power consumption, many of their advantages are being preempted by the CMOS logic families.

Table 2-2.-Numbering System for TTL Types

Types	–55° to +125°C	0° to +70°C
Regular	5400	7400
High-Power	54H00	74H00
Low-Power	54L00	74L00
Schottky	54S00	74S00
Low-Power Schottky	54LS00	74LS00

HIGH-POWER TTL.– The high-power TTL devices are designated with an *H* in the part number; 74H00 is the equivalent of a 7400 gate, and so on. Typically you get twice the speed for twice the power. Counters are good to 50 MHz. Within the high-power subfamily, the output-drive remains at 10, but the input is typically 1.3 times regular TTL loads. Thus, a regular TTL gate can drive at most only seven high-power TTL inputs. High-power TTL is largely being replaced by the newer Schottky TTL, which is faster and draws less supply power. Quite a few high-power devices remain available. One advantage they do have over the Schottky devices is that the outputs are “quieter,” a handy feature in high-speed digital-to-analog converters.

SCHOTTKY TTL.– Schottky TTL is an improved version of TTL that has a better speed/power tradeoff than the older types. To do this, Schottky diodes (a fast diode with a 0.3-volt forward drop) are placed across most of the transistors in the basic TTL gate. This prevents the transistors from saturating, and thus eliminates any storage-time delays inside the transistors. The part numbers have an *S* in them, as in 74S00. Propagation delays of 3 nanoseconds are combined with flip-flops that can run at 125 MHz.

Where high speed is essential, Schottky TTL is a logical choice. Its competitor is the emitter-coupled logic families that in general are much faster, but considerably more difficult to use.

A high-speed, unsaturated logic family, such as Schottky TTL, presents serious restrictions in the type and quality of test equipment you must have to work

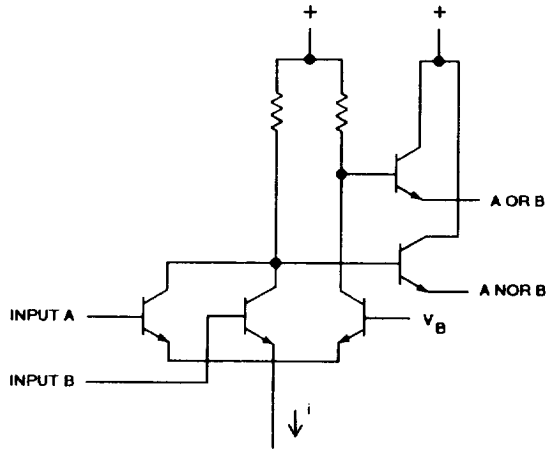


Figure 2-23.-ECL OR/NOR gate.

with it intelligently. A 60-MHz triggered oscilloscope is essential, and a 120-MHz oscilloscope is preferable. As might be expected, Schottky devices are much more critical as to layouts and supply decoupling than ordinary TTL because of their higher speed. Nevertheless, where high speed is essential, they are often the simplest solution to system problems in the 30- to 120-MHz range.

LOW-POWER SCHOTTKY TTL.– Devices such as the 74LS00 are emerging as a more recent variation on TTL. The low-power Schottky TTL family is slightly faster than regular TTL, but requires only one-fifth the power. It does this by using the Schottky diodes to eliminate storage-time effects, but then raises the circuit impedance levels to slow things down to normal and pick up power savings. For many applications, this represents a near-optimum combination of values.

Emitter-Coupled Logic (ECL) Devices

The basic ECL gate is shown in figure 2-23 to be composed of current-steering transistors that perform an operation on the inputs. Typically, the gate output is amplified by an emitter-follower transistor, and both the true and complement signals can be made available with no added delays at the output. This circuit makes the switching of the transistors very fast by never allowing them to turn all the way on, a condition known as saturation, giving us propagation delays of less than 1 nanosecond. This is not normally used in large, complex chips due to its characteristic of drawing large amounts of current, producing large power dissipation.

TYPES OF INTEGRATOR GALLIUM ARSENIDE CIRCUITS

Integrated circuits based on gallium arsenide (GaAs) have come into increasing use since the late 1970s. The major advantage of these circuits is their fast switching speed.

Gallium Arsenide FET

The gallium arsenide field-effect transistor (GaAs FET) is a majority carrier device in which the cross-sectional area of the conducting path of the carriers is varied by the potential applied to the gate (fig. 2-24). Unlike the MOSFET, the gate of the GaAs FET

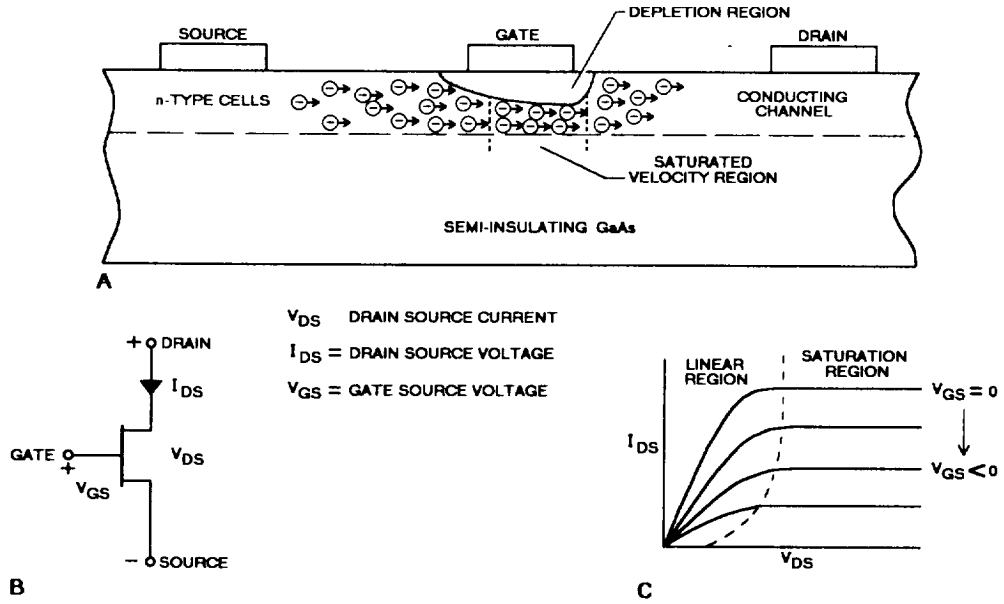


Figure 2-24.-Gallium arsenide FET; (a) cross section; (b) circuit symbol; (c) current-voltage characteristics.

is a Schottky barrier composed of metal and gallium arsenide. Because of the difference in work functions of the two materials, a junction is formed. The depletion region associated with the junction is a function of the difference in voltage of the gate and the conducting channel, and the doping density of the channel. By applying a negative voltage to the gate, the electrons under the gate in the channel arc repelled, extending the depletion region across the conducting channel. The variation in the height of the conducting portion of the channel caused by the change in the extent of the depletion region alters the resistance between the drain and source. Thus, the negative voltage on the gate modulates the current flowing between the drain and the source, as shown by the linear region of operation in figure 2-24, view C. As the height of the conducting channel is decreased by the gate voltage or as the drain voltage is increased, the velocity of charge carriers (electrons for n-type gallium arsenide) under the gate increases (similar to water in a hose when its path is constricted by passing through the nozzle). The velocity of the carriers continues to increase with increasing drain voltage, as does the current, until their saturated velocity is obtained (about 10^7 cm/s or 3×10^5 ft/s for gallium arsenide). At that point the device is in the saturated region of operation; that is, the current is independent of the drain voltage.

The high-frequency operation of a device is limited by the transit time of the carriers under the gate. The time during which the velocity of electrons (output signal) is modulated by the voltage on the gate (input signal) must be short compared to any change of the input voltage. Because electrons in gallium arsenide have a high saturated velocity, GaAs FETs operate at very high frequencies. The high-frequency performance is also improved by decreasing the gate length (the length of the path of the electrons under the gate) by using special lithographic techniques to define the gate during processing. Gallium arsenide FETs with gate lengths as short as 0.1 μm (4×10^6 in.) have been fabricated, resulting in a potential frequency of operation of approximately 100 GHz.

As noted above, the major advantage of gallium arsenide integrated circuits over silicon integrated circuits is the faster switching speed of the logic gate. The reason for the improvement of the switching speed of GaAs FET's with short gate lengths (less than 1 μm or 4×10^{-5} in.) over silicon FET's of comparable size has been the subject of controversy. In essence, the speed or gain-bandwidth product of a FET is determined by the velocity with which the electrons pass under the gate. The saturated drift velocity of electrons in gallium arsenide is twice that of electrons in silicon; therefore, the switching speed of gallium arsenide might be expected to be only twice as fast.

However, this simplified model neglects several important aspects of the problem. One way to determine the switching speed of a logic circuit is to calculate the total capacitance that must be charged or discharged as the logic level is switched, and the current drive available. The larger the current drive and the smaller the capacitance, the faster the switching speed. Since gallium arsenide integrated circuits are fabricated on semi-insulating substrates, the parasitic capacitance to ground is much smaller than for silicon integrated circuits. The only comparable small-capacitance silicon technology is CMOS/SOS (silicon on sapphire). Also, because of the higher mobility, the transconductance of a GaAs FET is much higher than for a silicon FET, and the associated parasitic resistances are lower. Thus, there is more current change for a given amount of input voltage. Finally, the mobility of gallium arsenide is six to eight times that of silicon, and even though the saturated velocities of gallium arsenide and silicon are within a factor of 2, the electrical field necessary for the carriers to reach velocity saturation in gallium arsenide (about 4 kV/cm) is much less than in silicon (about 40 kV/cm). Therefore, when operating at the low voltages typical of GaAs FETs, the speed ratio of similar gallium arsenide and silicon FETs is approximately proportional to their low-field mobilities. At higher voltages, the speed ratio decreases because the carrier velocity (current) continues to increase in silicon, whereas the carriers are saturated in gallium arsenide; however, this increase in speed is at the expense of increased power

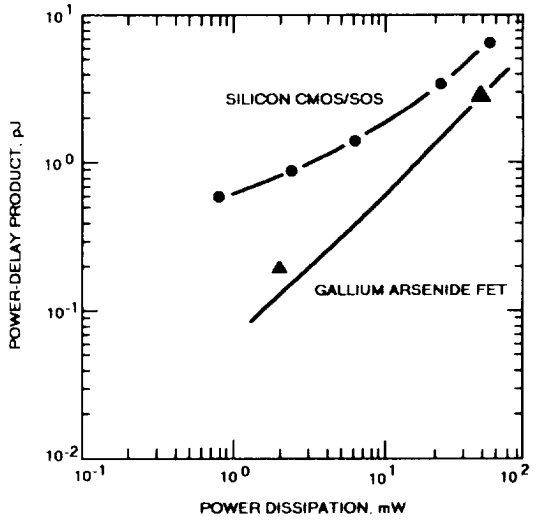


Figure 2-25.-Comparison of gallium arsenide and silicon inverter performance.

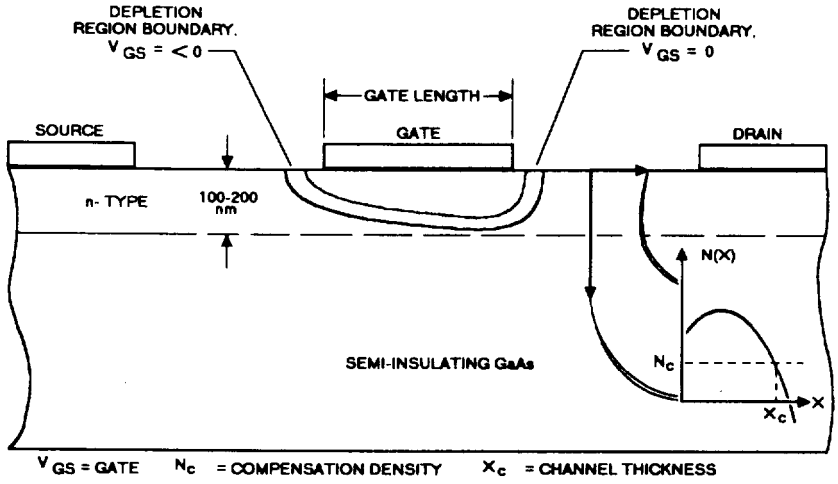


Figure 2-26.-Depletion-mode FET (DFET).

dissipation. This effect explains the experimental results plotted in figure 2-25, where the power-delay products of silicon and gallium arsenide inverters with 1-μm-gate-length (4×10^{-5} in.) FETs are plotted as functions of power dissipation. There are several device choices for high-speed gallium arsenide integrated circuits, each with certain advantages and disadvantages.

Depletion-Mode FET (DFET)

This is the most mature of the device technologies (fig. 2-26). The DFET has the largest current drive capacity per unit device width for an all-GaAs FET device. This contributes to its high speed and high power dissipation. The pinchoff voltage of the DFET is determined by the channel doping and thickness under the Schottky barrier gate. This voltage can be made quite large (about –2.5 V) in order to improve the noise immunity of logic gates in which they are used.

Enhancement-Mode FET (ENFET)

This low-current, low-power device is realized by increasing the pinchoff voltage to zero or above. The logic swing for the ENFET is limited to the difference between the pinchoff voltage (approximately 0 V) and the forward turn-on voltage of the Schottky barrier gate (approximately +0.5 V), thus providing a significantly lower noise immunity for logic gates using ENFETs. The realization of medium-scale integration (MSI) and LIS chips in which the noise margins are small requires stringent process controls to fabricate devices across the wafer with very small variations in pinchoff voltage.

Enhancement Mode Junction FET (E-JFET)

In this device the Schottky barrier of the ENFET is replaced with an implanted *p* region that forms a *pn* junction for the gate (fig. 2-27). The E-JFET has all the advantages of the ENFET with respect to low power,

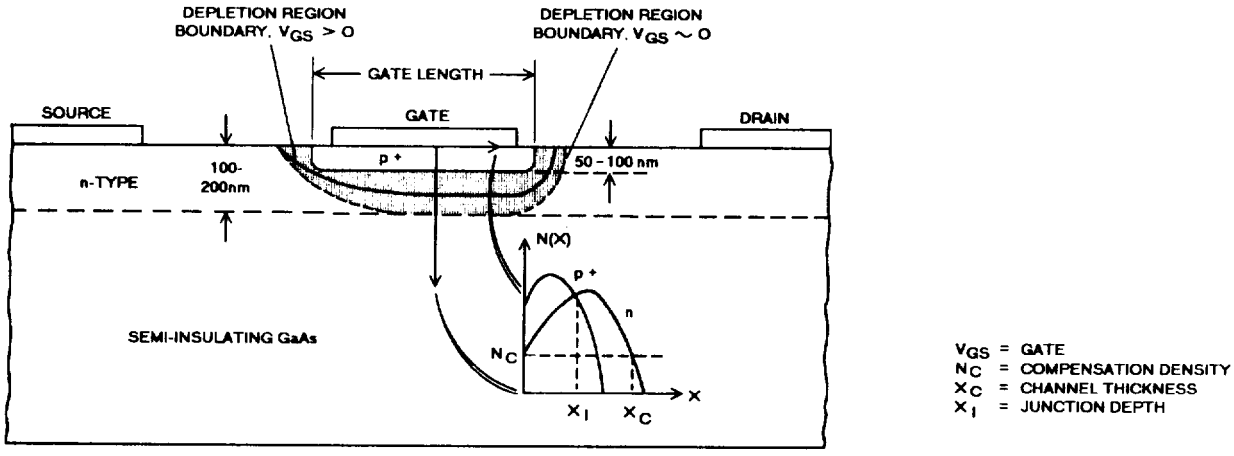


Figure 2-27.-Enhancement-made junction FET (E-JET).

plus the additional advantage of a slightly larger logic swing due to the larger turn-on voltage of the *pn* junction. The ultimate speed of the E-JFET will be less than an ENFET of similar dimensions because the added side wall gate capacitance of the *pn* - junction gate is a significant fraction of the total gate capacitance at submicrometer gate lengths.

LOGIC GATE CONFIGURATIONS

Three different logic gate configurations (fig. 2-28) are presently the most popular approaches to high-speed gallium arsenide logic circuits. The buffered-FET logic (BFL) gate is the fastest gate for reasonable fan-outs but dissipates the most power (approximately 5-10mW/per gate). The Schottky diode FET logic (SDFL) gate dissipates about one-fifth the power of the BFL; however, it is slower by about a factor of 2. Finally, direct-coupled FET logic (DCFL) gates using enhancement-mode FETs have the lowest power consumption (about 50μW/per gate) at gate delays two to four times those of BFL for complex logic circuits.

The BFL gate using DFETs requires level shifting to make the input and output logic levels compatible. This extra circuitry adds both delay to the switching time of the gate and extra power consumption; however, it provides buffering to the next stage, and therefore, has very good fan-out (fan-out is the number of identical logic gates it must drive) and on-off chip drive capabilities. Because of the high power dissipation and the huge device count per gate, BFL will not be suitable for circuits with the complexity of large scale integration (greater than 1000 gates).

The SDFL gate incorporates very small Schottky barrier diodes (1 x 2 μm or 4 x 10^{-5} by 8 x 10^{-5} in.) to perform the input logical OR function and to provide level shifting. The invert function is preformed by the DFETs in the second stage. Because of the lower power dissipation and small diodes, packing densities of more than 1000 gates/mm^2 (645,000 gates/in.^{2}) are achievable. Large fan-in does not require any significant chip area because of the small diodes; however, SDFL gates are very fan-out sensitive, and for fan-outs of greater than 3, either buffers or much wider DFETs must be incorporated to maintain the speed. Because of the medium power dissipation and high packing density, SDFL is suitable for large-scale integration applications,

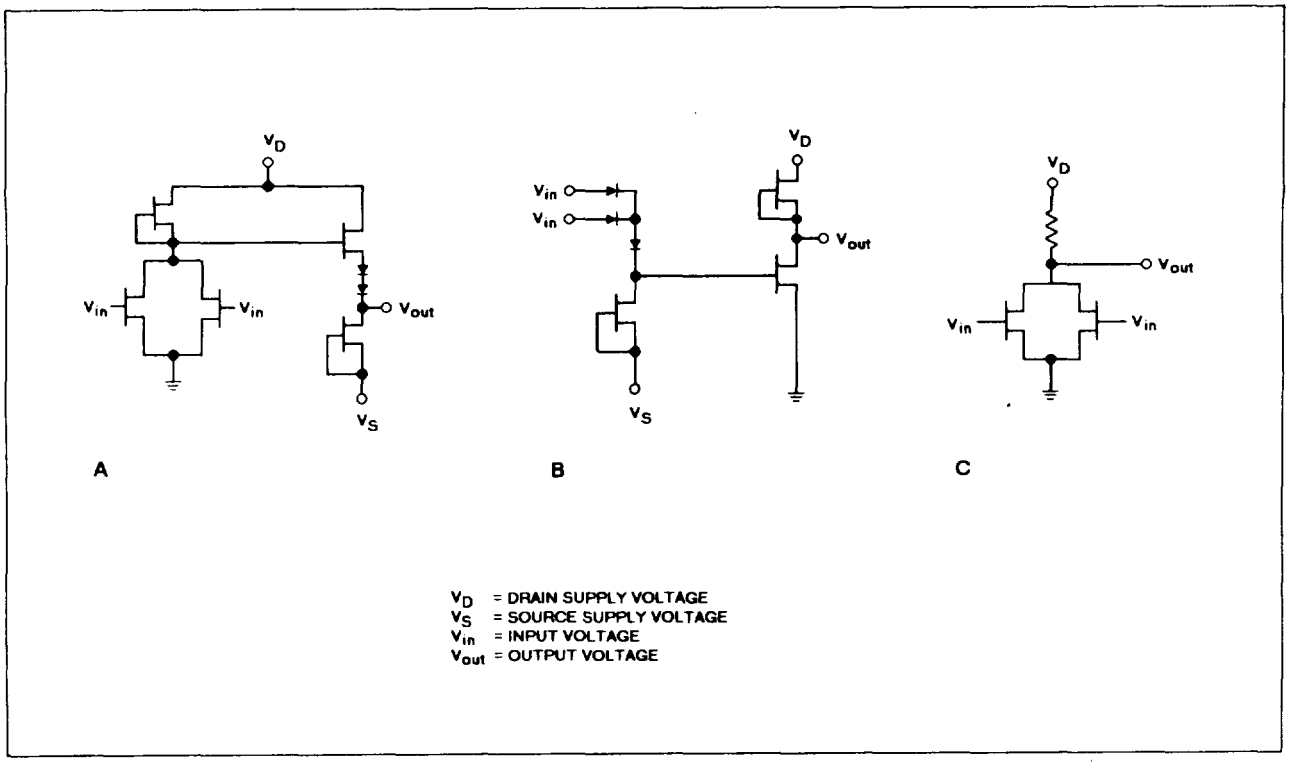


Figure 2-28.-High-speed gallium arsenide logic gates; (a) buffered FET logic (BFL) NOR gate; (b) Schottky diode FET logic (SDFL) NOR gate; (c) direct-coupled FET logic (DCFL) NOR gate.

but not for circuits with the complexity of very large-scale integration (more than 10,000 gates).

DCFL incorporating ENFETs is inherently much simpler than BFL or SDFL since there is no need for level shifting. The very low power consumption and circuit simplicity lead to high packing density (more than 5000 gates/mm^2 or 3.2 x 10^6 gates/$in.^2$) at only slightly slower speeds.

The table 2-3 lists the projected applications for each of the three logic gates, along with the competing silicon technology.

DIGITAL COUNTER

A digital counter is an instrument that, in its simplest form, provides an output that corresponds to the number of pulses applied to its input.

Counters may be categorized into two types: the Moore machine or the Mealy machine. The simpler counter type, the Moore machine, has a single count input (also called the clock input or pulse input), while the Mealy machine has additional inputs that alter the count sequence. Digital counters take many forms, such as geared mechanisms (tape counters and odometers are examples), relays (old telephone switching systems), vacuum tubes (old test equipment), and solid-state semiconductor circuits (most modern electronic counters). This section will stress solid-state electronic counters.

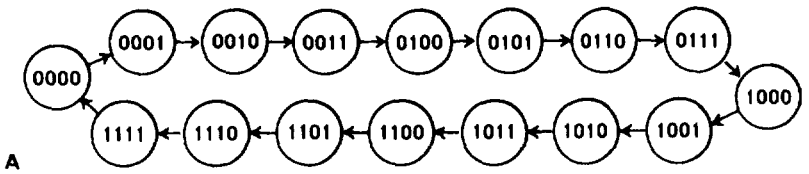
Most digital counters operate in the binary number system, since binary is easily implemented with electronic circuitry. Binary allows any integer (whole number) to be represented as a series of binary digits, or

Table 2-3.-Gallium Arsenide Logic Gate Applications and Issues

GALLIUM ARSENIDE TECHNOLOGY	APPLICATIONS	FEASIBILITY ISSUES	COMPETING SILICON TECHNOLOGY
Buffered FET logic (BFL)	SSI (small-scale integration) - MSI (medium-scale integration) superfast logic-prescalers, multiplexers, demultiplexers, fast cache memory	Most producible, uses large logic swings with good noise margin; tolerant of FET threshold variations; least area efficient.	Emitter-coupled logic (ECL) and submicrometer MOS
Schottky diode FET logic SSDFL)	High-speed LSI (large scale integration), for example, 8 x 8 multiplier; arithmetic-logic unit (ALU), gate arrays	Replaces FETs with diodes for logic function; usually smaller noise margin than BFL, but still fairly tolerant of threshold variations; circuit design is complicated by fan-out sensitivity	Bipolar LSI; 1-μm MOS LSI
Direct-coupled FET logic (DCFL)	Low power or VLSI (very large scale integration) applications, memory, gate arrays	Uses enhancement FETs; low-noise margin; requires excellent threshold control	1-μm MOS VLSI

bits, where each bit is either a 0 or 1 (off or on, low or high, and so forth).

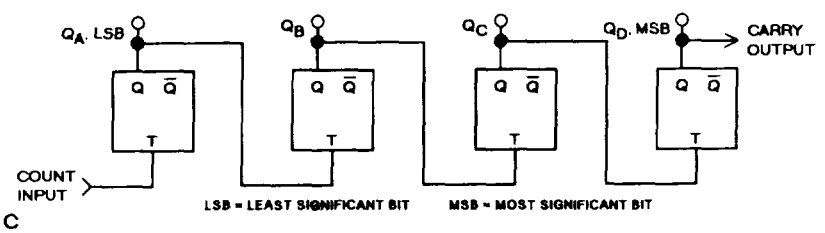
Figure 2-29 shows a 4-bit binary counter that can count from 0 to 15; the sixteenth count input causes the counter to return to the 0 output state and generate a carry pulse. This action of the counter to return to the 0 state with a carry output on every sixteenth pulse makes the 4-bit binary counter a modulo 16 counter. The four binary-digit outputs $Q_D, Q_C, Q_B,$ and Q_A are said to have an 8-4-2-1 "weighting" because, if Q_D through Q_A are all ones, then the binary counter output is $1111_2 = 1 \times 2^3 + 1 \times 2^2 + 1 \times 2^1 + 1 \times 2^0 = 8 + 4 + 2 + 1 = 15_{10}$, where the subscripts indicate the base of the number system. In figure 2-29, view A, the counter state-flow diagram is shown. Each possible state is represented by the numerical output of that state. Upon receiving a count pulse, the counter must change state by following an arrow from the present state to the next state. In figure 2-29, view B, a table is given showing the counter output after a given number of input pulses, assuming that the counter always starts from the 0 state. The counter output is listed in binary, octal, decimal, and hexadecimal. Figure 2-29, view C, shows a block diagram of the counter built with *T* flip-flops, and figure 2-29, view D, shows the counter waveforms through time, with a periodic count input. The *T* flip-flop is a device that has either a 0 or a 1 on its *Q* output at all times. When the count input *T* moves from the 1 state to the 0 state, the flip-flop output must change state, from a 0 to a 1 or a 1 to a 0. The carry output produces a 1-to-0



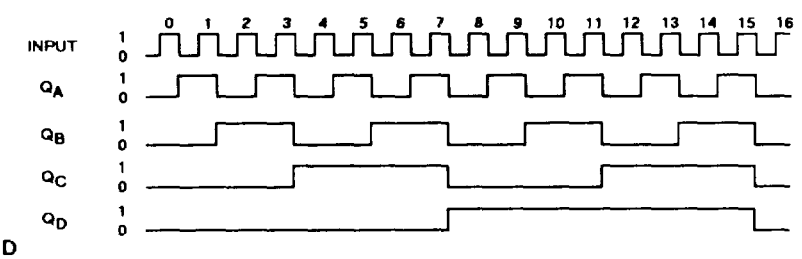
A

NUMBER OF COUNT PULSES	BINARY OUTPUT	OCTAL (BASE 8)	DECIMAL (BASE 10)	HEXADECIMAL (BASE 16)
0	0000	0	0	0
1	0001	1	1	1
2	0010	2	2	2
3	0011	3	3	3
4	0100	4	4	4
5	0101	5	5	5
6	0110	6	6	6
7	0111	7	7	7
8	1000	10	8	8
9	1001	11	9	9
10	1010	12	10	A
11	1011	13	11	B
12	1100	14	12	C
13	1101	15	13	D
14	1110	16	14	E
15	1111	17	15	F
16	0000	0	0	0

B



C



D

Figure 2-29.-Four-bit binary counter using trigger flip-flops; (a) state-flow diagram, (b) table of counter output in various number systems; (c) circuit block diagram; (d) output waveforms.

transition on every sixteenth count input, producing a divide-by- 16 function.

The 4 bits of the counter of figure 2-29 can be grouped together and used to represent a single hexadecimal digit; in figure 2-29, view B, each counter output state represents one hexadecimal digit. A 2-digit hexadecimal counter requires two sets of 4-bit binary counters, the carry output from the first set of counters driving the count input of the second set of counters.

A decimal counter built from four binary counters is shown in figure 2-30. Let 4 bits of data from the binary counter represent 1 decimal digit. The counter will work in the same way as the counter shown in figure 2-29, except that all the flip-flops are reset to the 0 state when the counter moves from the 1001_2 = 9 state, instead of advancing to the 1010_2 = 10 state. Besides the AND gate that is now used to detect the 1001 state of the counter and enable the resets, the circuit block diagram shows a new type of flip-flop. The "SR" flip-flop acts like a *T* flip-flop with an additional input that forces the *Q* output to a 1 state when the *S* (set) input is high and the *T* input has had a 1-to-0 transition applied. An *R* (reset) input acts as the S input does, except that the *Q* output goes to 0. This example decimal counter has an 8-4-2-1 weighted output that is known as binary-coded decimal (BCD). A seven-segment display is easily interfaced to the binary-coded-decimal counter using a binary-coded-decimal-to-seven- segment decoder/driver circuit that is widely available.

Digital counters are found in much modern electronic equipment, especially equipment that is digitally controlled or has digital numeric displays. A frequency counter, as a test instrument or a channel frequency display on a radio tuner, consists simply of a string of decade counters that count the pulses of an input signal for a known period of time, and display that count on a seven-segment display. A digital voltmeter operates by using nearly the same idea, except that the counter counts a known frequency for a period of time proportional to the input voltage.

Digital computers may contain counters in the form of programmable interval timers that count an integral number of clock pulses of a known period, and then generate an output at the end of the count to signal that the time period has expired. Most of the counters in a microprocessor consist of arithmetic logic units (ALU) that add one many-bit number to another, storing the results in a memory location. The program and data counters are examples of this kind of counter.

Counters have progressed from relays to light-wavelength-geometry very large-scale integrated circuits. There are several technologies for building individual digital counters. Single counters are available as integrated circuit chips in emitter-coupled logic (ECL), transistor-transistor logic (TTL), and CMOS. The three technologies are listed in the order of decreasing speed and decreasing power dissipation. ECL will operate to 600 MHz, TTL to 100 MHz

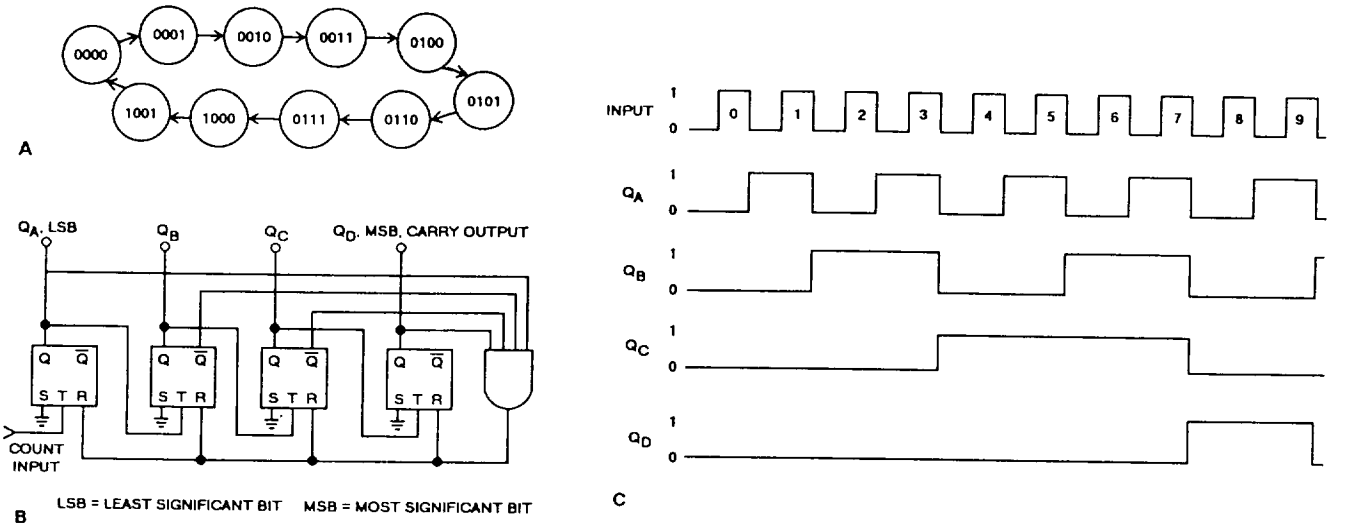


Figure 2-30.-Four-bit binary counter modified to be a decimal counter; (a) state-flow diagram; (b) circuit block diagram, (c) output waveforms.

(Schottky-clamped), and CMOS to 5 MHz. Standard, *high-volume* production *n*- channel metal oxide semiconductor (NMOS) LIS can implement a 1-bit binary counter in a 100 x 100 μm^2 (39 x 3.9 mil^2) area that will operate to 10 MHz. A gallium arsenide metal semiconductor field-effect transistor (MESFET) master-slave *JK* flip-flop has been reduced that operates at 610 MHz in a 390 x 390 μm^2 (15 x 15 mil^2) surface area while consuming the power of an NMOS.

COMPARATOR CIRCUIT

A comparator circuit is an electronic circuit that produces an output voltage or current whenever two input levels simultaneously satisfy predetermined amplitude requirements. A comparator circuit may be designed to respond to continuously varying (analog) or discrete (digital) signals, and its output may be in the form of signaling pulses that occur at the comparison point or in the form of discrete direct-current (dc) levels.

Linear Comparator

A linear comparator operates on continuous, or nondiscrete, waveforms. Most often one voltage, referred to as the reference voltage, is a variable de or level-setting voltage and the other is a time-varying waveform. One common application of the comparator is in a linear time-delay circuit. Inputs consist of a sawtooth waveform of linearly increasing magnitude (ramp function) and a variable de reference voltage. The reference voltage can be calibrated in units of time, as measured from the beginning of the sawtooth.

A clipper and a coincidence amplifier, together with a resistance-capacitance (RC) differentiating circuit, can perform the function of comparator. In figure 2-31, the series clipper, usually called a pick-off diode for this application, does not conduct until the input reaches level V_R. The diode input is a sawtooth as shown. Consequently, only the portion of the sawtooth, above V_R appears at the output of the clipper. This output is applied to the *RC* differentiating network, which passes only the initial part of the rise. This short pulse is then amplified to produce the resultant output waveform.

The particular amplifier illustrated is a two-transistor, high-gain amplifier with a relatively high input impedance and a low output impedance. A sharper pulse can be obtained if the amplifier is made regenerative. It may even take the form of a multivibrator or blocking oscillator to increase the gain at the point of coincidence.

Regenerative Comparator

Multivibrators can be used in several ways directly as comparators without need for the pick-off diodes; such comparators sense the required coincidence accurately and introduce little additional delay. A simple type is the direct-coupled bistable circuit, sometimes known as the Schmitt circuit, as shown in figure 2-32. This example employs enhancement mode *p*- channel field-effect transistors and can be made to function from either negative or positive-going input waveforms. The example compares a negative-going input waveform with reference voltage V_R.

Under a variety of choices of supply voltage and resistances, the circuit will be bistable; that is, either of the two transistors can be conducting for a particular voltage at input gate G_1. Until a predetermined value of

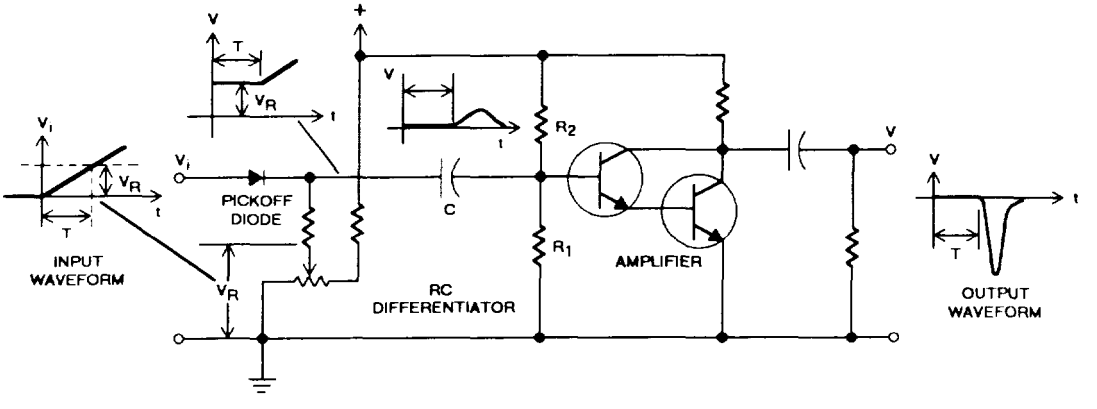


Figure 2-31.-Simple comparator circuit using pick-off diode.

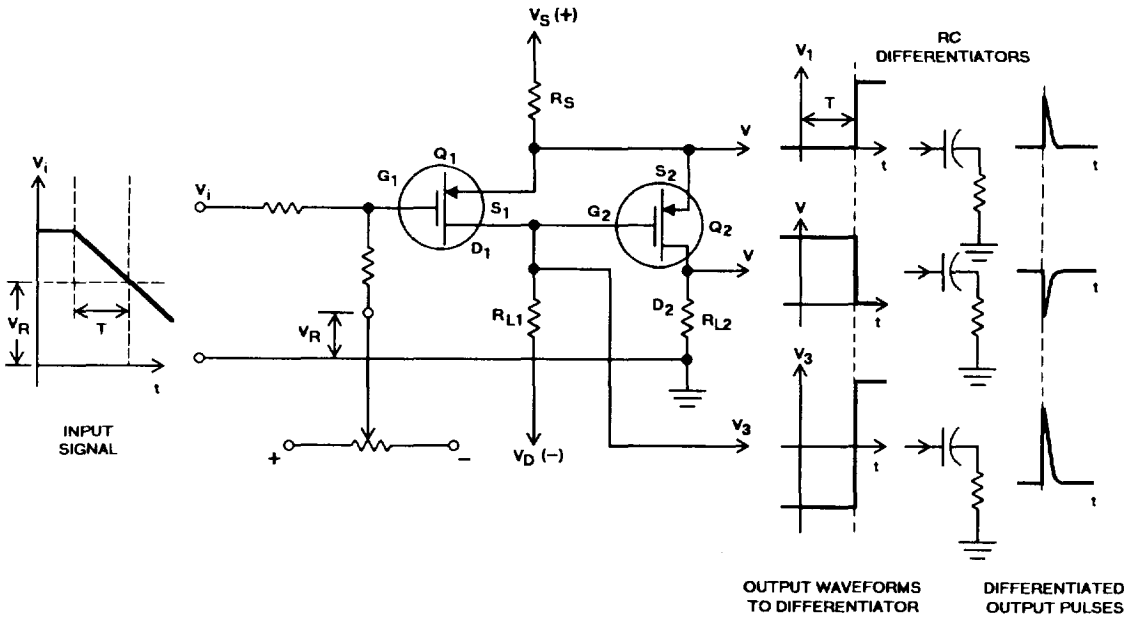


Figure 2-32.-Comparator using field-effect transistors in source-coupled bistable circuit.

reference voltage is reached, Q_1 is nonconducting, and at time *T*, it switches from nonconducting to conducting while Q_2 simultaneously switches from conducting to nonconducting. With dc coupling, as shown in figure 2-32, three outputs of differing dc levels and polarities are produced. If *RC* differentiating circuits are added as indicated, sharp pulses can then be obtained. When the input waveform ends, all points in the circuit return to their initial states.

Direct-coupled regenerative comparators such as the Schmitt circuit are usually bidirectional, responding to inputs approaching the reference level V_R from either the positive or the negative side. If the input starts at a value lower than V_R, the output voltage V_1 will be at its high value until V_R is reached, and it then shifts to its low value. Polarities of the other output signals will be correspondingly reversed. Thus, at the voltage coincidence of V_i and V_R, there will be generated one of two possible output states definable as logic level (1) or logic level (0) in digital terminology. Because of design limitations in practical circuits, the input voltage at which the bistable circuit changes state is slightly less or greater than V_R, depending upon whether the input signal is positive-going or negative-going. This slight difference in level is referred to as the hystersis of the circuit.

Integrated Circuit Comparators

High-gain dc operational amplifiers operated in the nonfeedback mode are often used to perform the comparator function, and many such amplifiers are classified as comparators because they are specifically designed to meet the needs for accurate voltage comparison applications. Such "op-amps" have two inputs, the output being inverting with respect to one and noninverting with respect to the other, as shown in figure 2-33. The voltage gain (amplification) of the amplifier is so high that its output will swing through its entire dynamic range, V_{min} to V_{max}, for very small changes in input voltage. Thus, for $V_i \langle V_R$, the amplifier will be cut off and the output voltage will be at V_{max}, and for $V_i \rangle V_R$, the amplifier will saturate and the output will be at V_{min}. For digital system applications, the output

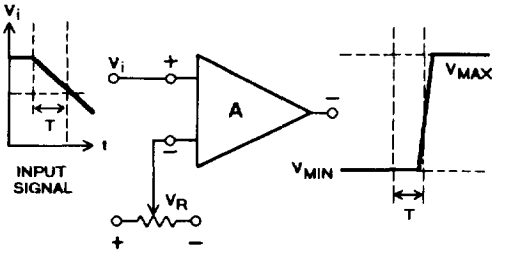


Figure 2-33.-Comparator circuit using integrated operational amplifier.

levels may be designed to coincide with logic level (0) and logic level (1) of the specific digital system, and thus be suitable for converting a specific level in a continuously varying signal to a specific logic number assigned to the level. Arrays of such comparators connected to a common input, each designed to respond at a distinct reference voltage, and with the outputs connected to appropriate logic gates may be used to convert a range of signal levels to a specific digital code, and as such form the basic building block of analog-to-digital converters.

The voltage gain, and hence the timing precision of the operational amplifier comparator, can be increased by converting it to a regenerative comparator, as shown in figure 2-34. It then becomes an integrated circuit form of the Schmitt circuit shown in figure 2-3.

Digital Comparator

The term *digital comparator* has historically been used when the comparator circuit is specifically designed to respond to a combination of discrete level (digital) signals; for example, when one or more such input signals simultaneously reach the reference level that causes the change of state of the output. Among other applications, such comparators perform the function of the logic gate such as the AND, OR, NOR, and NAND functions. More often, however, the term *digital comparator* is used to describe an array of logic gates designed specifically to determine whether one binary number is less than or greater than another binary number. Such digital comparators are sometimes called magnitude comparators or binary comparators.

Comparators may take many forms and can find many uses in addition to those that have been discussed. For example, the electronically regulated dc voltage supply uses a circuit that compares the de output voltage with a fixed reference level. The resulting difference

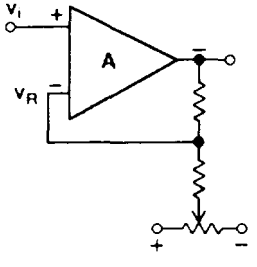


Figure 2-34.-Regenerative integrated circuit comparator.

signal controls an amplifier, which, in turn, changes the output to the desired level. In a radio receiver, the automatic gain control circuit may bethought of broadly as a comparator; it measures the short-term average of the signal at the output of the detector, compares this output with a desired bias level on the radio-frequency amplifier stages, and changes that bias to maintain a constant average level output from the detector.

ANALOG-TO-DIGITAL CONVERTER

The analog-to-digital converter is a device for converting the information contained in the value or magnitude of some characteristics of an input signal, compared to a standard or reference. This input is compared to information in the form of discrete states of a signal, usually with numerical values assigned to the various combinations of discrete states of the signal.

Analog-to-digital (A/D) converters are used to transform analog information, such as audio signals or measurements of physical variables (for example, temperature, force, or shaft rotation) into a form suitable for digital handling, which might involve any of the following operations: (1) processing by a computer or by logic circuits, including arithmetical operations, comparison, sorting, ordering, and code conversion; (2) storage until ready for further handling; (3) display in numerical or graphical form; and (4) transmission.

If a wide-range analog signal can be converted, with adequate frequency, to an appropriate number of two-level digits, or bits, the digital representation of the signal can be transmitted through a noisy medium without relative degradation of the fine structure of the original signal.

Conversion involves quantizing and encoding. Quantizing means partitioning the analog signal range into a number of discrete quanta and determining to which quantum the input signal belongs. Encoding means assigning a unique digital code to each quantum and determining the code that corresponds to the input signal. The most common system is binary, in which there are 2^n quanta (where n is some whole number), numbered consecutively; the code is a set of n physical two-valued levels or bits (1 or 0) corresponding to the binary number associated with the signal quantum.

Figure 2-35 shows a typical 3-bit binary representation of a range of input signals, partitioned into eight quanta. For example, a signal in the vicinity of 3/8 full scale (between 5/16 and 7/16 will be coded 011 (binary 3).

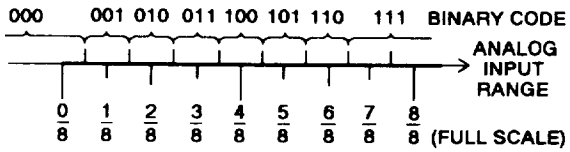


Figure 2-35.-A 3-bit binary representation of a range of input signals.

Conceptually, the conversion can be made to take place in any kind of medium—electrical, mechanical, fluid, optical, and so on (for example, shaft-rotation-to-optical). By far the most commonly employed form of A/D converters comprises those devices that convert electrical voltages or currents to coded sets of binary electrical levels (for example, +5 V or 0 V) in simultaneous (parallel) or pulse-train (serial) form, as shown in figure 2-36. The serial output is not always made available.

The converter depicted in figure 2-36 converts the analog input to a five-digit "word." If the coding is binary, the first digit (most significant bit, abbreviated MSB) has a weight of 1/2 full scale, the second 1/4 full scale, and so on, down to the *n* th digit (least-significant bit, abbreviated LSB), which has a weight of 2^{-n} of full scale (1/32 in this example). Thus, for the output word shown, the analog input must be given approximately by the following equation. The

$$\frac{16}{32} + \frac{0}{32} + \frac{4}{32} + \frac{2}{32} + \frac{0}{32} = \frac{22}{32} = \frac{11}{16}\ FS\ (fullscale)$$

number of bits, *n*, characterizes the resolution of a converter.

Figure 2-36 also shows a commonly used configuration of connections to an A/D converter: the analog signal and reference inputs; the parallel and serial digital outputs; the leads from the power supply, which provides the required energy for operation; and two control leads–a start-conversion input and a status-indicating output (busy) that indicates when a conversion is in progress. The reference voltage or current is often developed within the converter.

Second in importance to the binary code and its many variations is the binary-coded decimal (BCD), which is used rather widely, especially when the encoded material is to be displayed in numerical form. In BCD, each digit of a radix-10 number is represented by a 4-digit binary subgroup. For example, the BCD code for 379 is 0011 0111 1001. The output of the A/D converter used in digital panel meters is usually BCD.

There are many techniques used for A/D conversion, ranging from simple voltage-level comparators to sophisticated closed-loop systems, depending on the input level, output format, control features, and the desired speed, resolution, and accuracy. The two most popular techniques are dual-slope conversion and successive-approximations conversion.

Dual-slope converters have high resolution and low noise sensitivity; they operate at relatively low-speeds-usually a few conversions per second. They are primarily used for direct dc measurements requiring digital readout; the technique is the basis of the most

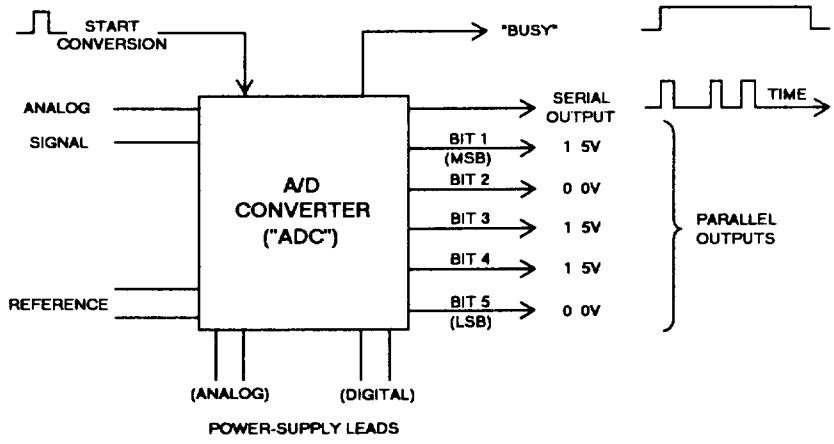


Figure 2-36.-Analog-to-digital converter, showing parallel and serial (return-to-zero) output formats for code 10110.

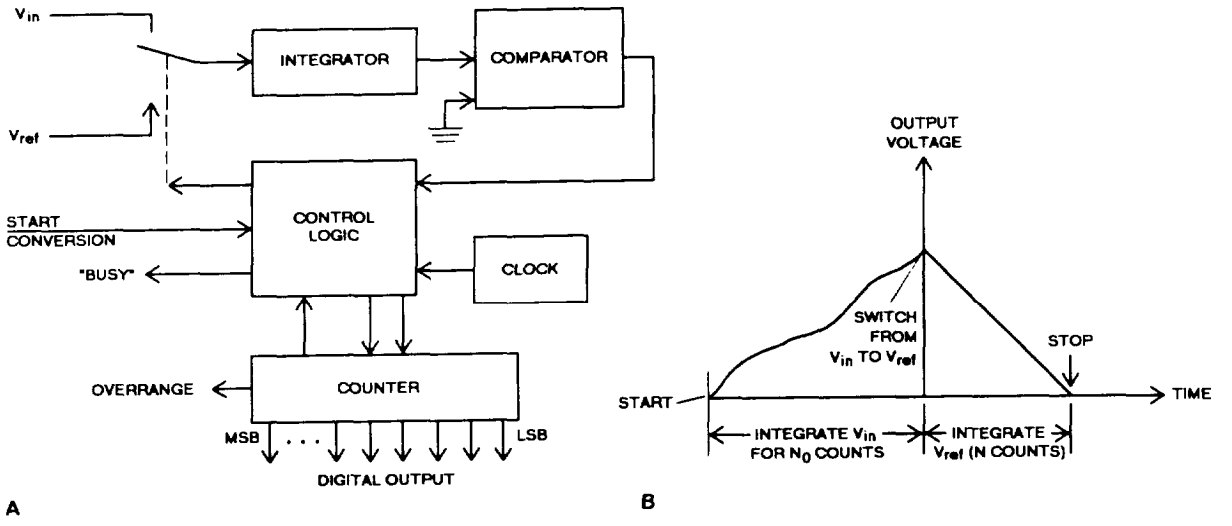


Figure 2-37.-Example of dual-slope conversion; (a) block diagram of converter; (b) integrator output.

widely used approach to the design of digital panel meters.

Figure 2-37, view A, is a simplified block diagram of a dual-slope converter. The input is integrated for a period of time determined by a clock-pulse generator and counter (fig. 2-37, view B). The final value of the signal integral becomes the initial condition for integration of the reference in the opposite sense, while the clock output is counted. When the net integral is zero, the count stops. Since the integral "up" of the input over a fixed time (N_0 counts) is equal to the integral "down" of the fixed reference, the ratio of the number of counts of the variable period to that of the fixed period is equal to the ratio of the average value of the signal to the reference.

Successive-approximations conversion is a high-speed technique used principally in data-acquisition and computer-interface systems. Figure 2-38, view A, is a simplified block diagram of a successive-approximations converter. In a manner

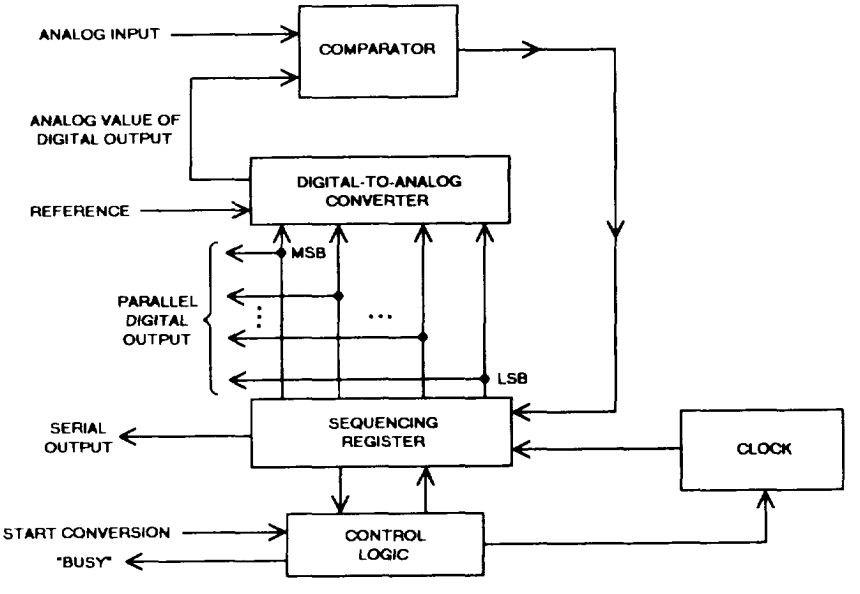


Figure 2-38.-Block diagram of a successive-approximations converter.

analogous to the operation of an apothecary's scale with a set of binary weights, the input is "weighed' against a set of successively smaller fractions of the reference, produced by a digital-to-analog (D/A) converter that reflects the number in the output register.

First, the MSB is tried (1/2 full scale). If the signal is less than the MSB, the MSB code is returned to zero; if the signal is equal to or greater than the MSB, the MSB code is latched in the output register (fig. 2-38, view B). The second bit is tried (1/4 full scale). If the signal is less than 1/4 or 3/4, depending on the previous choice, bit 2 is set to zero; if the signal is equal to or greater than 1/4 or 3/4; bit 2 is retained in the output register. The third bit is tried (1/8 full scale). If the signal is less than 1/8, 3/8, 5/8, or 7/8, depending on previous choices, bit 2 is set to zero; otherwise, it is accepted. The trial continues until the contribution of the least-significant bit has been weighed and either accepted or rejected The conversion is then complete. The digital code latched in the output register is the digital equivalent of the analog input signal.

The earliest A/D converters were large rack-panel chassis-type modules using vacuum tubes, requiring about 1.4 ft^3 (1/25 m^3) of space and many watts of power. Since then, they have become smaller in size and cost, evolving through circuit-board, encapsulated-module, and hybrid construction, with improved speed and resolution. Single-chip, 12-bit A/D converters with the ability to interface with microprocessors are now available in small integrated-circuit packages. Integrated-circuit A/D converters, with 6-bit and better resolution and conversion rates to beyond 50 MHz, are also available.

DIGITAL-TO-ANALOG CONVERTER

A digital-to-analog converter is a device for converting information in the form of combinations of discrete states or a signal, often representing binary number values, to information in the form of the value or magnitude of some characteristics of a signal, in relation to a standard or reference. Most often, it is a device that has electrical inputs representing a parallel binary number, and an output in the form of voltage or current.

Figure 2-39 shows the structure of a typical digital-to-analog converter. The essential elements, found even in the simplest devices, are enclosed within the dashed rectangle. The digital inputs, labeled u_i, $i = 1, 2, \ldots, n$, are equal to 1 or 0. The output voltage E_o is given by the following equation, where V_{REF} is an analog reference voltage and K is a constant.

$$E_o = KV_{REF}(u_1 2^{-1} + u_2 2^{-2} + u_3 2^{-3} + \ldots + u_n 2^{-n})$$

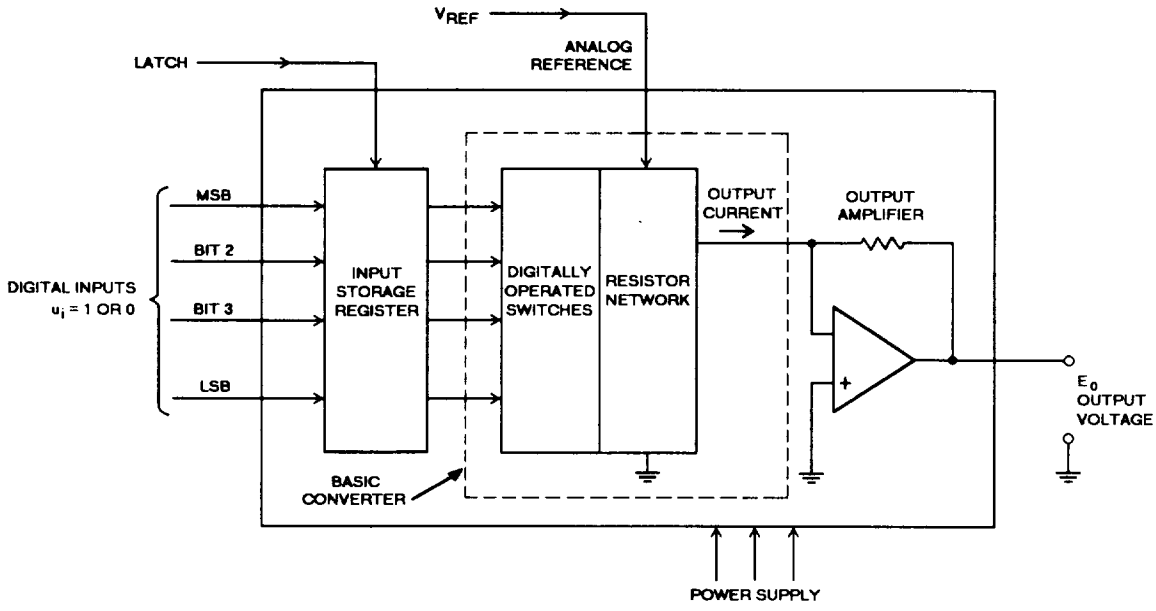


Figure 2-39.-Typical digital-to-analog converter.

Thus, for a 5-bit binary converter with latched input code 10110, the output is given by the following equation.

$$E_o = \left(\frac{16}{32} + \frac{0}{32} + \frac{4}{32} + \frac{2}{32} + \frac{0}{32}\right)KV_{REF} = \frac{11}{16}KV_{REF}$$

Bit 1 is the "most significant bit" (MSB), with a weight of 1/2; bit n is the "least significant bit" (LSB), with a weight of 2^{-n}. The number of bits n characterizes the resolution.

Digital-to-analog (D/A) converters (sometimes called DACs) are used to present the result of digital computation, storage, or transmission, typically for graphical display or for the control of devices that operate with continuously varying quantities. D/A converter circuits are also used in the design of analog-to-digital converters that employ feedback techniques, such as successive-approximation and counter-comparator types. In such applications, the D/A converter may not necessarily appear as a separately identifiable entity.

The fundamental circuit of most D/A converters involves a voltage or current reference; a resistive "ladder network" that derives weighted currents or voltages, usually as discrete fractions of the reference; and a set of switches, operated by the digital input, that determines which currents or voltages will be summed to constitute the output.

An elementary 3-bit D/A converter is shown in figure 2-40. Binary-weighted currents developed in R_1, R_2, and R_3 by V_{REF} are switched either directly to ground or to the output summing bus (which is held at zero volts by the operational-amplifier circuit). The sum of the currents develops an output voltage of polarity opposite to that of the reference across the feedback resistor R_f. Table 2-4 shows the binary relationship between the input code and the output, both as a voltage and as a fraction of the reference.

The output of the D/A converter is proportional to the product of the digital input value and the reference. In many applications, the reference is fixed, and the output bears a fixed proportion to the digital input. In other applications, the reference, as well as the digital input, can vary; a D/A converter that is used in these applications is thus called a multiplying DAC. It is principally used for imparting a digitally controlled scale factor, or "gain," to an analog input signal applied at the reference terminal.

Except for the highest resolutions (beyond 16 bits), commercially available D/A converters are generally manufactured in the form of dual in-line-packaged integrated circuits, using bipolar, MOS, and hybrid technologies. A single chip may include just the resistor network and switches, or it may also include a reference circuit, output amplifier, and one or more sets of registers (with control logic suitable for direct microprocessor interfacing).

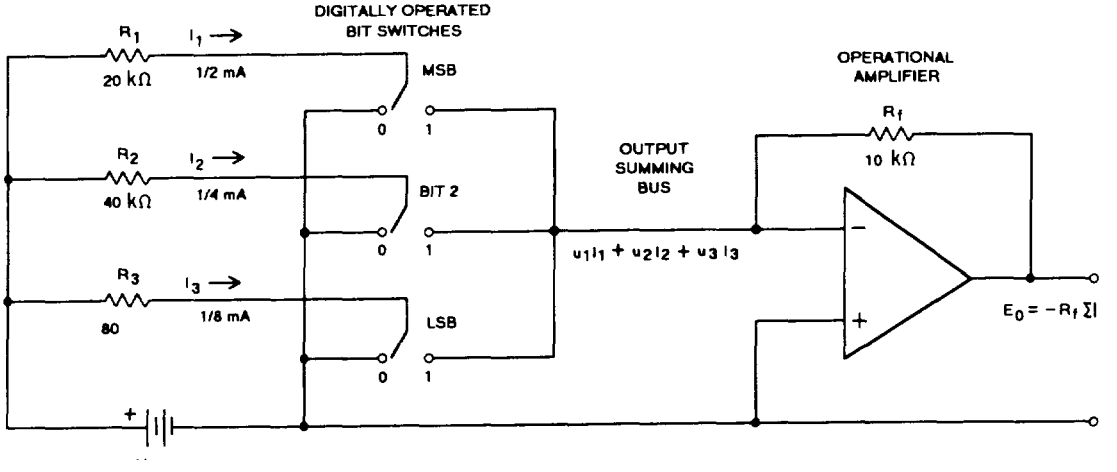


Figure 2-40.-Elementary 3-bit digital-to-analog converter.

Table 2-4.-Input and Output of Converter

DIGITAL INPUT CODE			ANALOG OUTPUT	
u_1	u_2	u_3	$-E_o$ volts	$\frac{E_o}{V_{REF}}$
0	0	0	0	0
0	0	1	1.25	1/8
0	1	0	2.5	2/8
0	1	1	3.75	3/8
1	0	0	5.0	4/8
1	0	1	6.25	5/8
1	1	0	7.5	6/8
1	1	1	8.75	7/8

CHAPTER 3

COMMUNICATIONS

Chapter Objective: *Upon completion of this chapter, you will have a working knowledge of communications procedures, to include receiver testing, transmitter testing, fiber optics, and basic intermediate-level (I level) equipment maintenance.*

As an Aviation Electronics Technician, you will be tasked to operate and maintain many different types of aircraft communications equipment. This equipment will differ in many respects, but in other respects, it will be much the same. As an example, there are numerous models of AM radios, yet they all serve the same function and operate on the same basic principles. It is beyond the scope of this chapter to discuss all radios or to present information that relates to all of the many different pieces of communications equipment. Therefore, only representative communication receivers and transmitters will be discussed, along with a brief overview of fiber optics. You will also be given a basic understanding of equipment repair at the I level of maintenance.

COMMUNICATION RECEIVER TESTING

Learning Objective: *Recognize the testing procedures for communication receivers, to include receiver sensitivity, bandwidth response and measurements, squelch, modulation and AFC measurements, and typical receiver alignment.*

Communication receivers are generally composed of a series of selective RF and AF circuits, each stage of which is designed to amplify the output of the preceding stage. The lowered efficiency of any amplifier, or a change in any one circuit parameter, usually results in lowered overall efficiency of the receiver. The sensitivity of the receiver may also be decreased by the misalignment of the successive circuits, although each of which may function in a suitable manner as a unit. The sole function of a communication receiver is to receive (selectively) a weak signal. Therefore, an objective overall test of sensitivity is the most significant single check that can be made on the condition of a receiver. Some receivers include a built-in output meter. Others require an external indicator, such as a dB meter or a spectrum analyzer to facilitate testing. The only other equipment needed for a sensitivity check is a calibrated signal for the excitation of the receiver on its various bands. Any decrease in sensitivity must be corrected.

In addition to sensitivity checks, qualitative checks, as outlined in the maintenance instruction manual or the equipment's technical manual, must be performed. Adjustment and servicing methods for a specific receiver are discussed in detail in its associated technical manual. When attempting to isolate receiver faults, first test the most accessible (or vulnerable) parts. Since a receiver could operate for years with reduced sensitivity before trouble was detected or a complete failure occurred, performing the maintenance prescribed in the equipment's technical manual is a necessity.

RECEIVER SENSITIVITY

Sensitivity measurements provide convenient and quantitative information regarding a receiver's ability to detect small (weak) signals in the presence of electronic noise. The sensitivity of a radio receiver is defined as the input carrier voltage with standard modulation required to develop a standard value of test output.

AM Receiver Sensitivity

Noise-figure measurements are a ratio of the signal-to-noise power ratio of an ideal receiver to the signal-to-noise power ratio of the receiver under test, Sensitivity measurements, however, are relative measurements that are arbitrarily calculated to be the 30-percent, 1,000-Hz modulated signal input required to raise the detected audio 10 dB or greater above the receiver's noise level. Sensitivity is measured in microvolts, or in dB below 1 volt. This arbitrary reference value was selected because a 10 dB change represents a x 10 change in voltage. Therefore, the detected output, as measured across a given resistance with zero signal input to the receiver, can be increased by a factor equal to or greater than 10. This increase can be achieved by increasing the input modulated signal level (in microvolt) to a predetermined level. Note that sensitivity (and selectivity) may be affected by alignment in all types of receivers.

As receivers become more complex, alignment becomes more of a problem. In amplitude-modulation receivers, improper alignment may result in the loss of weak signals through loss of sensitivity, and through inability to select the desired signal. If the oscillator is shifted off frequency, dial error will be introduced. Tracking error produces a varying intermediate frequency, which results in a loss of signal over portions of the frequency range of the receiver. In frequency-modulation receivers, the discriminator tuning becomes somewhat critical. In phase-modulated receivers, phasing of the carrier must be correct, adding to the alignment problem. When automatic gain control and automatic frequency control are added to a receiver, proper alignment procedures must be followed. In equipment employing crystals, as either reference generators, oscillators, or filters, the alignment must center on the crystals since, for all practical purposes, the frequencies of crystals are not variable. As a result, because of the wide variations in circuitry between models of receivers, the actual alignment procedures and specifications provided in the applicable technical manual must be closely followed if the sensitivity check indicates a need for alignment.

Single Sideband Sensitivity Measurement Considerations

Sensitivity measurements for single sideband (SSB) receivers are determined in a manner similar to that used for other amplitude-modulation equipment. However, the following considerations must be taken into account when performing measurements.

FM (F-3) SENSITIVITY MEASUREMENT.— The procedures for measurement of FM (F-3) receiver sensitivity are analogous to those for AM receivers; however, an FM signal generator must be used. The modulation signal is a 1,000-Hz tone with 2,500-Hz deviation. The modulation-on/modulation-off reference is still used as in AM, and the minimum signal required to obtain a 10 dB drop is a measure of the receiver's sensitivity.

PULSE-MODULATION SENSITIVITY MEASUREMENT.— Continuous-wave generator methods of measuring sensitivity do not provide an accurate indication of the ability of a receiver that is designed for the reception of pulse-modulated signals to receive weak pulse transmissions. A better method of determining the sensitivity of a pulse-modulation receiver involves performing a minimum-discernible signal measurement. This type of measurement consists of measuring the power level of a pulse whose level is just sufficient to produce a visible receiver output. Because of the relatively wide bandwidths associated with pulse-modulation receivers, a still better performance indication can be obtained. This is done by determining quantitatively how much noise is inherent in the receiver, since noise is the limiting factor in the determination of maximum sensitivity. This method of checking sensitivity uses a noise generator for a signal source. The noise in the receiver is related to a calculable noise figure.

DETERMINING IF BANDWIDTH RESPONSE

A graph showing the bandwidth response of an IF amplifier can be constructed by plotting frequency horizontally (from left to right) and signal amplitude vertically. This method would be ideal for record retention purposes, but it is not necessary for receiver checks and adjustments. For such checks and adjustments, a spectrum analyzer is used in conjunction with a tracking

generator. Figure 3-1 illustrates such an arrangement. In figure 3-1, the voltage control oscillator feeds both the spectrum analyzer's mixer and the tracking generator's mixer. Because of this simultaneous precision tracking, the tracking generator's output frequency acquires the same scan capabilities as the spectrum analyzer. Therefore, the analyzer's calibrated scan widths, which range from broadband to extremely narrow, are acquired by the tracking generator and can be applied to various IF strips. The configuration in figure 3-1 makes identification of any point on the display easy and unambiguous. The 3 dB point, 60 dB point, center frequency, or any point on the display can be measured by stopping the scan (either electronically or manually) at the point of interest and reading the indication on the tracking generator.

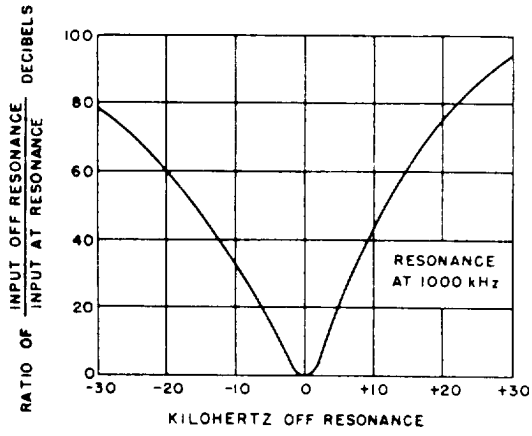


Figure 3-2.-Selectivity curve of typical AM receiver.

SELECTIVITY AND BANDWIDTH MEASUREMENTS

Selectivity is the property that enables a receiver to discriminate against transmissions other than the one to which it is tuned. It is usually expressed in the form of a curve obtained from a plot of the strength of a standard modulated-carrier signal required to produce a constant (standard) output, versus off-resonance frequency. Figure 3-2 shows a typical selectivity curve with the carrier signal strength at resonance used as a reference.

The bandwidth of a receiver is usually employed to define that portion of the selectivity curve that represents the frequency range over which the amplification is relatively constant. For most receivers, the bandwidth represents the usable portion of the curve, and has a direct relation to the fidelity of the modulated intelligence. Practically, the bandwidth is measured at the half-power down (3 dB down) or, for certain applications, at the 60 dB down points. This is represented by the frequency range between the two points on a response curve expressed as relative response in dB versus frequency, as shown in

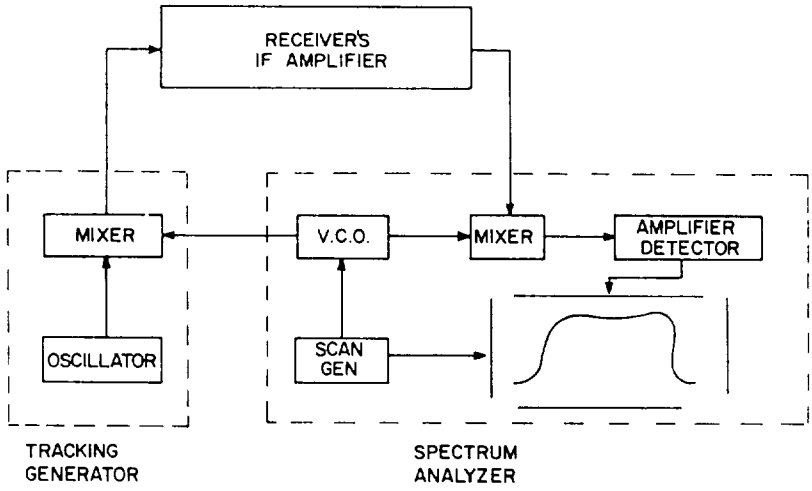


Figure 3-1.-Equipment arranged to obtain visual IF bandwidth response.

figure 3-3. The bandwidth at the 3 dB (or often the 6 dB) points, when compared with the bandwidth at the 6 dB down point, gives a good indication of the selectivity of the receiver. The character of the skirts of the curve becomes apparent. This comparison is referred to as the bandwidth or selectivity ratio. In most receivers, the overall bandwidth is determined by the IF amplifiers. Therefore, bandwidth is sometimes considered as fundamentally an IF characteristic measurement.

Overall Selectivity

Since the RF stages of a receiver are also of some importance in determining the selectivity, and are of fundamental importance in determining the image rejection characteristics, the selectivity factor is most often plotted as overall selectivity. The term *overall selectivity* usually refers to the frequency selectivity of a receiver as measured from (and including) the antenna to the input terminals of the final detector. It does not normally include any elements of the audio system. The overall selectivity of a superheterodyne receiver may be difficult to measure accurately with the equipment available in most operating installations, especially at frequencies above 1 MHz. If the lowest signal frequency is at least several times that of the lowest intermediate frequency used in the receiver, the overall selectivity is very likely to be practically the same as the lowest IF selectivity. Therefore, the lowest IF selectivity curve may suffice, and it is much easier to obtain.

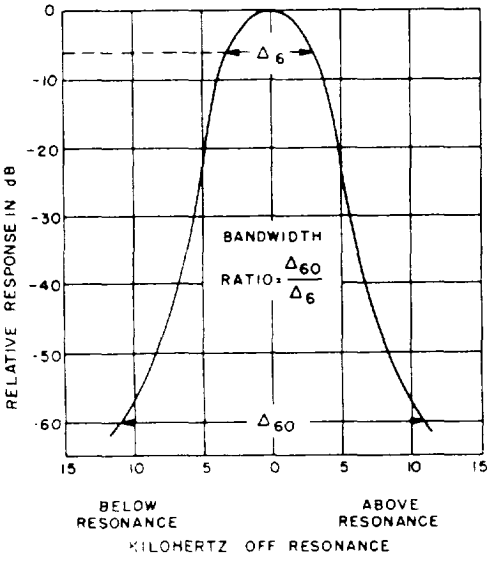


Figure 3-3.-Receiver response curve.

Bandwidth

When making bandwidth measurements, the receiver's automatic gain control (AGC) should be disabled (grounded), connected to a source of fixed bias or turned off, and the volume control set to maximum. Bandwidth curves can be obtained with the test setup illustrated in figure 3-1. This procedure can be used for narrow or wide-band receivers employing any type of demodulation. When making IF bandwidth measurements, the spectrum analyzer is set to the IFs center frequency. The scan width and scan speed controls are then adjusted to achieve an undistorted display on the CRT. If the scan time is too short with respect to the scan width of the spectrum analyzer, the response curve will appear wider than it actually is and the amplitude will be greatly reduced. This condition is illustrated in figure 3-4.

AGC Measurements

An automatic gain control (AGC) circuit reduces the effect of signal strength fading by maintaining a constant carrier level at the detector input of an AM receiver, despite variations of the in-signal carrier level. To determine the effectiveness of the AGC circuit, its characteristic should be measured at the center frequency of each band covered by the receiver. A curve can be plotted to compare the change of the receiver output to input signal levels.

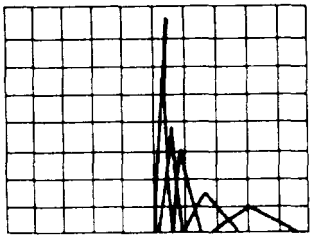


Figure 3-4.-Effects of decreased scan speed with constant scan width.

The standard method of measuring the AGC of amplitude-modulated receiver is to set the signal generator for 30 percent, 1,000-Hz modulation at carrier frequency. The receiver gain is set to maximum, and the carrier level is then varied over a wide range, such as .4 UV to 50 UV. The relative output is then plotted in dB as a function of carrier input voltage, which is also presented in terms of dB. Either the output meter incorporated in the receiver or an external indicating device, such as a vacuum tube voltmeter (VTVM) or a spectrum analyzer, can be employed to obtain the output measurement. Figure 3-5 illustrates a typical plot of carrier signal versus power output with different percentages of modulation used. In single sideband (SSB) receivers, the procedure transmission is similar except that the carrier is unmodulated.

DELAYED AGC CONSIDERATIONS.— Delayed AGC circuits are often incorporated in a receiver because even the weakest signal received in conventional AGC circuits tends to reduce the gain of the receiver somewhat. The delayed AGC adaptation incorporates a separate diode (AGC diode) in addition to the detector diode. Part of the signal fed to the detector diode is coupled to the AGC diode by a small capacitor. The AGC diode is maintained at a suitable bias; this bias keeps the diode until the peak voltage of the amplified signal voltage equals the bias introduced to the diode. For very weak signals that do not produce enough voltage on the anode of the AGC diode to overcome the existing negative potential, no AGC voltage is developed. Thus, the sensitivity of the receiver remains constant, just as if the automatic gain control were not being used. When normal strength signals are being received, which do not need the maximum sensitivity of the receiver, enough signal voltage will be coupled to the AGC diode to overcome the bias applied. AGC voltage will thus be developed normally for these stronger signals. Measurements on delayed AGC circuits are made in the same manner as described for conventional AGC circuits; however, particular attention should be given at the low-input portions of the curve.

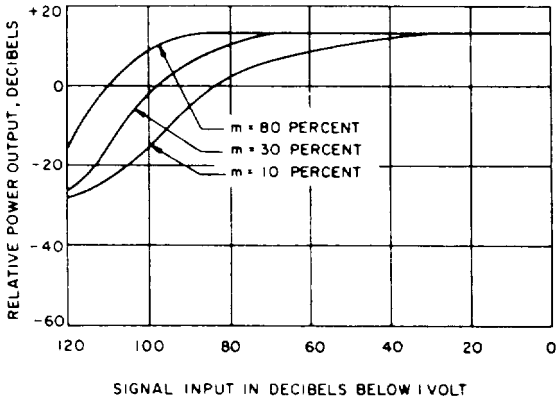


Figure 3-5.-Automatic gain control, characteristic curve.

RECEIVER STANDARD MEASUREMENTS.— In the measurement of single sideband reception, it is imperative that the standard oscillator used be precise. There are two methods of determining the precision of the frequency standard. One method involves measuring the output of the standard against a frequency counter. The frequency counter must therefore be more accurate than the standard to be measured. The other method, used in most SSB receivers, compares an extended primary standard with the internal receiver standard. The two signals are fed to a difference network, whose output is fed to either a meter with zero center swing or to a light that blinks on and off. In both instances the receiver's standard is compared against the primary standard input or a minimum zero change. Any adjustments are made only after the receiver has had time to warm up thoroughly (approximately 3 days). The adjustments are done in very small increments.

SQUELCH (SILENCER) CIRCUIT MEASUREMENTS

FM and high-frequency receiver circuits inherently have a high noise level when no signal is being received. During communications, where a receiver is tuned to a specific frequency for long stand-by periods in anticipation of signals that may appear at any time, the continuous roar of noise is annoying to anyone in the vicinity of the receiver. To silence the audio output during these periods, a squelch (or silencer) circuit has been incorporated. This circuit eliminates unwanted signal noise and other disturbances. Squelch circuits block the input to the audio stage of the receiver whenever the signal voltage is very low or is entirely absent at the detector. The squelch circuit accomplishes this silencing effect by applying a very large cutoff bias to the first audio amplifier, by actuating a relay to open the audio line, or by gating open the audio line with a field-effect transistor (FET), as shown in

figure 3-6. The high pass filter (fig. 3-6) removes all low-frequency signal components and passes the high-frequency noise components. The high-frequency noise increases with a decrease in signal strength, thus providing a gate control signal to the audio output gate FET, cutting it off in low to no signal condition.

When cutoff bias squelch is required, it must be in excess of cutoff to prevent the noise output from the intermediate amplifiers from causing current to flow in the first audio amplifier stage, even momentarily, on the noise peaks. For the determination of the squelch characteristic, connect the test equipment to the receiver, as shown in figure 3-7. The signal generator should be set to a frequency with 1,000-Hz, 30-percent modulation. With the signal generator RF output control set for zero output, the receiver output should be noted. It should be essentially zero. Gradually, the signal generator RF output should be increased until the squelch circuit operates. Operation of the squelch circuit is indicated by a sudden increase in the radio receiver output. The signal generator RF output required for the operation of the squelch circuit may be recorded as representing the squelch characteristic.

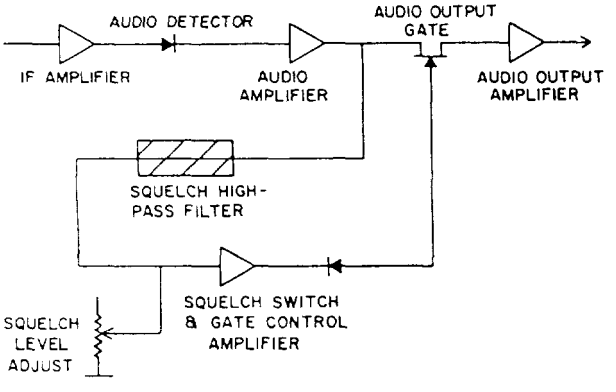


Figure 3-6.-Squelch circuit employing FET.

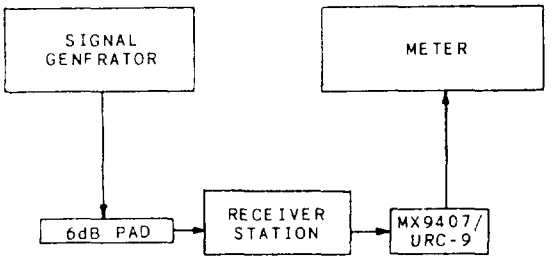


Figure 3-7.-Typical equipment arrangement for radio receiver testing.

MODULATION DISTORTION MEASUREMENTS

The distortion produced in the radio-frequency, intermediate-frequency, and detector stages of a receiver can increase significantly as a result of increases in the percentage of modulation. For example, the distortion generated by a square-law detector becomes prohibitive at percentages of modulation greater than about 50 percent. For this reason such detectors are rarely employed in communications receivers.

One method for determining the distortion of a receiver in terms of the percentage of modulation is to connect a signal generator to the receiver input, and to shunt a suitable resistor across the receiver output. The distortion meter should be connected across the resistor. The distortion meter will not respond to the resonant frequency, which is suppressed, but will provide the rms value of the other components of the distorted output signal. The meter will provide an indication of the amounts of distortion for calibrated percentages or modulation. The signal generator should be modulated at 1,000 Hz and set for an output of 50 microvolts. The receiver volume control should be adjusted for a low-level output of 50 milliwatts. This level should be maintained throughout the test. Maintaining this low power output level keeps the distortion contributed by the audio section to a low, constant level.

The percentage of modulation at the generator is then increased in convenient steps from 10 to 100 percent, and the results are plotted on linear graph paper, with the modulation percentage appearing horizontally and the value of distortion vertically. This test should then be repeated for different RF gain settings to determine whether the RF and the IF amplifiers affect the modulation.

AUTOMATIC FREQUENCY CONTROL CHARACTERISTIC MEASUREMENTS

Automatic frequency control (AFC) circuits are most often found in frequency-modulation receivers and in very-high-frequency and ultra-high frequency receivers because of the high degree of oscillator frequency stability required. Thus, FM receivers incorporate discriminator circuits, whose output voltage and polarity are

contingent upon the direction and deviation from a center, or mean value. For purposes of oscillator frequency control, a sampling of this voltage is filtered to remove any ac component. The resulting variation in dc voltage is applied to the local oscillator, which is a voltage-controlled oscillator (VCO). The VCO varies frequency as a function of the direction and magnitude of the applied correction voltage. The voltage-sensitive component of the VCO can be in the form of a reactance tube, a saturable reactor, or a varicap. All three vary the reactance of the oscillator tank circuit, thus changing the oscillator's frequency. The use of this technique can decrease the amount of frequency drift as much as 100 to 200 percent in an uncontrolled receiver.

To determine the locking range of the AFC circuit, connect a signal generator to the input of the receiver at some suitable level at the center frequency. Tune the signal generator both above and below the center frequency and note the break-off points.

RECEIVER ALIGNMENT

Sensitivity and selectivity may be affected in alignment of all types of receivers. As receivers become more complex, alignment becomes more of a problem. In AM receivers using conventional full-carrier signals, improper alignment may result in loss of weak signals, loss of sensitivity, and the ability to select the desired signals may be impaired. If the oscillator is shifted off frequency, dial error will be introduced. Tracking error produces a varying intermediate frequency, which results in loss of signal over portions of the frequency range of the receiver. In FM receivers the discriminator tuning becomes somewhat critical. In PM receivers, phasing of the carrier must be correct, adding to the alignment problem.

When AGC and AFC are added to a receiver, proper alignment procedures must be followed, or serious errors may be introduced. When multiple conversion is incorporated in the receiver with two or more heterodyne oscillators, additional variables are introduced, further complicating alignment. In equipment employing crystals, as either reference generators, oscillators or filters, the alignment must center on the crystals, since, as previously stated, the frequencies of individual crystals are not variable.

Alignment of Crystal Filter Circuits

Crystal filters are incorporated in communications receivers that require an extremely high order of selectivity. These filters are usually located between the receiver-mixer and intermediate-amplifier stages. The crystal is the major component of the filter. It is used because of the extremely high Q that can be obtained from its use. The filter is extremely adjustable so that a variation in bandpass can be obtained. Crystal filters are often used as wave traps, and are extensively used in SSB equipment because of the sharp frequency-cutoff property required for this type of equipment. However, filters are generally of the crystal-lattice type; they are usually hermetically sealed or potted and should not be tampered with. Some lattice or half-lattice crystal filters have adjustable trimmers, accessible as screwdriver adjustments. Some adjustments are labeled as factory adjustments and should never be disturbed. In any case, the manufacturer's data should always be consulted before any adjustments are attempted. Plug-in filters are readily replaced. Mechanical filters should never be tampered with or repairs attempted under any conditions.

Schematic diagrams of crystal filters usually indicate variable capacitors and variable inductances. Such diagrams may be misleading to those unfamiliar with filter circuits. Capacitors in parallel with filter crystals are usually of very small value, on the order of 1 to several picofarads. These frequently consist only of leads given a slight wrap or twist. They may be a piece of wire bent near the crystal holder or electrode. Such capacitors are factory-adjusted, and are usually not accessible without dismantling the filter. The schematic symbols for a crystal and its equivalent electrical circuit are shown in figure 3-8, view A. A crystal in its holder is actually a combination of both series and parallel resonant circuits. As

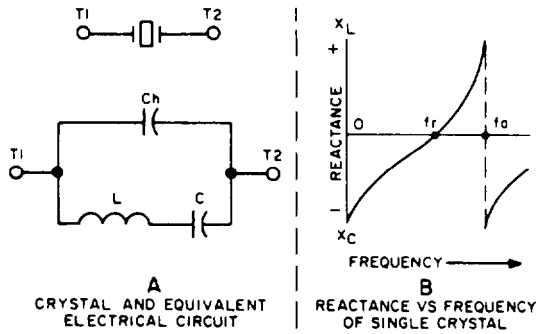


Figure 3-8.-Equivalent electrical circuit and reactance curve of a quartz crystal.

such it has two resonant frequencies, as shown in figure 3-8, view B. The series-resonant frequency occurs at the point where the reactance curve crosses the zero-reactance line. The parallel-resonance (anti-resonance) frequency occurs at the point where the reactance curve rises to a high inductive reactance. The frequency then falls sharply through the zero-reference line to a high capacitive reactance. In most crystals, the two resonant frequency points will occur within a few hundred cycles of each other. The points can be spread (or narrowed) by shunting them with a lump constant so that a suitable filter network can be designed. Phasing controls on interference filters are examples of capacitance introduced into the filter circuit to shift the crystal rejection slot (parallel-resonant frequency) so that specific unwanted signals can be rejected.

When aligning filter circuits, the circuit in which it is integrated must be considered. When connected in the intermediate-frequency amplifiers of a communications receiver, either conventional AM or SSB, the alignment will consist principally of properly tuning the resonant input circuit to the filter and to the resonant circuit at the output of the filter (fig. 3-9) for maximum output. The points of parallel resonance must be aligned for sharp cutoff (maximum attention) at the design frequency. This is an especially important consideration for equipment containing AFC circuits. If sufficient response is not allowed, the carrier may be severely attenuated at a slightly too low (or too high) frequency. This will cause the AFC circuit to drop control. Thus, the desired limits of AFC operation are also considered in the bandpass of the filter and vice versa

Alignment of Wave Traps

The term *wave trap* usually refers to a resonant element used as an auxiliary device to provide additional frequency selectivity in a radio circuit. It may take a distributed form (resonant stub or cavity), or it may consist of a lumped reactor combination (inductor and capacitor) (fig. 3-10). A trap normally provides a means of rejecting (or accepting) signals over only a relatively narrow band of signal frequencies. The width would depend on the effective Q of the trap circuit. The trapping desired may result from the "shorting" effect of a series-resonant circuit shunted across the signal path; from the selective opposition to the flow of current afforded by a

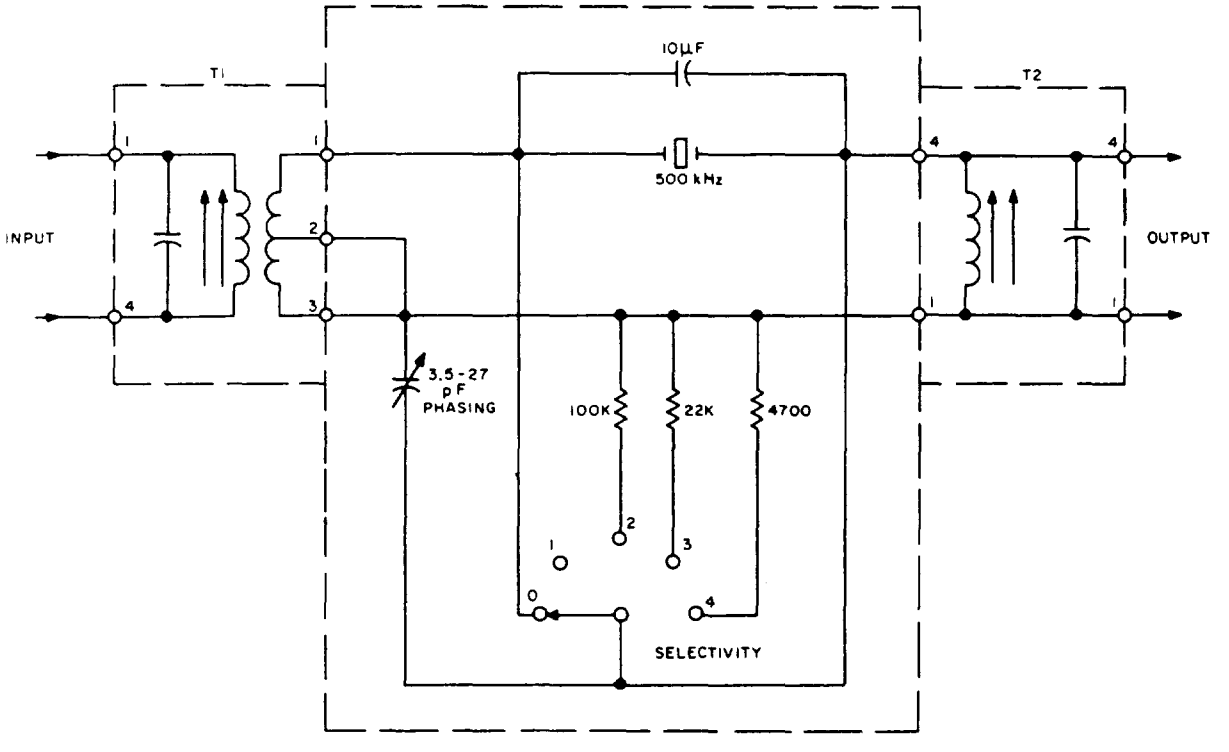


Figure 3-9.-Crystal filter circuit.

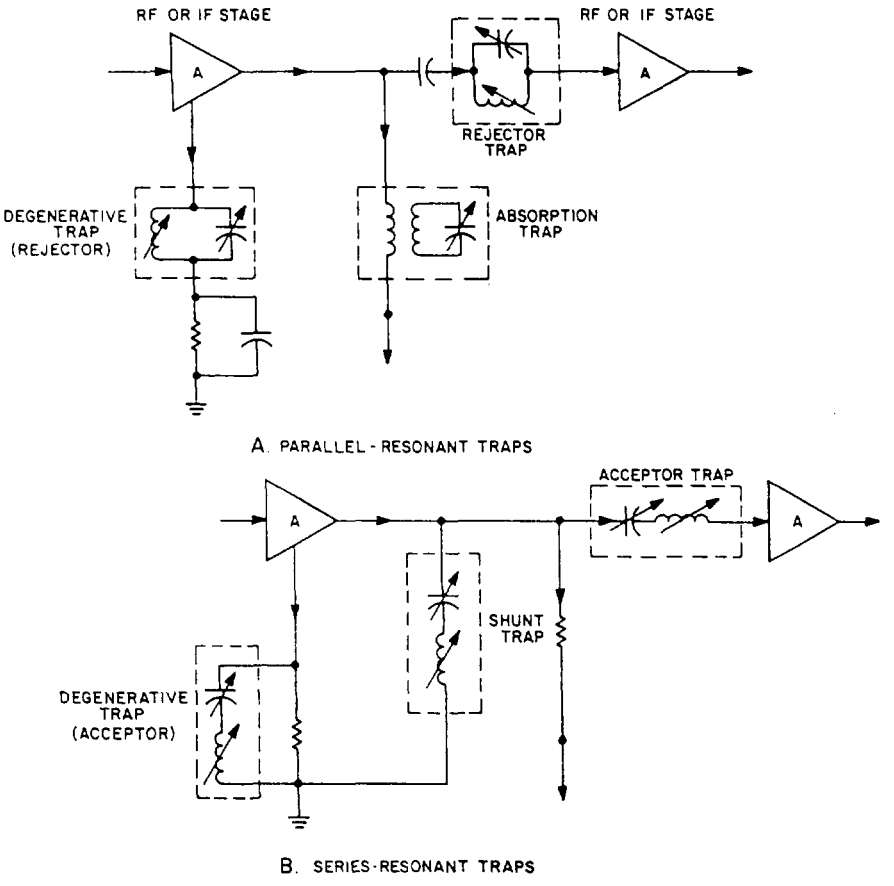


Figure 3-10.-Wave-trap circuits.

high value of resonant impedance in series with the path; from selective degeneration in an amplifier, produced by using resonant circuits to provide frequency-dependent feedback, etc.

Some of the more common lumped-reactor wave-trap applications are shown in figure 3-10. In addition to those shown, many other forms of wave traps may be incorporated in radio equipment. In some applications, a wave trap is used to suppress response at a frequency not desired in one channel. The resulting trap resonance at that frequency is used as a means or supplying the signal to a second channel. Resistance-capacitance (RC), resistance-inductance (RL), and inductance-capacitance (LC) networks, affording high-pass and low-pass characteristics, are also employed to provide band elimination or band-pass effects for wave-trapping purposes.

The operating frequencies and apparent effects of wave traps differ from one type of equipment to another. In general, the traps will be left until the last steps in a prescribed alignment procedure because of their auxiliary corrective nature. Adjustment of wave-trap trimmers, on the other hand, must usually be accomplished at very specific frequencies and under particular conditions, which should be rigidly observed. If adequate instructions for wave-trap alignment are lacking in an equipment technical manual, immediate steps should be taken to obtain further instructions. An incorrectly adjusted trap circuit may produce serious shortcomings in equipment operation that are not apparent to the operator under ordinary conditions. The signal generator and output indicator commonly employed in the alignment of receiver-tuned circuits will usually serve for wave-trap alignment in receiving equipment. Other forms of radio equipment employing traps, such as field-strength meters and oscilloscopes, may require special instrumentation.

Alignment of Beat Frequency Oscillators

Beat frequency oscillators (BFOs) are incorporated in communications receivers to

provide an audible indication of received continuous-wave transmission. They are also used to calibrate dials in receivers containing internal crystal calibration oscillators. Figure 3-11 shows a schematic diagram of a typical manually operated BFO that is heterodyned against the intermediate frequency. The procedure for aligning this type of oscillator requires feeding an unmodulated signal into the input of the receiver, with the receiver switched to the CW mode of operation. A frequency counter is connected to the phone jack to record the frequency of the detected signal caused by heterodyning the BFO with the incoming IF signal. Then L1 is adjusted so that the BFO vernier control will cause a readout of the frequency counter of 6,000 Hz above and below the zero beat.

AM Receiver Alignment

Prior to the alignment of an AM receiver, the automatic gain control (if possible) should be turned off. The gain should be adjusted by means of the manual radio-frequency gain control. The gain level should be set to give the standard 6 milliwatts of audio output with about 100 to 1,000 microvolt of signal input at the receiver antenna terminals. This alignment condition is desirable to reduce the detuning effect of receiver gain variations as reflected in changes of overall selectivity. It ensures the circuits will be resonated under average load conditions at approximately the middle working value or amplifier's input reactance and with freedom from serious regeneration. With most receivers, this condition also reduces receiver noise to a degree that renders it unnecessary to quiet the receiver by removing the amplifier stage preceding the point of alignment-signal injection.

DISABLING AUTOMATIC GAIN CONTROLS.— Amplitude modulation receiving equipment that operates with automatic gain control as a permanent condition (with a built-in provision for the alternative manual control of RF and IF gain) may present a problem, especially if considerable regeneration is normally present at full gain. It may not be feasible to disable the automatic gain control in order to add a temporary battery-biased manual gain control potentiometer in its place. In such cases, it will be necessary to align each section of the receiver (with suitable signal-input levels at the various points of signal injection) to produce final detector operation below the threshold of AGC action. The amplifier stage preceding the point of alignment signal injection should be disabled to preclude the presence of unwanted signals and noise.

DISABLING LOCAL OSCILLATORS.— The heterodyne oscillator or oscillators should not be disabled when aligning a receiver, except for the beat frequency oscillator (BFO), which is used to provide tone output from the final detector in CW reception. The heterodyne oscillator injection voltage is ordinarily a major factor controlling the mixer's operating bias and impedance, with consequent influence on gain and both mixer input and output circuit resonance. In some cases, adjustment of heterodyne oscillator tuning may not be possible as a means of preventing undesired beats or random signals that may result from the interaction of the alignment signal and the heterodyne injection voltage. The oscillator must then be disabled. Stopping an oscillator by short-circuiting its input to ground or by shorting its tank circuit may cause serious damage to the oscillator and to other electronic parts. Therefore, removal of the oscillator output is the safest way

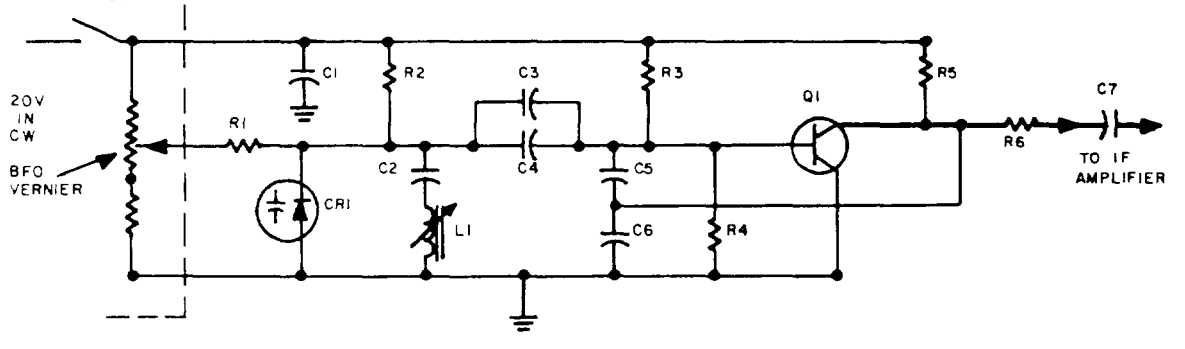


Figure 3-11.-Beat frequency oscillator circuit.

to disable the oscillator. The receiver's technical manual will detail the procedure for oscillator removal if it is required for alignment.

BEAT FREQUENCY OSCILLATOR CONSIDERATIONS.— The beat frequency oscillator (BFO) injection voltage in a properly designed CW receiver usually produces a large fixed bias at the final detector. This will mask its rectified voltage changes. This masking is very objectionable when the rectifier signal voltage is employed as an output indication for alignment. Before starting the actual alignment, the technician must disable all those auxiliary functions provided in the receiver that may interfere with proper output indication or circuit resonance. This includes silencer (squelch) circuits and noise limiters.

IF AMPLIFIER ALIGNMENT.— With a few exceptions, such as some trap circuits, IF resonant circuits are aligned by adjusting their trimmers to produce maximum signal voltage. The IF trimmers of the typical AM receiver are thus adjusted to produce maximum final-detector signal input voltage, using the input-signal frequency or frequencies prescribed in the technical manual for the equipment. In many cases, this will be the nominal band center frequency of the particular IF amplifier. In other instances, usually involving relatively wide IF passbands, "peaking" of some or all trimmers for maximum response at one or more frequencies off the band center will be specified. In general, the last IF transformer preceding the detector should be aligned first, unless a different order is specified in the equipment technical manual.

The input from the signal generator should be adjusted to produce a signal output level which is well above the noise level at the output indicator, but which is also well below the saturation level of the amplifier stages. The signal input should be progressively reduced as needed, as more circuits are brought into proper alignment. The progression of circuit adjustment should move toward the mixer stage. After the first round of alignment adjustments of the IF amplifier stages is completed, an overall check of the IF alignment should be made. A similar procedure should be used for the alignment of the preceding IF amplifier(s) in receivers employing more than one frequency conversion. The IF signal input should, in each case, be injected at the input electrode of the mixer preceding that particular IF amplifier. This ensures inclusion of the transformer located in the output circuit of the mixer. The associated conversion oscillator should be disabled if necessary, as previously discussed.

RF STAGE ALIGNMENT.— In addition to a suitable signal generator, use the dummy antenna specified by the receiver instruction book to simulate an ideal antenna for the receiver. Adjust the signal generator (modulated as required) to the upper alignment frequency specified for the particular receiver tuning band, using an external frequency standard if necessary. If an antenna trimmer control is provided on the front panel of the receiver, set it to the middle of its range. Tune the receiver to that signal frequency and the generator output to produce the desired maximum (or other specified optimum detection on the receiver out put indicator).

The tuning-dial frequency indication should coincide closely with the signal frequency being supplied. If it does not, reset the tuning dial to indicate the proper frequency. Adjust the high-frequency (shunt capacitance) trimmer of the oscillator tank circuit to produce optimum output from the test signal. Following these adjustments, adjust the interstage and antenna circuit shunt-capacitance trimmers for optimum output, with the test signal input level reset, as needed, to avoid receiver saturation effects.

Oscillator shunt trimmers occasionally have an unusually wide range of adjustments. For this and other reasons, it is possible to misalign the circuit so the heterodyne oscillator is on the wrong side of the signal frequency. In many instances, this mistake will be revealed as an inability to obtain good circuit tracking over the tuning band. Sometimes, however, the mistake will not be so clearly apparent. Therefore, always ensure that the oscillator is being trimmed on the proper side of the desired signal frequency. Determination of the proper relationship from the equipment instruction book, together with careful observations of shunt trimmer positioning (whether its capacitance is increasing or decreasing relative to the two positions of heterodyne response that it produces), will help to prevent error. Next, you should check the oscillator alignment at some specified frequency near the low-frequency end of the tuning band.

In many military receivers, iron core or eddy current trimmers are used in the RF coils to permit tank inductance adjustments for optimum low-frequency tracking of all RF circuits. Make the inductance adjustments on all coils except the oscillator coil before the oscillator series padder

is checked. Then trim the series padder to produce optimum output while the receiver tuning control is "rocked" back and forth through the region test signal response. When this process is complete, "touch up" the shunt trimmer adjustments for optimum response and correct tuning-dial reading (calibration) at the high frequency alignment point in the band. Also touch up the low-end padder adjustments (for optimum response) and check the tuning-dial reading against the test-signal frequency.

If oscillator tracing relative to the other RF circuits is poor over the band as the tuning control is operated throughout its range (indicated by abnormal variations of gain and/or output noise), it may be necessary to adjust the oscillator tank inductance trimmer. You can determine the correction needed to produce better tracking by trial readjustment of the oscillator shunt-capacitance trimmer. If the tracking (as checked by tuning from the high-frequency to the low-frequency alignment points) is improved by increasing the shunt trimmer capacitance, the oscillator tank inductance is low. If the shunt trimmer capacitance must be decreased to obtain improvement, the tank inductance is high. Correction adjustment of oscillator tank induction will necessitate some changes in oscillator series-padder and shunt-trimmer adjustments. Also, the preselector alignment procedure must be repeated. If the tuning dial is still in error over part of the band, it may be possible to correct the calibration to some degree by further slight readjustments. In either case, the entire preselector alignment procedure must be repeated. If the tuning dial is still in error over part of the band, it may be possible to correct the calibration to some degree by further slight readjustments to the oscillator and other trimmers. This realignment should be undertaken only after careful study of the tracking discrepancies and calibration errors over the entire band. Ensure that you fully understand the superheterodyne tracking problems if adequate directions are not available.

In general, it is inadvisable to sacrifice receiver gain and selectivity for the minor convenience of accurate tuning-dial calibration. Check receivers that incorporate IF traps in their RF circuit by applying a signal, at the intermediate frequency, to the receiver input. The trimmers for such traps are usually adjusted for minimum output at the center frequency of the first IF amplifier, and may require large input signal amplitude at that frequency.

FM Receiver Alignment

The basic difference between receivers used for the reception of frequency or phase-modulated signals and those used for the reception of amplitude modulated signals lies in the types of demodulator and IF amplifier circuit employed. In an FM receiver, a frequency-sensitive demodulator is used. IF amplifiers are designed to cause, rather than avoid, amplitude limiting. When testing several amplifier stages that have similar operating functions, such as successive IF stages, it is possible (but not recommended) to test immediately for an "overall" response curve like that shown in figure 3-12. You may see this curve at the output of the last IF stage or at the grid of the limiter.

When using an FM signal generator for testing wide-band equipment, such as an FM receiver, you can see the response curve directly on the screen of an oscillator scope. Improperly applied, this procedure could consist of varying "at random" the different adjustments in all the stages until the overall response curve appears to be satisfactory. But, this good-looking curve may result from a compromise. This generally means that a poor alignment in one stage is compensated by overemphasized and shifted alignment in other stages. The reason for this is that one stage may

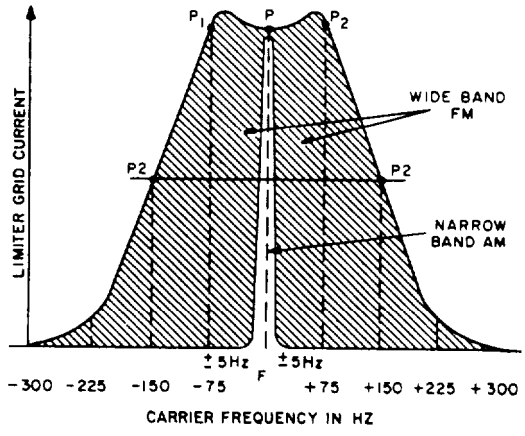


Figure 3-12.-FM versus AM resonance curves.

be peaked unsystematically, another stage may have a center peak, and the remaining stages may have two response peaks. This refers to all IF stages. No stage by itself satisfies the condition required for linear networks with respect to amplitude and phase. Distortion is bound to result.

Regardless of the method of aligning the receiver, the recommended practice is to first align the discriminator. You must have a sufficient signal source, the sensitivity of the indicator must be high, and the IF transformers must not be excessively detuned. The remainder of the set, up to the discriminator, is then aligned. The correct IF alignment should always be made by first aligning the IF stage ahead of the limiter, then the preceding IF stage, etc. As far as the RF stage or stages before the mixer stage are concerned, ordinary single peaking is generally practiced, since coil and tube damping nearly always provide the required broadness of the response curve. The receiver can be aligned most conveniently by using an FM signal generator. However, if such a generator is not available, an AM generator can be used. A meter on an oscilloscope may be employed as an output indication. Normal receiver alignment consists of the following sequence:

1. Alignment of the demodulation (discriminator) stage

2. Alignment of the limiter stage

3. Alignment of the IF amplifier stages

4. Alignment of the RF stages

LIMITER-TYPE DISCRIMINATOR ALIGNMENT.— Figure 3-13 is a schematic diagram of a limiter-type detector. In the double-tuned circuit shown, the primary and secondary are tuned to the carrier frequency. At the carrier frequency, the voltages developed across the diodes are equal to each other, and the diode currents are also equal. Thus, the opposing voltages developed across the output diode resistors are equal and therefore cancel. As a result, no voltage is developed at point A. At frequencies lower than that of the carrier frequency, there is more current flow through the lower diode than through the upper diode. This causes a negative voltage to be developed at point A. Conversely, at frequencies higher than that of the carrier frequency, there is more current flow through the upper diode. This causes a positive voltage to be developed at point A. Within the range of carrier-frequency swing, the dc output changes in proportion to the frequency change.

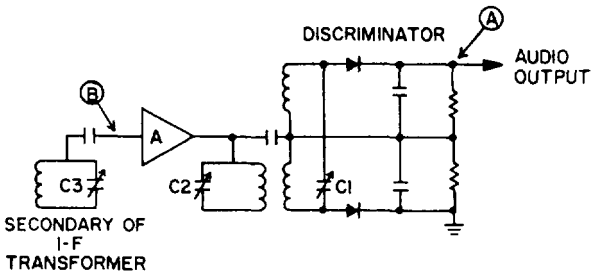


Figure 3-13.-Limiter-type discriminator circuit.

There are several ways of measuring the linearity of a discriminator. The most straightforward method requires the use of a high-resistance voltmeter to measure the output voltage between point A and ground (fig. 3-13), while varying the applied intermediate frequency in known steps. A voltage of the center intermediate frequency is applied to the limiter (or to the mixer if the IF amplifiers are properly aligned and the limiters are properly set). The setting of the secondary capacitor C1 is then varied until zero output is noted. (Both terminals of this capacitor are above ground potential.)

WARNING

An insulated screwdriver must be used for this adjustment to prevent electrical shock.

The signal generator is then set above and below the intermediate frequency. The voltmeter should indicate equal but opposite direct voltages for equal but opposite frequency deviations. If unequal voltages are obtained, the setting of primary trimmer capacitor C2 is incorrect and must be adjusted until equality is obtained. If necessary, repeat these operations until the proper indications are obtained. You can determine the linearity over the entire range by plotting the values obtained for steps of frequency deviation,

as shown in figure 3-14, view A. The output of the generator should be constant, or the limiter should be in full operation. A visual method of aligning the discriminator, using an FM signal generator and an oscilloscope, has an advantage over the meter method in that the discriminator curve may be observed. Since the effects of the adjustments are visible, no guesswork is involved.

There are two methods of setting the discriminator to its proper center frequency. The first method is more accurate and has more easily observed results. It consists of applying an amplitude modulated RF signal of the correct center frequency to the discriminator. The discriminator secondary is then adjusted for minimum signal output. The output signal will disappear if the discriminator characteristic is symmetrical about the center frequency, since the output will be zero at the center frequency. In the second method, a marker pip is made to appear on the discriminator response curve. The pip is set for the crossover point at the center frequency. The center frequency is determined by noting where the marker disappears and reappears. Because of the difficulty in observing the exact points of appearance of this pip, this method sometimes leads to inaccurate results. To use the AM signal method for aligning the discriminator to its center frequency, connect an AM signal generator to point B (fig. 3-13) and the oscilloscope vertical input to point A. Set the signal generator for 400 Hz amplitude modulation, and adjust the oscilloscope controls for a convenient pattern size. When the discriminator secondary trimmer capacitor (C1) is not adjusted to the current frequency but is close to it, a pattern similar to figure 3-14, view B, will appear. To align the discriminator to the correct center frequency, adjust the secondary trimmer capacitor slowly in one direction, and then in the other direction until the 400-Hz signal disappears and then reappears with a further movement of the trimmer. Set the capacitor midway between these points. Then connect an FM signal generator to point B (fig. 3-13). Leave the oscilloscope vertical input at point A and connect the horizontal input to the modulation circuit of the signal generator. Adjust the signal generator for full frequency deviation and set the oscilloscope controls for a convenient pattern size. Then, adjust the primary trimmer capacitor, C2, (fig. 3-13) for a symmetrical curve similar to the one shown in figure 3-14, view A.

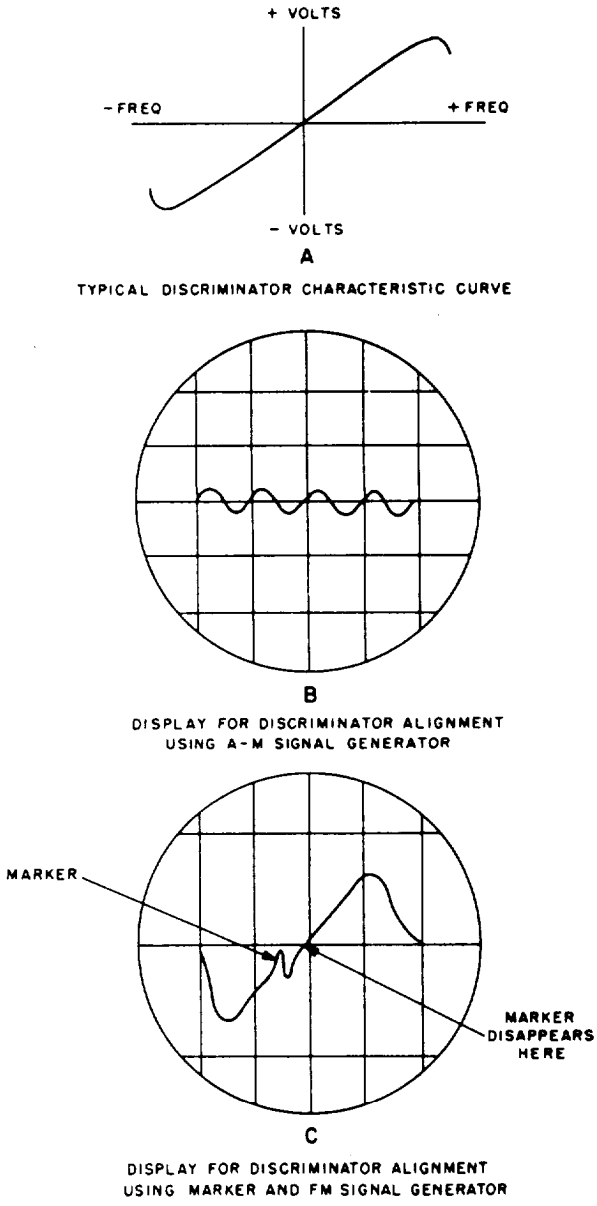


Figure 3-14.-Discriminator characteristic measurements.

When using the marker method, connect the FM signal generator and oscilloscope as described above. Then couple point A, marker generator or wavemeter with the signal generator output to point B (fig. 3-13) to produce a pip on the discriminator response curve. With the marker signal generator or wavemeter set at the discriminator center frequency, adjust the secondary trimmer capacitor until the marker disappears at the crossover point at the center of the response curve. Then, adjust the primary trimmer capacitor for a symmetrical curve.

RATIO DETECTOR ALIGNMENT.— Another type of FM detector (fig. 3-15) is called a "ratio detector." This circuit is based on changes in the ratio of the voltage across the two diodes rather than on differences in voltage. A ratio detector is virtually insensitive to amplitude variations. The tuning and coupling provisions are about the same as in a limiter-type discriminator. As a result, the RF voltage developed across the diode at any instant depends upon the amount of frequency deviation from the carrier center frequency. Unlike the arrangement in the limiter-type discriminator, however, the diodes are connected to conduct simultaneously, so that a negative voltage is developed across the load resistor. A filter capacitor connected across the load resistor has enough value to hold the voltage constant even at the lowest audio frequencies to be reproduced. The voltages across the diodes differ according to the instantaneous frequency of the carrier. The rectified voltages across capacitors C3 and C4 are proportional to the corresponding diode voltages. Although each of these voltages vary, their sum is held constant by the filter. As a result of this reaction, an audio output is developed across C4. Many modified versions of the ratio detector are in common use. However, the operation and alignment procedures are similar to those described for the ratio detector.

To align a ratio detector, connect an FM signal generator between point B and ground, and connect the input to the oscilloscope between point A and ground (fig. 3-15). Set the generator to the center frequency, with maximum frequency deviation. Adjust the discriminator, the primary trimmer capacitor, C2, for a curve of maximum amplitude. The curve will appear somewhat S-shaped if the secondary trimmer capacitor C1 is not excessively detuned. Keep the generator attenuation control set for an output below the level where limiting occurs. Next, adjust the secondary trimmer capacitor until the S-shaped curve is symmetrical (fig. 3-14, view A). Set the curve to exact center frequency as described previously for discriminators. If necessary, retune

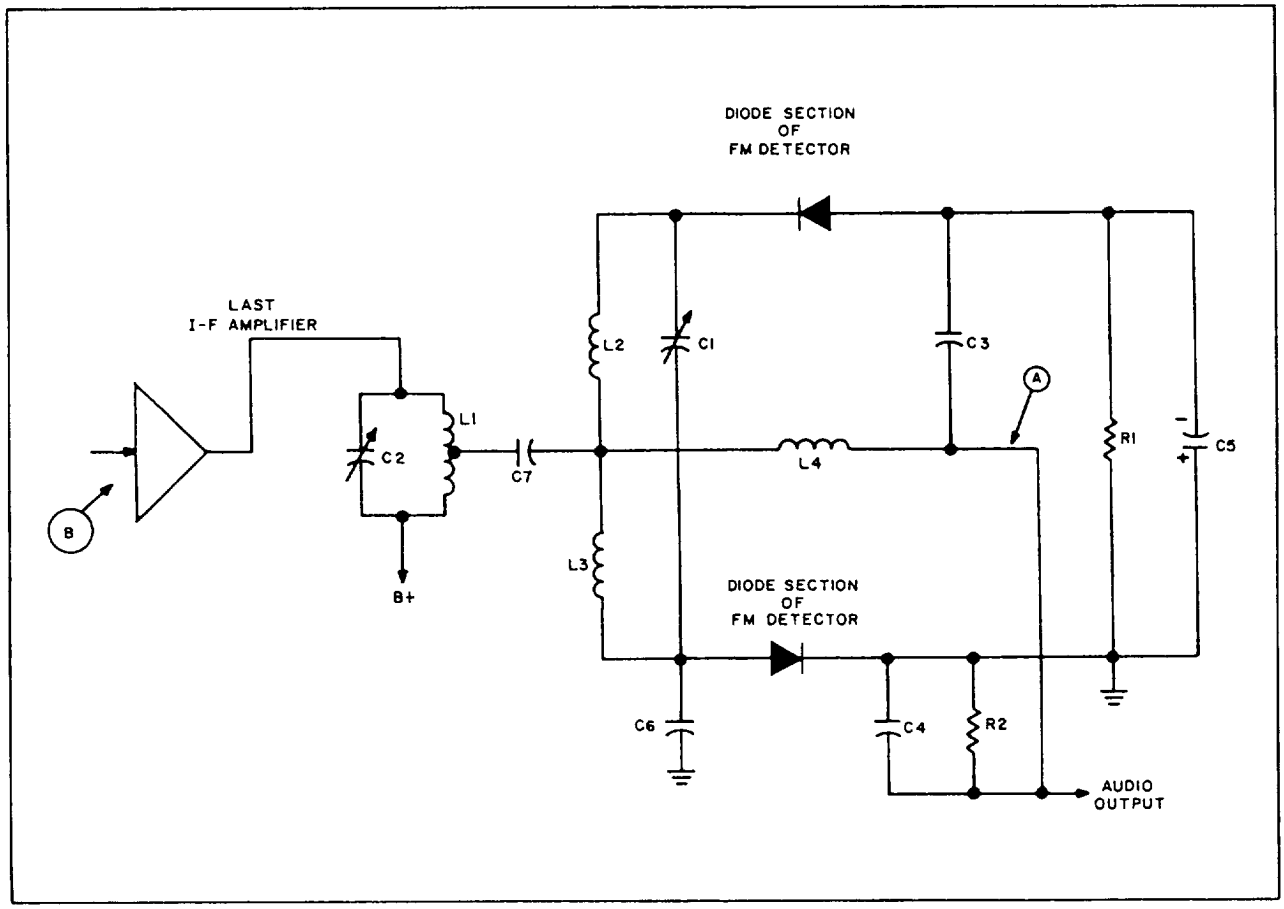


Figure 3-15.-Ratio detector circuit.

the detector primary trimmer for a symmetrical response of maximum amplitude.

IF AMPLIFIER ALIGNMENT.— Specific alignment procedures are generally included in technical manuals for particular receivers. In these cases, specific response curves for each transformer may be given so that a particular overall response curve may be obtained. However, in general, the IF amplifiers may be aligned by feeding an FM signal from the generator to the grid of the IF amplifier just preceding the limiter, while observing the discriminator output. The secondary of this IF amplifier is then adjusted for a symmetrical S-shaped curve, having a proper frequency response of maximum amplitude. The procedure is repeated to tune the primary. If the output can be kept below the threshold of limiting by reducing the generator output, adjust each IF secondary and primary in sequence, proceeding from the last IF stage curve of maximum amplitude. Should limiting occur, the response curve of the amplifier can be observed at the grid of the limiter and the IF amplifiers tuned for proper bandpass (fig. 3-12).

RF AND OSCILLATOR STAGES ALIGNMENT.— The RF and oscillator stages may also be aligned by using of an FM signal generator and an oscilloscope. To align these stages, connect the output from the generator to the antenna terminals of the receiver through a matching network. Connect the oscilloscope input to the discriminator output. Set the generator to a frequency in the approximate center of the band being tested. Set the frequency deviation greater than the receiver bandpass. Observing the receiver output response, tune the shunt (high-frequency trimmer capacitor) for maximum output. Set the signal generator and receiver to the low end of the band. Use a tuning wand and observe the oscilloscope pattern. If the signal amplitude decreases when either end of the wand is inserted into the oscillator coil, the tracking is satisfactory. If the output increases with the brass end of the wand inserted, spread the turns of the oscillator coil. If the output increases with the iron end of the wand inserted, compress the turns of the coil. Do not bend the coil excessively. Only a slight physical change is necessary at the high frequencies at which this type of equipment generally operates. Repeat these steps until no further change is noted.

The last adjustment should be of the shunt (high-frequency) trimmer capacitor. Return the signal generator and receiver to the center frequency. Adjust the shunt trimmer capacitor of the mixer-grid circuit for maximum output. If an RF stage is employed, also adjust the shunt trimmer capacitor of the RF stage for maximum output. Lower the frequency setting oft he receiver and generator. Check the tracking of the mixer and RF grid circuits with the tuning wand. If the output increases with the brass end of the wand inserted in the coil, spread the coil turns. If the output increases with the iron end inserted, compress the coil turns. If the output decreases when either end is inserted, the tracking is correct. Do this for both the mixer and RF coils. Repeat these adjustments until no further improvement is noted.

An oscilloscope may also be used to check the response of the RF and mixer sections of a receiver. Connect the FM signal generator to the receiver input through a matching network. Connect the oscilloscope vertical input to the mixer plate decoupling network or, by means of a high-frequency detector probe, to the mixer's output. The first IF amplifier is disabled during this test to reduce the loading effect of the oscilloscope input. Couple a marker signal-generator with the FM signal generator to the receiver input to determine the frequency points of the response curve. Set the FM signal generator for the desired frequency deviations. In many communications receivers, the front-end response curve is from 150 to 200 kHz wide. The bandwidth of the front end is largely fixed by the number and Q of the RF circuits, since all of the circuits are usually tuned to identical frequencies. Usually, the bandwidth of the RF circuits is considerably greater then the IF bandwidth, so that the latter mainly determines the bandwidth of the entire receiver. When making any adjustment of the front end, both the RF and overall response must be considered, since it is important that the RF response be wide enough to pass all of the important frequency components of the signal. This check may be made by the previous procedure for measuring the RF response.

COMMUNICATIONS TRANSMITTER AND TRANSCEIVER TESTING

Learning Objective: *Recognize the testing procedures for communications transmitters, to include frequency and modulation measurements and IF and RF amplifier measurements.*

When testing communications transmitters and/or transceivers, the configuration in which

the equipment is installed must be considered. In the newer installations, system monitoring is used (more so than equipment monitoring). These newer configurations employ built-in test equipment (BITE) to perform system monitoring and to a varying degree, fault isolation. In the older installations, front panel meters and dials are relied upon for equipment monitoring and some fault isolation. A number of pieces of equipment can be placed in either installation, while other equipment is unique to either the system concept or the individual equipment concept. In both instances, however, the same parameters are monitored. Only the monitoring method is changed.

Temperature is another consideration in transmitter/transceiver maintenance. High-powered transmitters emit a great deal of heat in their power amplifier and driver stages. If the heat is poorly dissipated, premature failure of equipment will occur. Two factors that contribute significantly to poor heat dissipation are water and dirty filters/heat exchanges. Although the first factor is usually unpredictable, the latter is always avoidable with routine maintenance. In addition, routine maintenance can usually prevent most circulatory system failures. If forced air (blower) circulation is employed to dissipate the heat in a highpowered transmitter, dust settling in the equipment can contribute to problems. Because dust forms a film that absorbs moisture, insulation resistance is lowered to a point where flashover may occur. Strict adherence to scheduled equipment cleaning is mandatory if the equipment is to be cooled effectively. Insulators must be wiped down, corroded metal parts cleaned, and arc-overs repaired. It is possible to detect poor contacts by inspecting for evidence of local overheating or arching. Such contacts must be thoroughly cleaned and tightened.

FREQUENCY GENERATION

On any given day, several thousand different frequencies may be used simultaneously by several transmitting units. Since atmospheric conditions dictate which frequency will best be propagated at any given time, the transmission frequency in use may need to be changed quite often so that the individual units may maintain efficient and reliable communications. Regulations require each transmitted frequency to be precise and that it comply with harmonic frequency emission limits and sideband emission limits. To meet these rigid standards, crystal oscillators and frequency synthesis are employed to generate the required transmission frequency. When frequency synthesis is used, a calibrated frequency standard with a very high degree of accuracy (on the order of 1 part in 10^9 per day) serves as the basic oscillator. This standard is external to the transmitter or transceiver, and its output is fed to each transmitter or transceiver via a frequency standard distribution system, as shown in figure 3-16.

A back-up internal standard is incorporated in each transmitter/transceiver for use in case of primary standard malfunction, The output of the primary standard is multiplied and/or divided in the transmitter to obtain the desired frequencies for use in the frequency synthesis

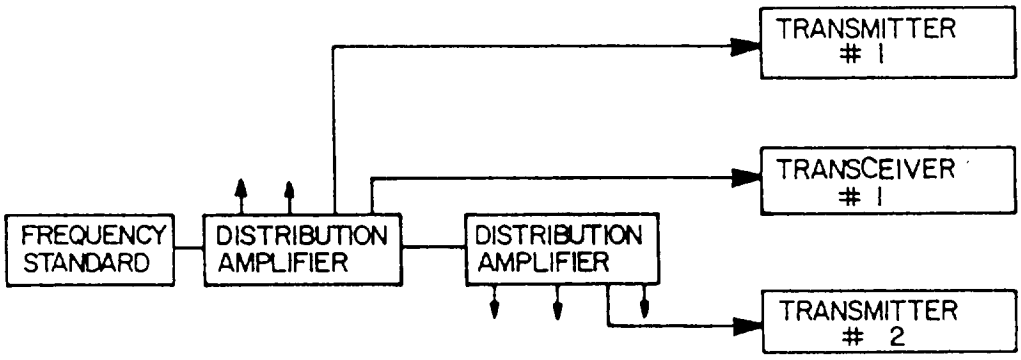


Figure 3-16.-Frequency standard distribution system.

process. Figure 3-17 shows an example of an HF transmitter's (frequency) synthesizer, In the example shown in figure 3-17, all of the output operating frequencies are derived from the 5 MHz primary standard input frequency. This ensures each operating frequency is as accurate as the primary standard.

FREQUENCY MEASUREMENT

The primary standard used in the frequency standard distribution system requires routine calibration by a class A calibration facility. The standard must be more accurate than any other general-purpose test equipment maintained aboard ship. Primary power must be applied to the standard at all times to ensure stable operation. As a provision against a loss of ac power, a battery source of alternative power is incorporated in the standard. The battery becomes switched in automatically when ac power is lost, but its staying time is limited to approximately 8 hours on a full charge.

The primary power system is alarmed and must be checked frequently to ensure battery-power readiness. This is accomplished by manually switching to the battery power source. The alarm should then sound and the standard

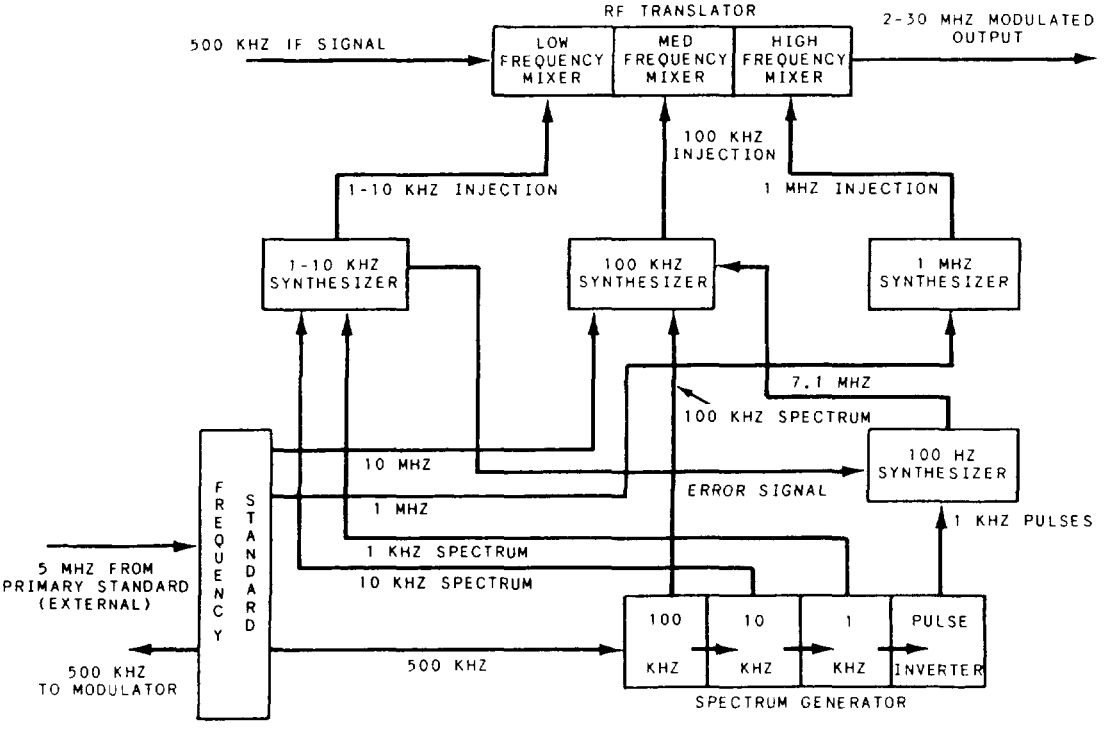


Figure 3-17.-Frequency synthesizer.

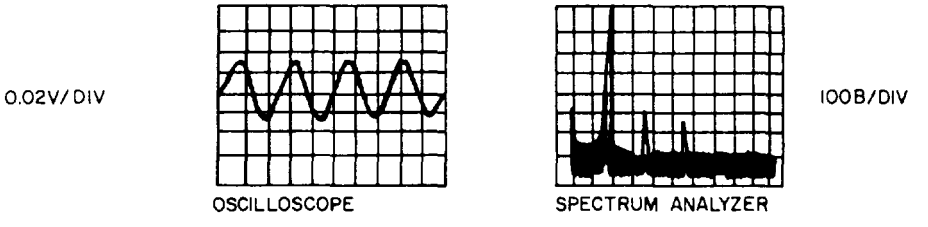


Figure 3-18.-Harmonic distortion displays.

should operate automatically from the battery source and without interruption on the batteries, Before switching manually to battery operation, make sure the standard's front panel meter indicates a battery charge exits. If all power to the primary standard becomes lost, the transmitters and transceivers in the frequency distribution system must then be shifted over to the internal frequency standard. The internal frequency standard should normally be allowed 24 hours to become stable. In emergency operations, the internal standard requires at least 4 hours of warm-up time. Although the frequency accuracy of the primary standard cannot be checked realistically, a check with a calibrated frequency counter will indicate whether or not the standard is grossly off frequency. Harmonic distortion of the primary standard's frequency must be guarded against because of the unwanted changes this frequency will undergo during the frequency synthesis process in the transmitters and transceivers. Spectrum analysis is used to measure harmonic distortion because low-level distortion will not appear in the time-domain display of an oscilloscope. Such distortion will show up, however, in the frequency domain display of the spectrum analyzer (fig. 3-18). High-level distortion will show up in either display. If more than 0.2 percent of distortion is encountered in the standard's output, the standard must be sent to a shore facility for repair and/or calibration.

The two best means of measuring a transmitter's out put frequency are the frequency counter and the spectrum analyzer. The frequency counter is by far the more accurate of the two because interpolation is not required. The spectrum analyzer, on the other hand, can measure second and third order harmonic emissions, sideband emissions, and intermodulation distortion, as well as output frequency. In addition, when the spectrum analyzer is used in conjunction with a directional coupler and calibrated attenuators, it can also measure power output and reflected power. The spectrum analyzer can thus provide a better indication of transmission quality than can a frequency counter.

AMPLITUDE MODULATION MEASUREMENTS

If an AM signal is applied to the vertical input of an oscilloscope, the oscilloscope will display a wave-envelope pattern of the AM signal (fig. 3-19). When the modulation voltage is used as an external sweep voltage, a trapezoid pattern (fig. 3-20) is displayed. In this display, the amplitude-modulated carrier is plotted as a function of modulation rather than of time. The resultant pattern in figure 3-20 will remain stationary, and its shape is determined by the percent of modulation. The presence of an abnormal pattern does not necessarily mean that there is a defect in the operation of the transmitter. A pattern with elliptical sides merely implies that with respect to the signals applied to the oscilloscope, there is a phase difference between the modulating signal and the modulated radio-frequency signal. The curved appearance of the sides is simply a Lissajous effect denoting phase difference. In fact, the straight sides of a properly formed trapezoid simply represent the characteristic Lissajous pattern of a zero phase difference. When the modulating signal is not a sine wave, a trapezoid is still formed, but a series of bright vertical bands appear in the pattern. These bands indicate points at which the lateral motion of the electron beam is relatively slow.

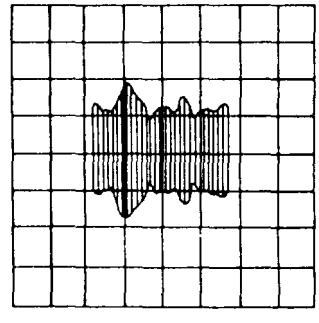


Figure 3-19.-Amplitude-modulated carrier.

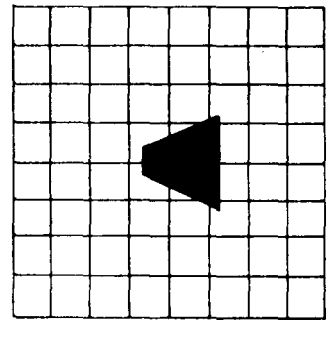


Figure 3-20.-Trapezoidal AM carrier pattern.

Distorted trapezoid waveforms due to certain circuit malfunctions are shown in figure 3-21. View A of figure 3-21 shows an amplitude modulated waveform and its corresponding trapezoidal display. This display results from excessive plate voltage applied to the radio-frequency power amplifier in a radio transmitter. View B shows the reverse situation, where there was insufficient voltage applied to the plate of the transmitter power amplifier. Views C and D show the effects obtained from imperfect neutralization in a radio transmitter power amplifier. There is nonuniform density in the amplitude modulated waveform. Light and dark bars, such as those shown in the figure, usually indicate spurious oscillation. To calculate the percent of modulation, the trapezoid pattern provides the most convenient form to use. The horizontal and vertical gain controls are adjusted for a suitable display on the screen, such as shown in figure 3-22. The modulation percent is then calculated, using the formula

$$\frac{H1 - H2}{H1 + H2} \times 100.$$

H1 is the greatest vertical height (amplitude) and H2 is the lesser vertical height. Using figure 3-22 as an example, the percent of modulation would be

$$\frac{5 - 1}{5 + 1} \times 100 = \frac{4}{6} \times 100 = 66.6 \text{ percent.}$$

The longer side or the trapezoidal pattern represents modulation peaks, or crests. The shorter side indicates modulation troughs, or low points, at 100-percent modulation. The wedge-shaped pattern assumes a point on the shorter side. Modulation over 100 percent causes this

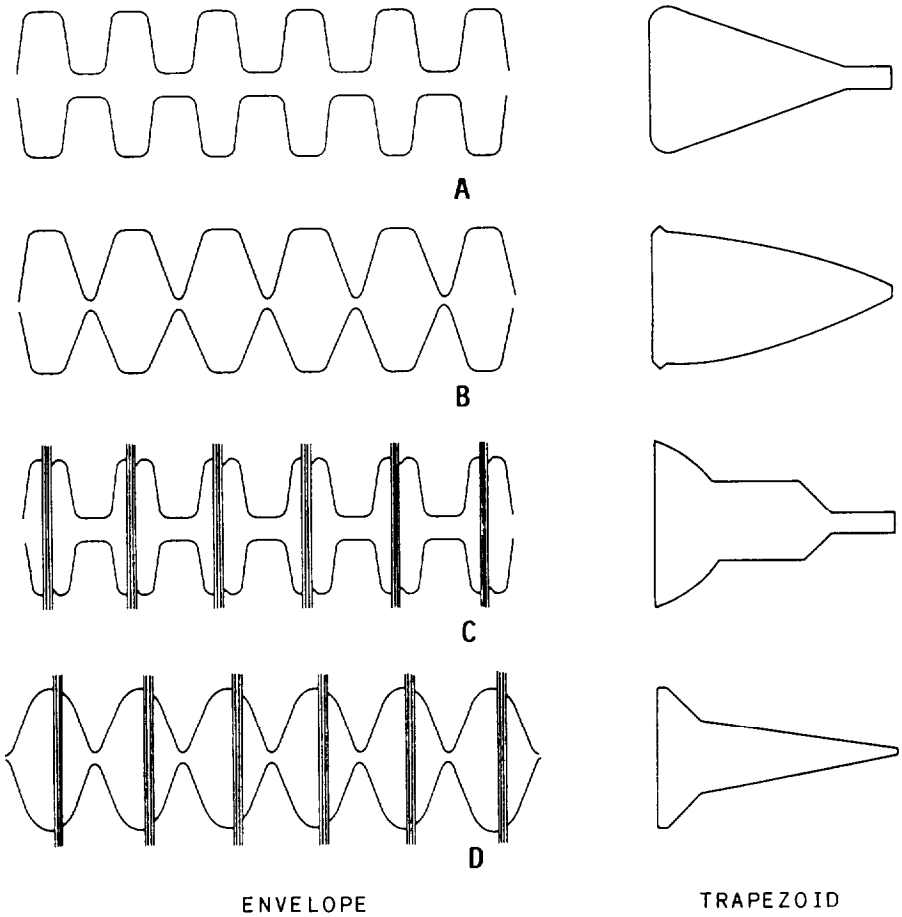


Figure 3-21.-Distorted amplitude-modulated waveforms.

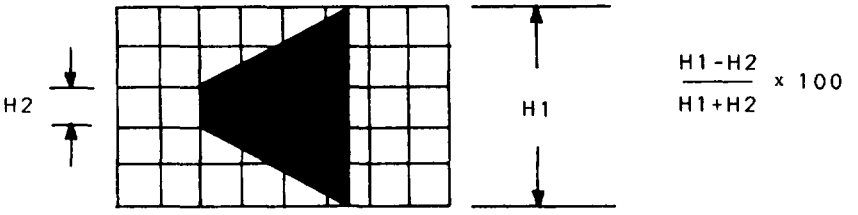


Figure 3-22.-Trapezoid method of determining percent of modulation.

point to extend and form a horizontal line or tail, as shown in figure 3-23. Because the trapezoid type of display retains its triangular characteristics even with varying degrees of modulation, it provides a more easy discernible indication of overmodulation as well as the modulation percentage. To obtain correct results, you should take care to avoid stray radio-frequency pickup that may distort the oscilloscope presentation.

FREQUENCY MODULATION MEASUREMENTS

The concept of percentage of modulation (as discussed in connection with amplitude modulation) does not apply to frequency modulation. The amplitude of the FM wave is constant. The extent of modulation must be described in terms other than those of the amplitude-modulated wave. When referring to a class of stations, a certain maximum frequency swing is established as representing 100-percent modulation. For example, in the case of FM broadcast stations, a frequency swing of plus or minus 75 kHz from the unmodulated center frequency (frequency deviation) is commonly considered as being the equivalent of 100-percent modulation. However, the more widely accepted method of describing the extent of modulation is to state the value of the modulation index. This index (m) is the ratio of the amount by which the transmitted frequency swings from its average frequency (frequency deviation) to the frequency of the modulating signal. The relationship of these quantities is shown by the following equation:

$$m = \frac{Fd}{FM}$$

where:

m = modulation index

Fd = frequency deviation

FM = frequency of modulating signal

By means of this basic relationship, it is possible to determine the frequency deviation when the modulation index and the modulating frequency are known. It should be carefully noted (in describing the extent of frequency modulation) that the modulation percentage and the modulation index are defined in a different manner. The percentage is proportional to the frequency swing.

The modulation index is also directly proportional to the frequency swing, but in addition, it is inversely proportional to the highest modulating frequency. Thus, in contrast to amplitude modulation, the modulation index of a frequency-modulated wave is not the decimal equivalent of the modulation percentage. The modulation index of a frequency modulated wave, for example, will exceed 1 (unity) by many times when the frequency swing is large and the modulating frequency is low. The frequency modulated output is the sum of a center frequency component and numerous pairs of sideband frequency components. The center frequency component has the same frequency as the unmodulated carrier. The two components of the first sideband pair have frequencies respectively higher and lower

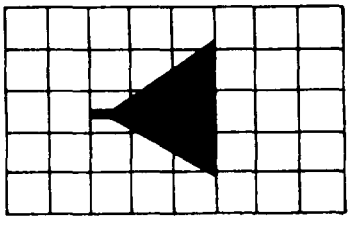


Figure 3-23.-Overmodulated carrier.

than the center frequency by the amount of the modulating frequency, just as in amplitude modulation. In frequency modulation, however, there are additional pairs of sideband components that can have appreciable amplitude. For example, the second pair of sidebands (having frequencies that are higher and lower than the center frequency by twice the amount of the modulating frequency) can also be important. The same can be true of the third pair of sidebands. These are removed from the center frequency by three times the modulating frequency, and of even higher orders of sideband pairs, whose frequencies differ from the center frequency by correspondingly greater amounts. When the modulation is only slight, only the pair of sidebands nearest in frequency to the carrier frequency component will have sufficient amplitude to be important. Under this condition, the bandwidth required is no greater than that for an amplitude modulated wave. As the frequency modulation is increased, however, more pairs of sidebands acquire appreciable amplitude, and the bandwidth requirements become greater than the amplitude modulation.

The actual amplitudes of the frequency-modulated-wave sidebands and carrier, as compared with an unmodulated carrier amplitude of 1, may be read directly from table 3-1 for modulation indices up to 6. To find the amplitude of any sideband pair, determine the modulation index (m), read the corresponding amplitude factor for the sideband pair, and multiply this factor by the amplitude of the unmodulated carrier. The amplitude of the carrier during modulation is found in the same manner, taking the amplitude factor from the Jo(m) column. Where no value is given in a column, the amplitude factor is less than 0.005, and the sideband pair will not be important for normal considerations, The values of $J_0(m), J_1(m),$ and $J_2(m)$ over the range m = 0 to m = 16 are shown plotted in figure 3-24. A study of these curves reveals some interesting facts about the composition of frequency modulated waves. $J_0(m)$ is less than 1 for all values of m greater than zero, This indicates that as sideband components appear with modulation, the amplitude of the center frequency component is less than its amplitude in the absence of modulation. This fact is evident if you remember that the amplitude of the frequency modulated wave is constant. So, the average power during each radio frequency cycle is the same as the power during any other radio

Table 3-1.-Bessel Factors for Finding Amplitudes of Center and Sideband Frequency Components

m	$J_0(m)$ F	$J_1(m)$ F ± Fm	$J_2(m)$ F ± 2Fm	$J_3(m)$ F ± 3Fm	$J_4(m)$ F ± 4Fm	$J_5(m)$ F ± 5Fm	$J_6(m)$ F ± 6Fm	$J_7(m)$ F ± 7Fm	$J_8(m)$ F ± 8Fm	$J_9(m)$ F ± 9Fm
0.0	1.000									
0.1	0.9975	0.0499								
0.2	0.99	0.0995								
0.3	0.9776	0.1483	0.0112							
0.4	0.9604	0.196	0.0197							
0.5	0.9385	0.2423	0.0306							
0.6	0.912	0.2867	0.0437							
0.7	0.8812	0.329	0.0589	0.0069						
0.8	0.8463	0.3688	0.0758	0.0102						
0.9	0.8075	0.4059	0.0946	0.0144						
1.0	0.7652	0.4401	0.1149	0.0196						
1.2	0.6711	0.4983	0.1593	0.0329	0.005					
1.4	0.5669	0.5419	0.2073	0.0505	0.0001					
1.6	0.4554	0.5699	0.257	0.0725	0.0150					
1.8	0.3400	0.5815	0.3061	0.0988	0.0232					
2.0	0.2239	0.5767	0.3528	0.1289	0.034	0.007				
3.0	−0.2601	0.3391	0.4861	0.3091	0.1320	0.0430	0.0114			
4.0	−0.3971	−0.066	0.3641	0.4302	0.2811	0.1321	0.0491	0.0152		
5.0	−0.1776	−0.3276	0.0466	0.3648	0.3912	0.2611	0.131	0.0534	0.0184	
6.0	−0.1506	−0.2767	−0.2429	0.1148	0.3576	0.3621	0.2458	0.1296	0.0565	0.0212

frequency cycle. So the power in the wave will not change when frequency modulation causes sideband currents to appear, the amplitude of the center frequency component must decrease sufficiently to keep the total of the I^2R products of all the components equal to the power of the unmodulated wave.

FREQUENCY DEVIATION MEASUREMENTS

Regardless of the differences between amplitude modulation and frequency modulation, it is possible to make an analog between percentage of amplitude modulation and frequency deviation. Specifically, frequency deviation is proportional to the amplitude of the modulating signal, as is the percentage of amplitude modulation. Because of this analogy, it is convenient to extend the concept of percentage of modulation to frequency modulation by arbitrarily designating the maximum allowable frequency deviation of a class of operation as 100-percent modulation. An important distinction to remember is that no distortion results from modulation percentages greater than 100 in FM transmission. However, any percentage larger than the figure sanctioned by the proper authorities will produce excessive channel width, making interference with other stations possible. For example, the maximum frequency deviation for commercial FM stations is limited to 75 kHz; for military applications, the maximum deviation is limited to 40 kHz, and is classed as narrow-band FM transmission. The sound transmission of television stations is restricted to a deviation of 25 kHz.

It was stated earlier that modulation index (m) determines the relative amplitude of the carrier and sideband frequencies emitted by an FM transmitter. The modulation index may be measured by using the fact that the carrier amplitude becomes zero whenever the modulation index is such that $J_0(m) = 0$, where J_0 is a Bessel function of the zero order. The values of the modulation index for these conditions are given

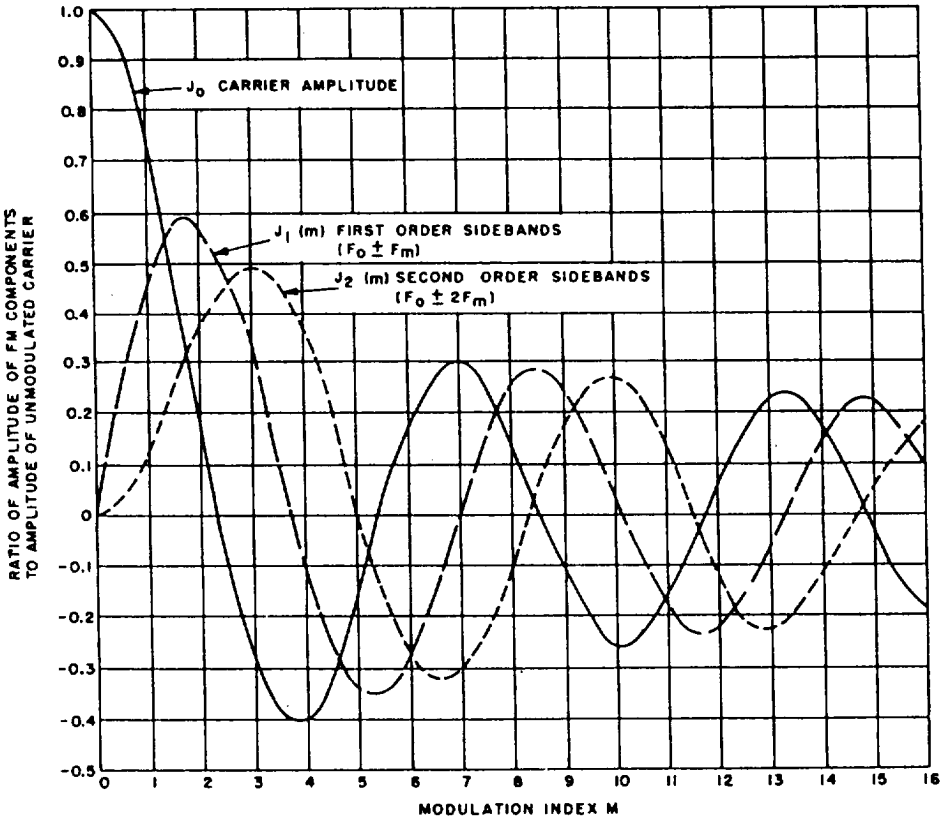


Figure 3-24.-Variation of FM wave component with degree of modulation.

in table 3-2. Specifically, the carrier component disappears completely for certain values of m; that is, m = 2.405, 5.52, 8.654, etc. (Note that $J_0(m)$ = 0 in figure 3-24 for these values of m.) For these specific values of m, all of the transmitter power is contained in the sidebands. This fact allows the measurement of specific values of the modulation index by measuring the amplitude of the carrier component only. The level of modulation on the FM carrier is increased from zero to the first point at which the detected carrier disappears. The point at which the carrier first disappears corresponds to m = 2.405. Upon increasing the modulation further, carrier reappears and then disappears a second time. The second vanishing of the carrier corresponds to m = 5.52. Further increases in modulation will produce the higher carrier zeros (or null points). For example, the frequency deviation at the first null point is F_d = 2.405 F_m. This means of determining frequency deviation is generally known as the "Bessel zero method." Modulation indices between carrier zero-points would involve considerable interpolation, leaving room for large error if the Bessel zero method were used. In addition, the modulating frequency must be started at zero amplitude to determine which zero point is being displayed.

A more accurate method involves comparing the carrier amplitude with respect to the sideband amplitudes. Figure 3-25 shows a spectrum analyzer display of an FM signal where deviation has caused the carrier to be at 30 percent of its unmodulated level. In figure 3-25, views B and C, the carrier is in the negative region of its curve.

Table 3-2.-Values of Modulation Index for Which a Carrier Wave Has Zero Amplitude

ORDER OF CARRIER ZERO	MODULATION INDEX
1	2.405
2	5.52
3	8.654
4	11.79
5	14.93
6	18.07
m(m6)	$18.07 + \pi(m - 6)$

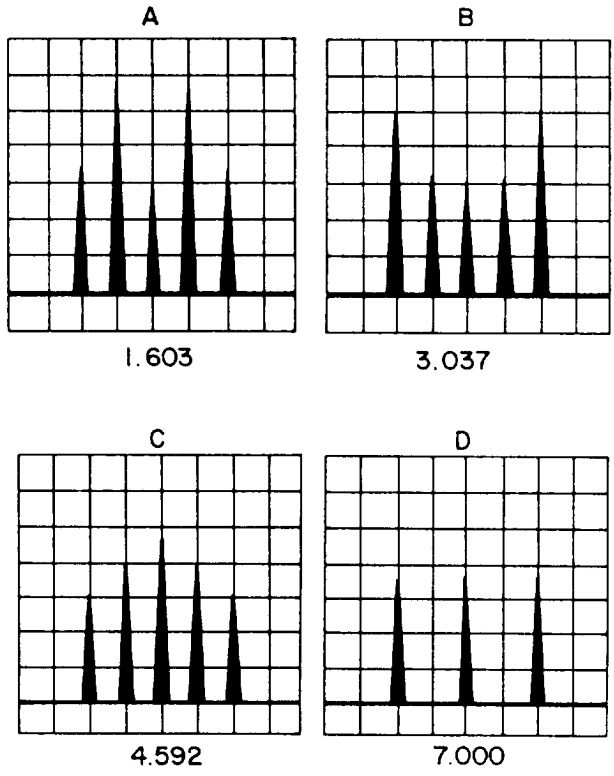


Figure 3-25.-FM spectral display at indices of 1.603, 3.037, 4.592, and 7.000, respectively (carrier @ 30 percent).

The spectrum analyzer will still display the carrier level as 30 percent, but by measuring sideband levels (two or three should prove sufficient), the correct index can be determined, as shown in figure 3-25. From this procedure, the precise frequency deviation can be readily determined.

IF AND RF AMPLIFIERS

There are three basic methods used to obtain a transmitter's operating frequency. One method mixes the output of various crystal oscillators of different frequencies. Another method involves multiplying the basic crystal frequency by certain factors. A third method employs frequency synthesis, whereby the basic oscillator's frequency is used to generate harmonics, which are then amplified and mixed. On occasion, more than one method is used. In each method, intermediate (IF) and output (RF) frequencies are developed that require amplification to attain the rated output power of the transmitter. Both IF and RF-type amplifiers can distort a modulated signal if the

amplifiers are not operated in the linear portion of their characteristic curve. Linear operation is not always the most efficient or desirable method of operating an amplifier. The output power requirements and the type of modulation used will be the determining factor in the design of IF and RF amplifiers.

IF Gain Measurement

When the gain factor of an IF signal is the main consideration and the IF stage is not amplifying an AM or SSB signal, nonlinear amplification is used for maximum efficiency. Since the IF is not amplitude-modulated, distortion products can be eliminated by installing fixed or tunable filters in the output stage on each amplifier. There are two prime considerations to keep in mind when testing this type of IF amplifier—the gain of the amplifier and the selectivity (response) of the filter, In general, the undesired, out-of-band signals should be reduced by more than 40 dB because IF gain will vary by each piece of equipment. The technical manual for each piece of equipment must be considered to determine what the IF gain factor should be. Figure 3-26 shows a basic setup for measuring IF gain.

FM Requirements

In FM transmissions, the basic frequency is modulated. It is this frequency that is mixed, translated, or multiplied to obtain the transmitter's output frequency. Since the distortion generated in nonlinear amplification primarily affects the amplitude of the carrier more than the frequency, FM is less susceptible than AM to the distortion created in nonlinear amplification. Therefore, the IF and RF amplifiers of an FM transmitter can be operated for maximum efficiency. The primary consideration in FM is that the IF filter's response be broad enough to pass the required frequency deviation of the FM signal with sufficient amplitude. The test setup for checking an FM transmitter's IF stages is similar to that for testing an AM transmitter's IF stages except that a sweep generator is used as a signal source instead of a CW signal generator.

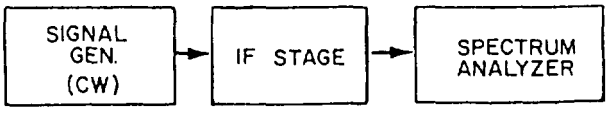


Figure 3-26.-IF gain and distortion measurement, test equipment arrangement.

AM/FM Considerations

Both AM and FM transmitters are rated in terms of average power. In AM transmitters, the power contained in the carrier does not change with an increase or decrease in the percent of modulation. Sideband power can increase the power output of a transmitter by as much as 50 percent at 100-percent modulation. In FM transmitters, the dispersement of power is averaged throughout the sidebands and the carrier; therefore, there is neither increase nor decrease in power as modulation becomes changed.

FIBER OPTICS

Learning Objective: *Describe fiber optics to include a basic system, advantages, and fiber construction. Describe light transmissions, fiber types, cables, and couplings.*

Fiber optics has revolutionized the telephone industry and will become the preferred norm of aviation and electronics technology. No longer will you see the cumbersome myriad of wires, connections, and cabling we have today. Weight will be reduced, and capabilities will be increased. As an AT you should see fiber optic technology in the very near future.

Fiber optics is not new. In the mid 1800s, William Wheeler patented a device for piping light from room to room. Alexander Graham Bell's photophone could reproduce voices through detection of the amount of light received from a modulated light source. In the last decade, a practical means of sending light has evolved—in the form of glass fibers.

BASIC SYSTEM

The principles of fiber optics follow the basic properties of light, and include refraction and reflection. Light traveling within a fiber obeys the laws of propagation. Fiber optics is the technique of sending data in the form of light through long, thin, flexible fibers of glass, plastic, or other transparent materials. A basic fiber optic system

(fig. 3-27) consists of a transmitter, a fiber medium, and a receiver. The transmitter converts electrical signals into current to drive a light source for injection into a fiber. The fiber or fibers guide(s) the light to a light detector that converts the light back into an electrical signal. The receiver is a low noise and large voltage gain receiver that provides further processing.

ADVANTAGES OF FIBER OPTIC SYSTEMS

There are many advantages of using fiber optics over systems in use today. Some of these advantages are shown below:

- Fiber optics can be used in flammable areas because light, not an electrical pulse, is the energy sent.
- Fiber optic systems are immune to radio frequency interference (RFI), electromotive interference (EMI), and noise caused by lightning and cross talk.
- Fiber optic systems are immune to electromagnetic pulse effects induced by nuclear explosions.
- Fiber optic aren't affected by moisture or temperature changes.
- Fiber optic systems are easy to repair.
- Fiber optic systems have very high data transmission rates.
- Fiber optic devices are small in size and lightweight.

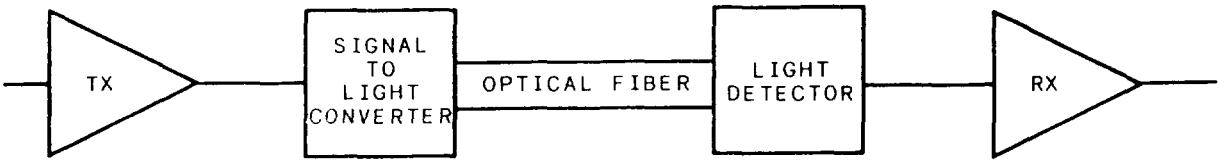


Figure 3-27.-Basic fiber optic system.

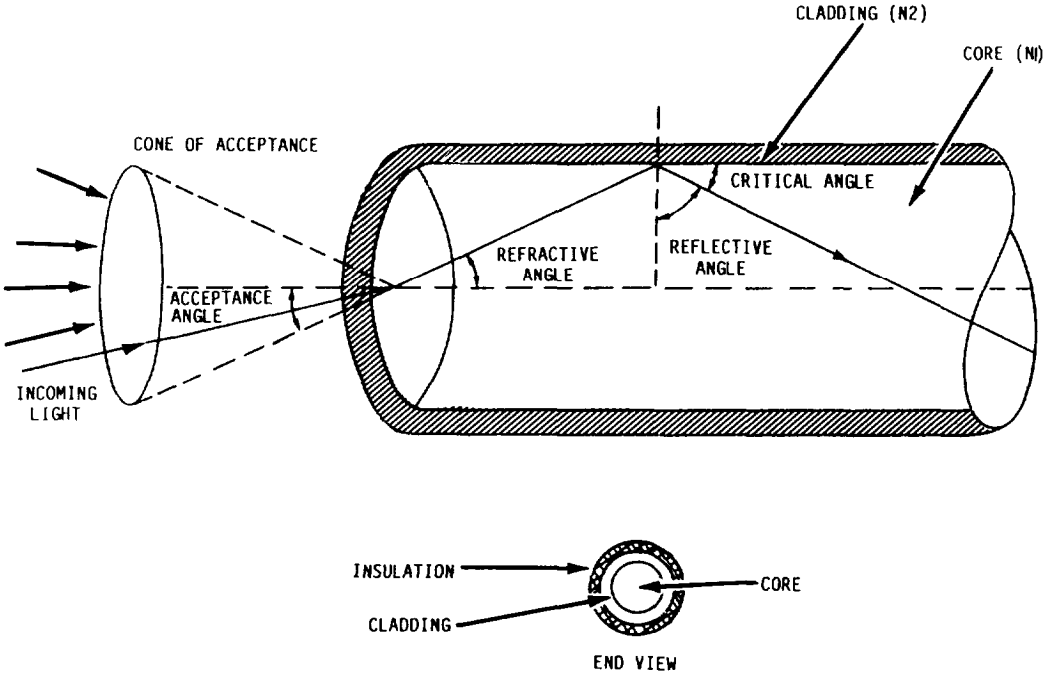


Figure 3-28.-Transmission of light in a fiber.

OPTICAL FIBER CONSTRUCTION

A typical optical fiber is a transparent, dielectric cylinder (core) enclosed within a second transparent dielectric cylinder (cladding). The core and cladding are enclosed by insulation (fig. 3-28). The dielectric cylinders consist of various optical glasses and plastics. The *cladding,* which has a relatively low index of refraction, encloses the core, which has a very high index of refraction. Cladding contains most of the transmitted light within the core. Its low index prevents light leakage and increases efficiency. The *insulation* protects a single fiber or several fibers from stress and the environment.

LIGHT TRANSMISSION

The light injected into a fiber travels in a series of reflections from wall to wall between the core and cladding. The reflections depend on the cone of acceptance and resulting angles of refraction and reflection propagation (fig. 3-29). The *cone of acceptance* is the area in front of the fiber that determines the angle of light waves it will accept. The *acceptance angle* is the half-angle of the cone of acceptance. The light enters the core and refracts to the interface of the core and cladding. The light reflects at the same angle of impact. The light, reflecting from wall to wall, continues at the same angle to the end of the fiber at the detector. As in the physics of light, the maximum *critical angle* is that angle that, when surpassed, won't reflect; in this case, it is lost in the cladding of the fiber. As long as the light wave is at a lesser angle than the maximum critical angle of the fiber (as determined by the function of the fibers' core and cladding indexes of refraction), light will travel to the receiver.

TYPES OF OPTICAL FIBERS

There are two types of optical fibers—step-index and graded-index. The step-index type has large differences in the core and cladding indexes of refraction. When held constant these differences cause light to reflect from the interface back through the core to its opposite wall. The graded-index type has a decreasing core refractive index as the radial distance from the core increases. This causes the light rays to continuously refocus as they travel down the fiber. Both types operate in either single mode or multi mode. Single mode accepts a specific wavelength, otherwise large attenuation will result. The multimode type operates over a range of wavelengths with minimum signal loss (fig. 3-29).

PROPERTIES OF OPTICAL CABLES

Optical cables are affected by many physical properties. Some of these are discussed in the following text.

Numerical Index

The numerical index of optical cables deals with the sine of the angle of acceptance. The

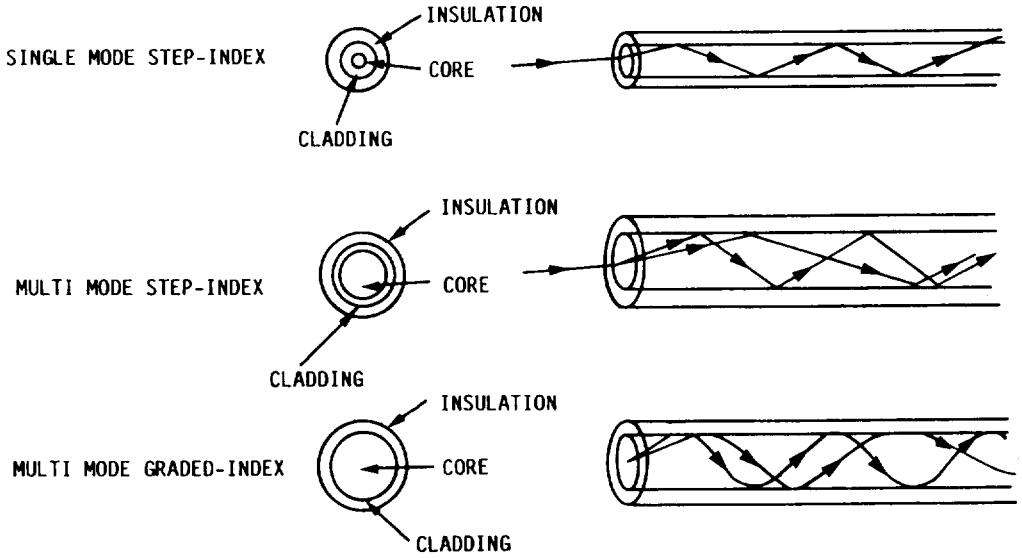


Figure 3-29.-Types of optical fibers.

numerical aperture (NA) (or numerical index) can be found using the formula shown below:

$$NA = \text{sine } i = n1^2 - n2^2$$

where i = acceptance angle, n1 = Core Index of Refraction, and n2 = Cladding Index of Refraction.

The acceptance angle is a measure of the numerical aperture (NA) or numerical index of a fiber. This lets the manufacturer select the proper fiber for the desired specific light waves and for optimum power coupling. NA is a measure of the light capture angle (half-acceptance angle). It describes the max core angle of light rays that will be reflected down the fiber by total reflection.

The refractive index (Index of Refraction) of a material is the ratio of the speed of light in a vacuum to the speed of light in the material. Review chapter 1 for more information on refraction if you don't understand this section. The higher the refractive index of a material, the lower the velocity of light through the material. Also, there will be more refraction or bending of the light when it enters the material.

If NA increases, angle i must have increased, and the fiber sees more light. NA can never be greater than 1.0; normal values are low (0.2 and 0.6).

Dispersion

Dispersion is the spreading or widening of light waves due to the refractive index of the material and the wavelength of the light traveling in the fiber. There are two types of dispersion—intermodal and intramodal.

Intermodal (multi-mode) dispersion. Intermodal dispersion is the propagation (travel) of rays of the same wavelength along different paths through the fiber. These wavelength rays arrive at the receiving end at different times.

Intramodal dispersion. Intramodal dispersion is due to variations of the index of refraction of the core and cladding.

Attenuation

Attenuation is the loss or reduction in amplitude of the energy transmitted. These losses are due to differences of refractive indexes and imperfections in fiber materials. Also, man-made scratches or dirt and light scattering within the fiber cause unwanted losses. Efforts to reduce these losses include the forming of the following standard parameters:

- Bandwidth parameters. Bandwidth parameters include attenuation curves, which provide all designers the ability to choose the best fiber. These parameters are plotted in decibels per kilometers (dB/km). They measure the efficiency of the fiber as a comparison of light transmission to light loss through a fiber.
- Rise time parameters. These parameters set speed requirements for operation.
- Fiber strength parameters. These parameters set tensile strength standards to help reduce flaws and microcracks in the fiber.

FIBER COUPLING

One important aspect of a fiber system is the connection between the fiber and the other parts. The coupling efficiency is the ratio of power accepted by the fiber to the power emitted by

$$\text{the source } \left(\frac{P_f}{P_s}\right).$$

Coupling efficiency increases with the square of the NA (numerical aperture) and decreases with source and fiber mismatches. Optical power coupled into the fiber is a function of the radiance of the source and the NA.

CHAPTER 4

NAVIGATION SYSTEMS

Chapter Objective: *Recognize components, operating principles, and features of airborne navigation systems used by the Navy.*

Airborne navigation systems encompass various equipment and instruments that are used to determine an aircraft's position, altitude, and heading. Your portion of this equipment includes automatic direction finder (ADF) systems, VHF omnidirectional range (VOR) systems, instrument landing systems (ILS), tactical air navigation (TACAN) systems, long range navigation (LORAN) systems, inertial navigation systems (INS), Doppler navigation systems, navigational computer systems, and electronic altimeter systems.

You will be tasked to operate and maintain the navigation equipment. It is beyond the scope of this manual to discuss all of the specific equipment; therefore, only representative equipment will be discussed. This chapter will provide you with the basic concepts, capabilities, and operating principles of the various types of navigation sets. Although there maybe newer and more sophisticated equipment in use than those depicted as examples in this chapter, keep in mind they all operate on the same basic principles.

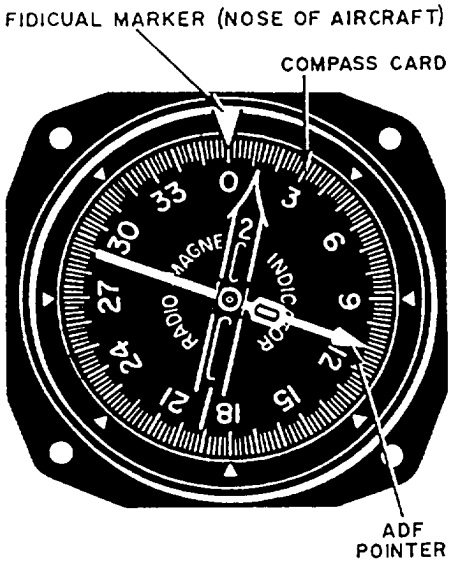


Figure 4-1.-Typical ADF indicator.

AUTOMATIC DIRECTION FINDER (ADF) SYSTEM

Learning Objective: *Identify the ADF system theory to include a typical UHF ADF system.*

An automatic direction finder is a radio receiver equipped with a directional antenna, which is used to determine the direction from which a radio signal is received. The antenna is motor driven by signals from the ADF receiver, and is connected by means of servomotors to an indicator pointer (needle). See figure 4-1. The pointer will point toward the radio station, giving the operator a relative bearing to the station with respect to aircraft heading. By using a map and knowing the location (city) of the radio station, an ADF operator can determine the aircraft's relative position from the city. By plotting a two-station fix (relative bearings), the operator can determine the aircraft's exact position on the map. See figure 4-2.

Most ADF units provide for manual operation of the directional antenna (referred to as the loop antenna) in addition to the automatic mode. The automatic mode

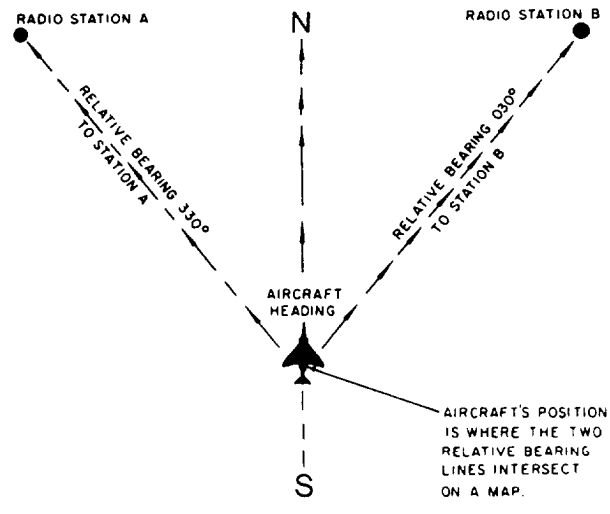
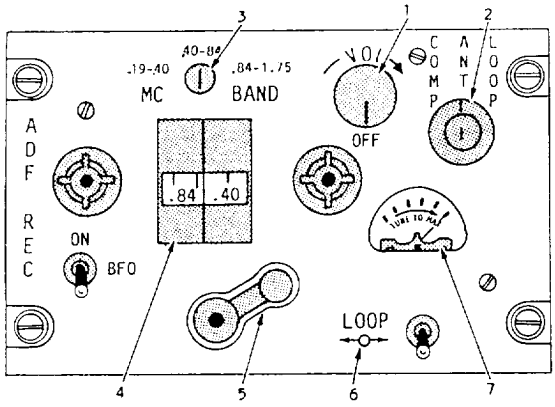


Figure 4-2.-Two-station ADF fix to determine aircraft's position.



CONTROL		FUNCTION
1.	OFF/ON Vol	Applies power to equipment and sets headset audio level
2.	Mode switch	Selects mode of operation of the equipment Comp–Automatic mode ANT–Selects sense antenna only. Pointer inoperative Loop–Selects loop antenna only
3.	Band select	Selects receiver frequency band of operation
4.	Frequency dial	Indicates frequency to which the receiver is tuned
5.	Tuning knob	Tunes the receiver to desired frequency as indicated by frequency dial
6.	Loop switch	Manually drives the loop antenna in the loop mode of operation
7.	Signal strength meter	Indicates the amplitude of the received signal

Figure 4-3.-Typical ADF control panel.

involves the use of closed-loop control circuits to operate the antenna. While in manual mode, the operator drives the antenna by use of a control switch located on the ADF control panel. See figure 4-3. When a signal is received, the indicator pointer will point to the radio station, just as it would in automatic mode.

BASIC PRINCIPLES

When a conductor is cut by magnetic lines of force, or lines of flux, a voltage is induced in the conductor. In order to cut lines of flux, the conductor must be perpendicular or must have a component that is perpendicular to the lines of flux. The relative motion between flux and conductor must have a comment in

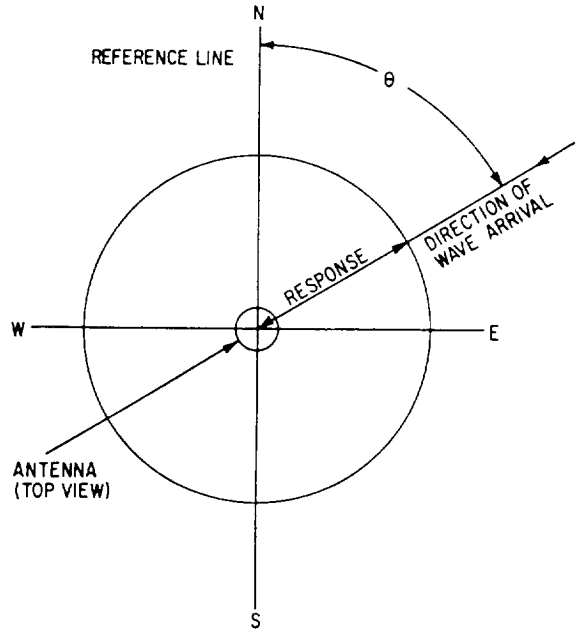


Figure 4-4.-Polar response of a monopole (sense) antenna.

a direction that is perpendicular to both the lines of flux and the conductor.

A vertically polarized wave has a vertical electric (E) field and a horizontal magnetic (H) field. The wave induces voltage in vertical conductors only. A vertical wire, or monopole, is the simplest type of antenna. When a vertically polarized radio wave induces voltage in a monopole, the induced voltage is in phase with the incident wave and is the same for all horizontal angles of incidence. See figure 4-4. This similarity of response pattern, in all directions simultaneously, suggests the name omnidirectional (omni means "all") for this antenna.

The response of a loop antenna is different from that of a vertical monopole. The vertical monopole antenna is also known as a sense antenna. A rectangular, single-turn loop with dimensions that are small compared to the wavelength of an incident radiation field is shown in figure 4-5, view A. As the loop is rotated about the XX′ axis, the angle between the plane of the loop and the direction of propagation of the wave is changed.

If the loop is placed in the radiation field like the one shown in figure 4-5, view B, the H vectors of the field cut the sides AB and CD at slightly different times because the wave travels at a finite speed. At any instant the voltage induced in arm AB is slightly different from

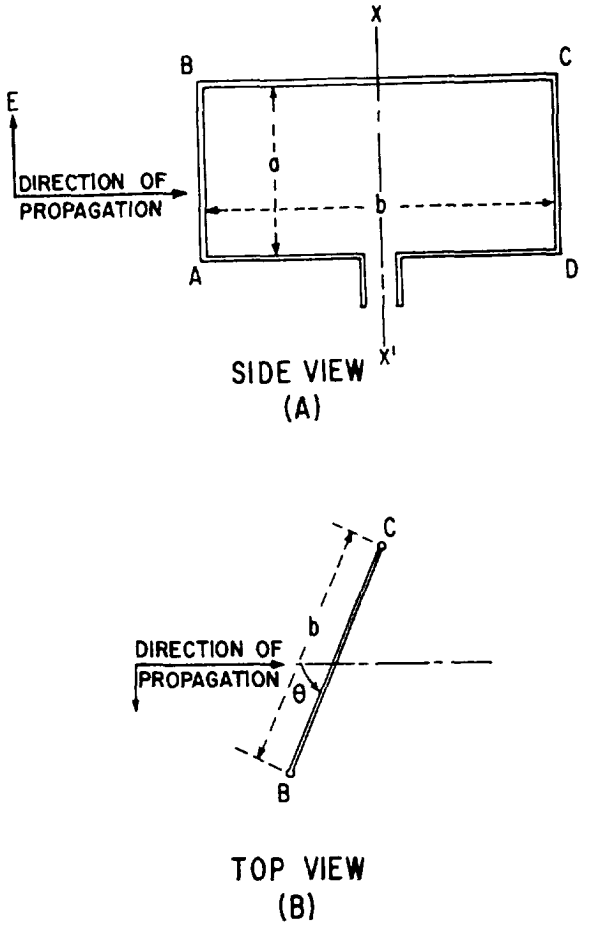


Figure 4-5.-Loop antenna in a radiation field.

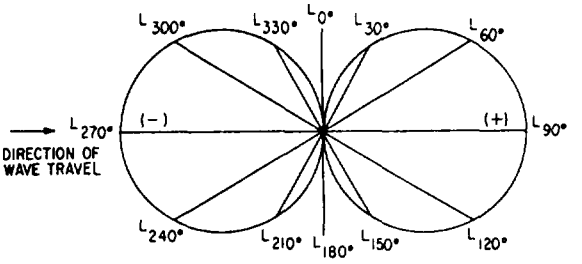


Figure 4-6.-Polar response of a loop antenna.

the voltage induced in arm CD. The arms BC and AD are not affected by the H lines of a wave polarized at right angles to them. The arms do not contribute to the induced voltage in the loop because the horizontal members are parallel to the H lines.

If the loop is turned so that its face is perpendicular to the direction of arrival of the wave, the sides AB and CD are cut by the incoming wave, that is, $\theta = 90°$. The sides AB and CD are cut by the H vector at the same instant. The voltages induced in arms AB and CD are then the same magnitude and phase. They neutralize each other, so that no current flows in the antenna loop.

Since the magnetic field of the radio wave alternates at the frequency of the wave, the instantaneous flux density at any point along the path of arrival varies sinusoidally. Sinusoidal voltage is voltage that varies with the sine function of the phase angle. Thus, the voltage induced into arms AB and CD are sinusoidal voltages with a phase difference θ. The total loop voltage is the sinusoidal voltage, which represents the integration of the sums of all the instantaneous voltages induced into the two arms. It may be shown mathematically that this resultant voltage is proportional to the cosine of θ. The pattern of the response is similar to the figure eight. See figure 4-6. All medium-frequency, direction-finding equipments obtain bearings by using the response as shown in this figure.

The directional characteristic of the loop antenna is called a cosine, or figure-of-eight pattern. When the loop is oriented so that the received signal is maximum, a small change in orientation produces a small change in signal. When the orientation of the loop is such that the received signal is minimum, a large change occurs in output voltage. Furthermore, there is a reversal in phase of the signal as the loop passes through a null point. For these reasons, the null points, rather than the maximum-response points, are used in radio direction finding to obtain a line bearing or line of arrival of a radio wave.

As there are two null positions 180 degrees apart, the loop can give a line of bearing (the actual bearing or its reciprocal). The absolute direction of the transmitter from the direction tinder is not determined directly from the loop antenna. The determination of absolute direction, or sense, is obtained by adding the output of a vertical sense antenna (monopole antenna) to that of the loop antenna. When the two antennas are properly connected, the combined response is not ambiguous.

SIGNAL COMPARISON

The figure-of-eight pattern of a loop has two null positions for one incident radio wave. See figure 4-6. If the outputs of a loop and a sense antenna are combined in phase, the response of the two antennas is the

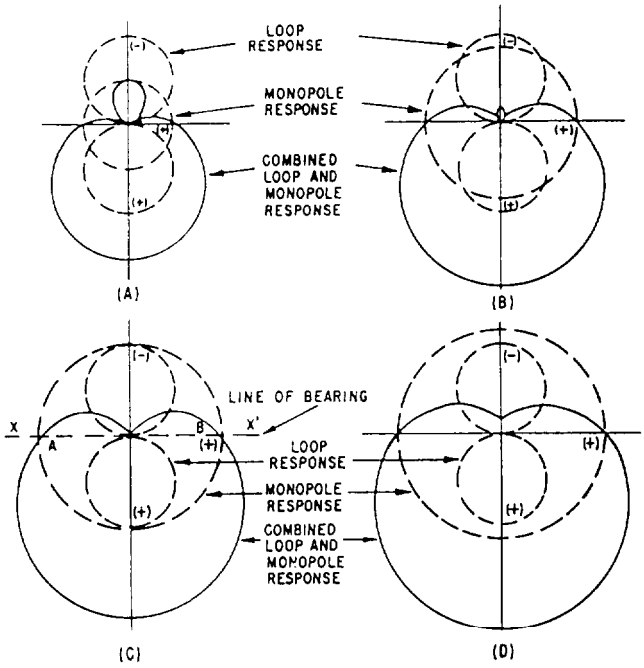


Figure 4-7.-Response of the combined sense and loop antenna.

algebraic sum of their individual diagrams. See figure 4-7. This figure shows four possible responses caused by differences in the relative amplitudes of the sense and loop outputs. The desired response, shown in view (C), has one sharp null.

The output of the sense antenna is independent of the horizontal direction of arrival of the wave, so it may be considered to have a positive polarity. The phase of the loop voltage changes as the loop passes through a null. One-half of the figure-of-eight pattern has a positive polarity and the other half a negative polarity. The addition of the loop and sense curves gives the responses shown. The shape of the resultant curve is called a cardioid because of its similarity to a valentine heart.

The output of the sense antenna is in phase with the radio wave. The output of the loop antenna, however, is 90 degrees out of phase with the radio wave. This means that the loop output voltage is maximum when the sense output is zero, and vice versa. The cardioid pattern produced by the combined loop and sense antenna can also be produced by a rhombic antenna. This type of antenna is used without a sense element.

DIRECTION FINDER SET AN/ARD-13

Direction Finder Set AN/ARD-13 is a low-frequency radio navigation device that operates at frequencies between 90 kHz and 1800 kHz. It is capable of receiving both AM and CW transmissions within its operating range. We will use this set as a representative direction finder set.

The direction finder has three modes of operation that are selected remotely at the control unit. In the ANT mode, the RF input to the receiver is from the sense antenna. The direction finder in this mode operates as a nondirectional, low-frequency receiver. The RF input in the loop mode is from the loop antenna. The loop mode may be used for manual direction finding by rotating the loop for an audio output null or a tuning meter null. A 180-degree ambiguity indirection is possible in the loop mode, since the loop antenna pattern has two nulls 180 degrees apart. The direction from which signals are best received can be chosen manually from the control box by positioning the loop. In the ADF mode, the loop antenna's signal determines whether the loop is pointed to the left or to the right of the signal source. The receiver commands clockwise rotation of the loop if the loop axis is pointed to the left of the signal source. Counterclockwise rotation is commanded if the loop axis is to the right of the signal source. Rotation ceases when the loop axis is pointed directly at the signal source.

The position of the loop is continuously transmitted to the bearing indicator. The bearing indicator reads the bearing to the station in the ADF mode because the loop is kept pointed directly at the station. The bearing indicator combines the bearing information from the direction finder with navigation data received from other equipment. Audio signals in all three modes are supplied to the intercommunication system (ICS) in the aircraft. The audio level to the ICS is varied manually from the control box.

Theory of Operation

To understand the theory of operation of a direction finder, refer to the simplified block diagram of figure 4-8.

ANT (ANTENNA) MODE.– Radio frequency signals from the sense antenna are coupled to the RF amplifier through the impedance-only signal input to the RF amplifier in the ANT mode. The loop amplifier and the balanced modulator are disabled. The oscillator and mixer convert the output of the RF amplifier to 455.7 Hz, the intermediate frequency of the receiver. The oscillator is tuned from the control box by the tuning servo. One of two mechanical fibers passes the desired signal and attenuates the undesired signals. The broad filter provides selectivity of 3.1 kHz, and the sharp filter

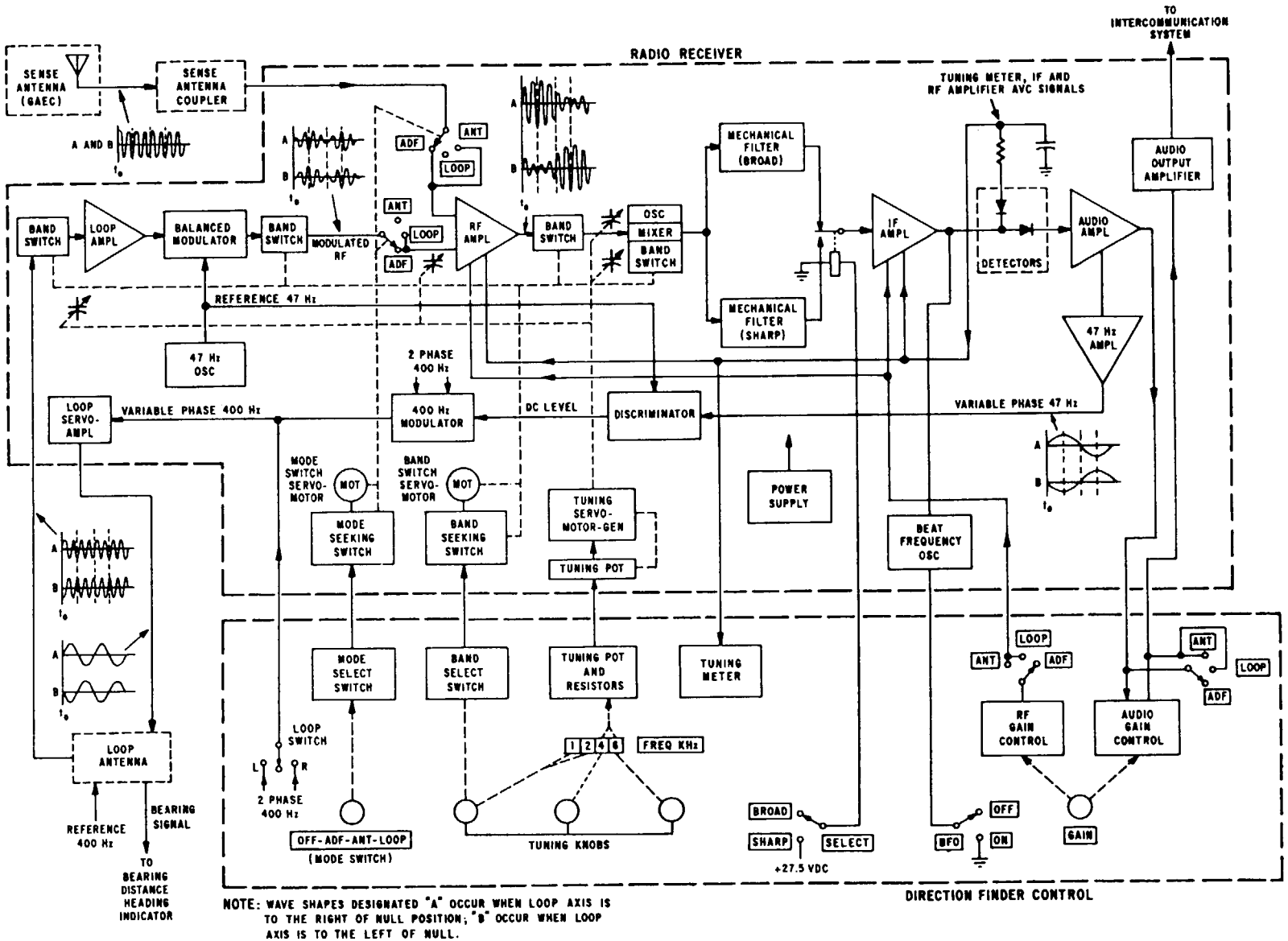


Figure 4-8.–AN/ARD-13 simplified block diagram.

provides receiver selectivity of 1.5 kHz. The narrow band output of the mechanical filters is amplified by the IF amplifier, and it is applied to the detectors. The output of one detector is used as the automatic volume control (AVC) signal to limit the gain of the RF and IF amplifiers. The AVC signal is also applied to the tuning meter. The output of the other detector is applied to the audio amplifier. The audio gain control is bypassed in the ANT mode, and the receiver gain is manually controlled by using the RF gain control. The audio amplifier increases the output of the detector to the level required by the ICS in the aircraft.

LOOP MODE.– Rotation of the loop antenna is controlled by the loop switch on the control box. The loop switch applies either of two phases of 400-Hz ac from the receiver to the loop. The operator uses the loop switch to drive the loop antenna to the position of minimum reception of the received signal. This position of minimum reception, which occurs when the loop antenna is pointed directly at the signal source, is called the null position. The angle of the transmitting station, with respect to aircraft heading, can then be read accurately on the bearing indicator. The RF output of the loop antenna is applied to the balanced modulator through the loop amplifier. The balanced modulator is unbalanced during loop operation, and it couples the output of the loop amplifier to the RF amplifier. Signals from the balanced modulator are the only input to the RF amplifier in the loop mode. The input to the RF amplifier is the same as in the antenna mode of operation.

ADF MODE.– The RF output of the loop antenna has either of two phases relative to signals from the sense antenna. Phase A, as shown in figure 4-8, occurs when the loop antenna is to the right of the null position. Phase B occurs when the loop is to the left of the null. The loop antenna has no output when in the null position. Either output phase of the loop antenna is modulated by a 47-Hz signal in the balanced modulator stage. The output of the balanced modulator is added to signals from the sense antenna by the RF amplifier. Note that there is a 180-degree difference in phase between the envelope of A, present when the loop antenna is to the right of the null, and envelope B, present when the loop antenna is to the left of the null. The output of the IF amplifier is amplified and detected. The output is applied to the audio amplifier. The output of the audio amplifier is processed and applied to the 47-Hz amplifier.

The amplified 47-Hz component of the output of the audio amplifier is applied to the discriminator. The discriminator compares the phase of the 47-Hz signal from the 47-Hz amplifier with the reference phase from the 47-Hz oscillator. If the two are in phase, the discriminator applies a positive de level to the 400-Hz modulator. If the two are out of phase, the discriminator applies a negative dc level to the 400-Hz modulator.

The 400-Hz modulator applies either a phase A or phase B 400-Hz signal to the loop servo amplifier, which is connected to one winding of the loop antenna drive motor. Phase A applies if the dc level from the discriminator is positive. The signal that drives the loop antenna motor will always cause the loop antenna to rotate toward the null position.

The bearing indicator, electromechanically coupled to the loop antenna by synchros, reads the position of the loop, and thus the direction to the signal source.

In the ADF mode, the RF gain control is inoperative. The audio gain control is used to vary the audio output level of the receiver.

ADF Bearing Limitations

Various factors contribute to inaccuracy in radio bearings. As a maintenance technician, you should keep these in mind when analyzing reported ADF discrepancies. Some of the more important factors are discussed in the following text.

NIGHT EFFECT.– This is caused by the reflection of sky waves from the ionosphere. Night effect is most noticeable for about 1 hour, around sunrise and sunset. At these times, the height of sky waves vary in their intensity and range. This fluctuation interferes with the reception of the ground wave. Since the operation of the ADF depends on the reception of ground waves, the loop antenna tends to hunt, causing the bearing needle to fluctuate.

ELECTRICAL DISTURBANCE.– Radio waves are distorted by electrical storms. This results in extremely erratic hunting of the loop antenna and bearing needle. The antenna tends to home in the direction of the electrical storm.

PRECIPITATION STATIC.– An aircraft may accumulate a static charge when moving through the air, especially air that is laden with particles (dust, ice crystals, etc.). These particles may already have a charge on them, or create one through frictional contact with the aircraft surfaces. These charges tend to discharge from surface to surface or off into the air. In so doing, these changes intermittently cause interference with the ADF equipment.

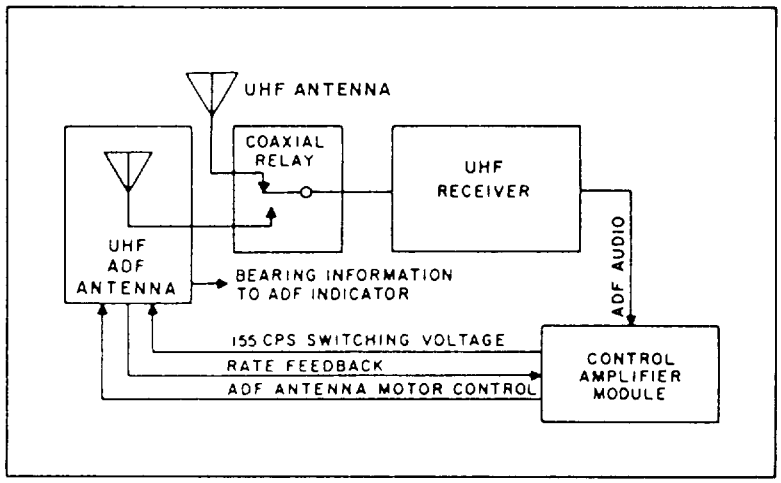


Figure 4-9.–Typical UHF ADF system block diagram.

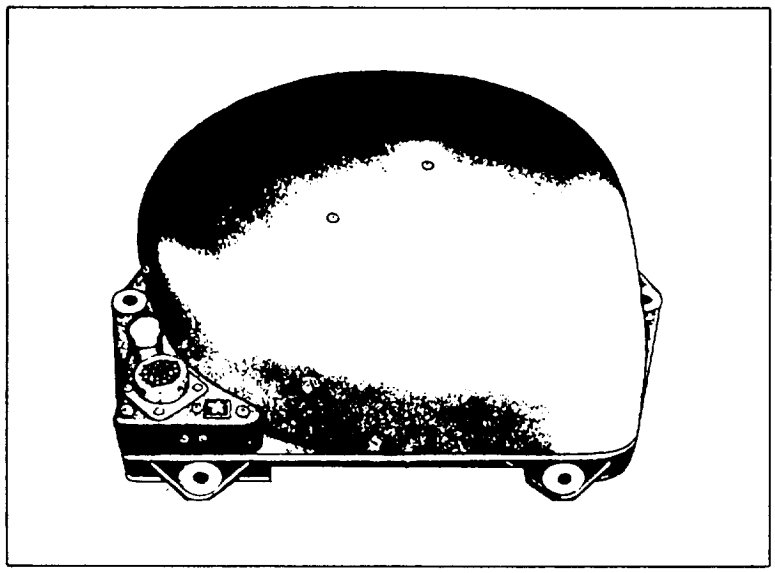


Figure 4-10.-Antenna AS-909/ARA-48.

QUADRANTAL ERROR.– When incoming radio waves strike an aircraft's surface, a number of reradiated fields are created around the metallic portions of the aircraft. These fields bend the radio waves prior to reception by the loop antenna. This error, known as quadrantal error, is maximum when incoming radio signals must cross the wings or stabilizer surfaces before striking the loop antenna. Quadrantal error is usually adjusted for by compensating circuits installed in the loop antenna unit. Whenever anew loop antenna unit is installed in an aircraft, it must be calibrated in accordance with the maintenance manual for the equipment.

TYPICAL UHF ADF SYSTEM

The total typical UHF ADF system is comprised of an ADF antenna, a coaxial relay (to select either normal UHF antenna or the ADF antenna), a control amplifier (to drive the ADF antenna), a UHF transceiver that is wired for ADF, and a bearing indicator. Figure 4-9 is a block diagram of a typical UHF ADF system. Note the receiver in the figure could be an ARC-159, since it is wired for ADF operation. If you place the mode selector switch to the ADF position, it will actuate the coaxial relay shown in figure 4-9. The control amplifier module contains the circuitry to steer the ADF antenna, much the same as the ARD-13 drove its loop antenna.

The heart of a UHF ADF system is the directional antenna. Because of the design of the typical antenna, no sense antenna is required as with the loop antenna. A typical UHF ADF antenna is the AS-909/ARA-48. See figure 4-10. It is a flush-mounted type antenna. It receives signals in the range of 225.0 to 400.0 MHz.

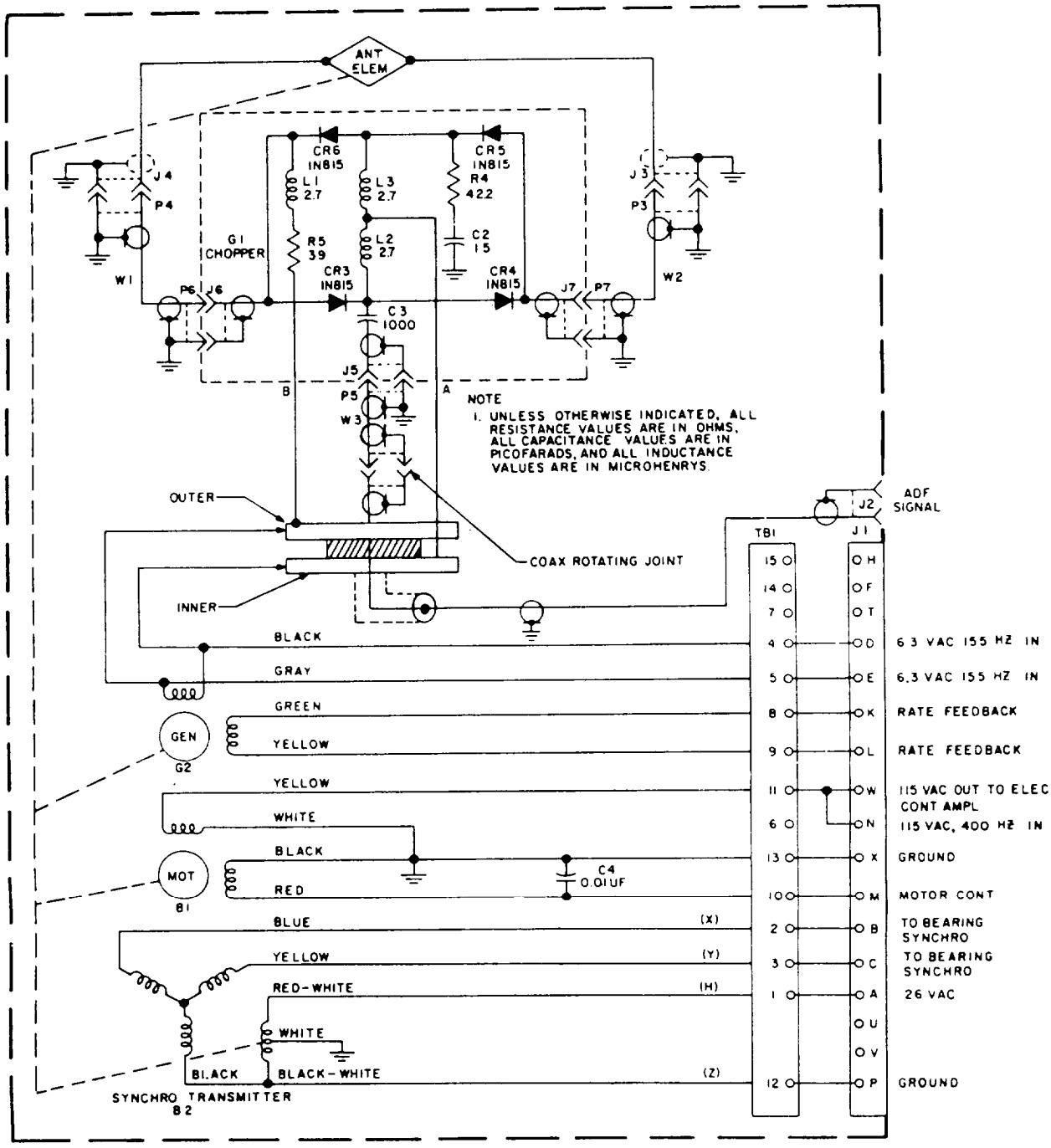


Figure 4-11.-Antenna AS-909/ARA-48 schematic diagram.

The antenna unit consists of a directional receiving element, an antenna drive motor, a rate generator, a lobing switch, and the associated gear assembly necessary to mechanically link the items. See figure 4-11. The ante ma is a cavity-backed complementary slot radiator that is formed by the position of a rhombic-shaped metal plate. The antenna element is terminated alternately at either end by use of the antenna lobing switch G1. This action allows the antenna field, which is a cardioid, to be reversed 180 degrees, 155 times each second, by a signal from the control amplifier module. It is this switching of the field that prevents ambiguous 180-degree readings of the received signal and eliminates the need of a sense antenna.

The switching of the cardioid antenna pattern causes the received RF signal to be square-wave modulated. The degree of modulation is received by the antenna element. See figure 4-12. The AS909/ARA-48 antenna

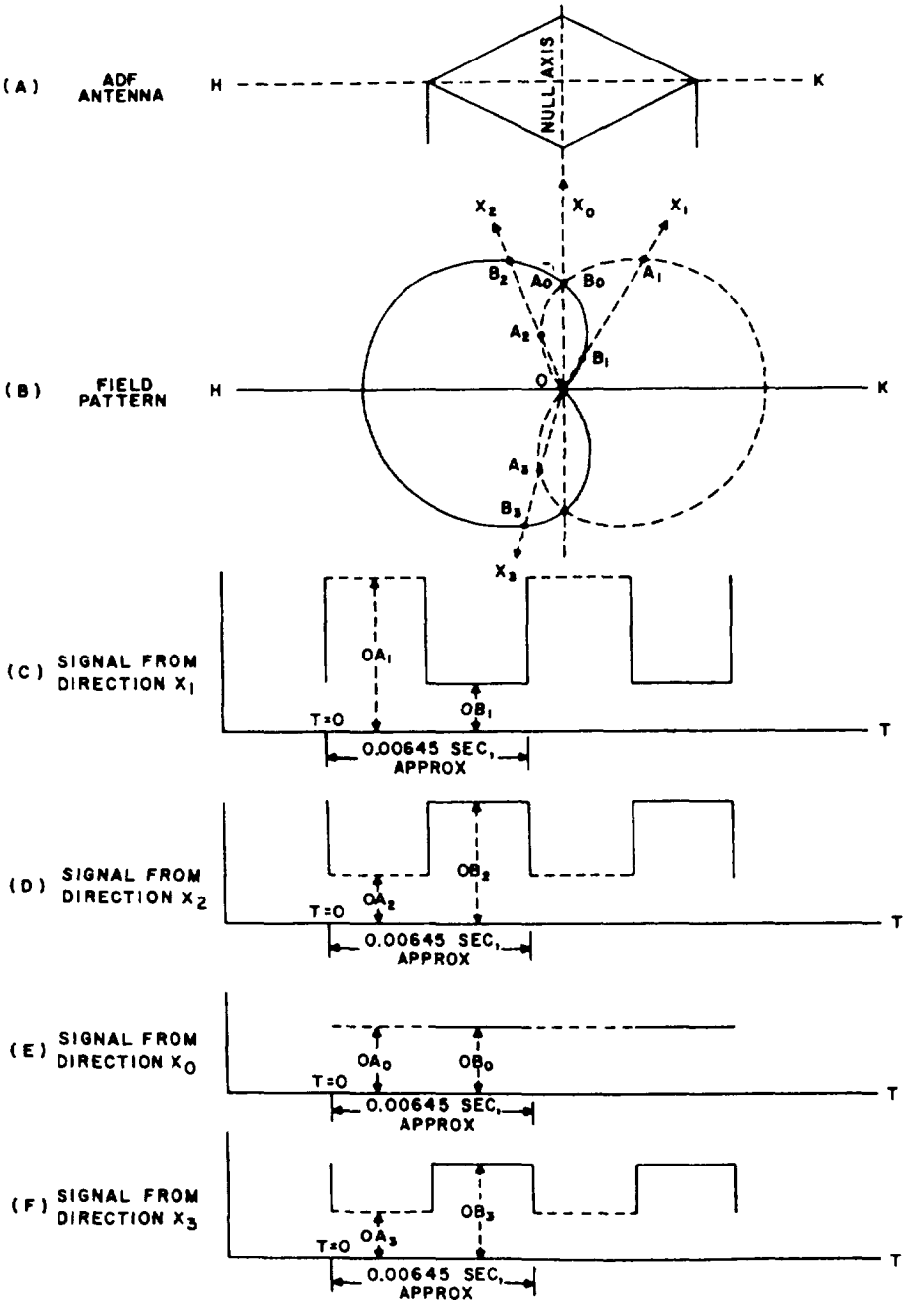


Figure 4-12.-Receiver input levels for various ADF signal paths.

develops the modulation as follows: Assume that a signal is received from the direction OX_1, figure 4-12, view (B). The resulting modulated input to the control amplifier module (fig. 4-9), after being detected in the UHF receiver, is of the form shown in view (C) of figure 4-12. A motor control voltage proportional to the difference between OA_1 and OB_1 is then applied from the control amplifier module to the antenna drive motor B1 (fig. 4-11). B1 drives the antenna element toward the null position indicated by OX_0, figure 4-12, view (B). When a signal is received along this null axis, the difference in modulation resulting from the switching of the antenna field pattern is zero. Under these conditions, the motor control voltage applied to B1 is zero, and the antenna ceases to rotate.

Synchro transmitter B2 (fig. 4-11) transmits antenna position information to the bearing indicator needle.

A rate feedback voltage proportional to the speed of the antenna rotation is developed by the rate generator G2 (fig. 4-11) and fed to the control amplifier module,

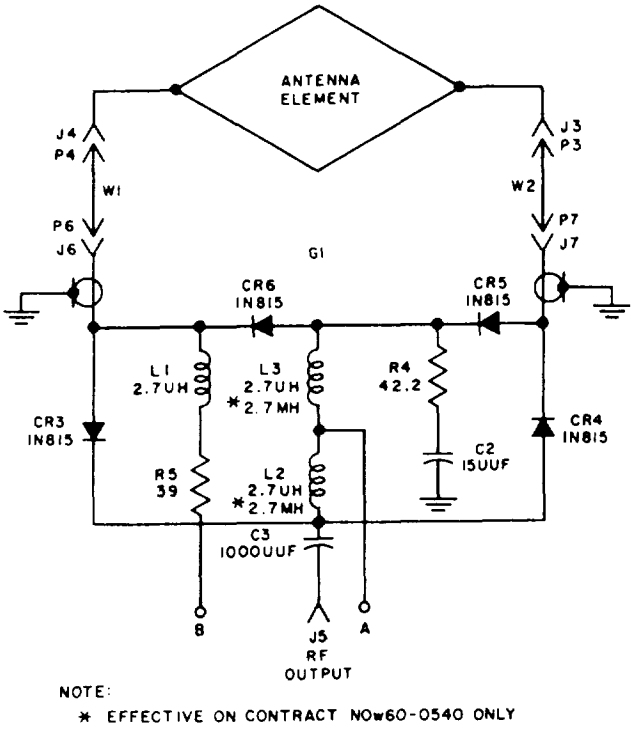


Figure 4-13.–Electronic lobing switch simplified schematic diagram.

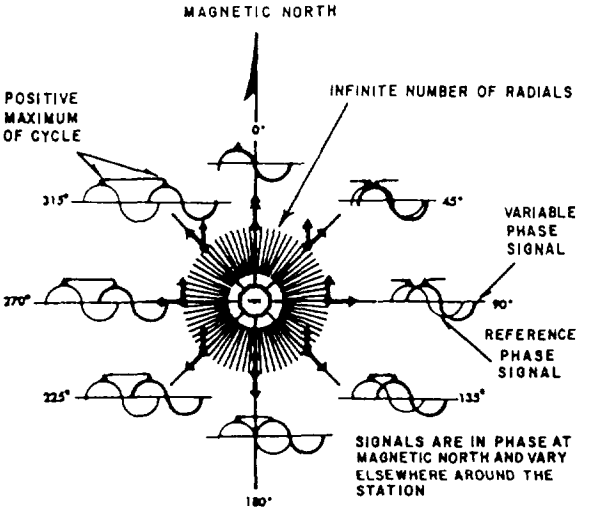


Figure 4-14.-VOR transmitter.

where it combines with the input from the UHF receiver to prevent oversteering of the antenna and similar problems.

The electronic lobing switch G1 (fig. 4- 13) uses four diodes to perform the switching function. In the simplified schematic diagram shown in figure 4-13, the ADF antenna element is represented as a diamond-shaped plate, the ends of which are connected to J3 and J4, respectively. To examine the operation, assume a 6.3-volt square-wave voltage at 155 Hz incoming from the control amplifier module and applied between points A and B. A positive voltage at point B places forward bias on CR3 and CR5 (through the antenna element), causing both diodes to conduct and appear as a low-value RF impedance. The same potential will bias CR6 and CR4 in the reverse direction, so they are nonconducting. Under these conditions, the J3 end of the antenna element is connected through the resistance of CR5 to the terminating network which consists of R4 and C2. The RF signal is coupled from the J4 end of the antenna element, through CR3 to J5. When the input square wave drives point B negative with respect to point A, CR3 and CR5 are biased off, and CR6 and CR4 are conducting due to the forward bias. The J4 end of the antenna element is then connected to the terminating network R4 and C2, while the signal is passed from J3 through CR4 and C3 to J5. Inductors L1, L2, and L3 isolate the RF energy from the 155-Hz control amplifier circuits. Capacitor C2 offers a low impedance to the high-frequency RF signals effectively placing the end of R4 at RF ground, while isolating the 155-Hz circuitry by the relatively high reactance at 155 Hz.

Although the typical UHF ADF antenna is better than the standard LOOP/SENSE antenna, it is still vulnerable to some of the factors that could cause bearing inaccuracies. Radio signals experience minimal “night effect” errors while operating in the UHF frequency spectrum, but they are subject to quadrantal and electrical disturbance errors.

VHF OMNIDIRECTIONAL RANGE (VOR) SYSTEM

Learning Objective: *Identify VOR* basic *operating principles as well as the operation of a typical VOR system.*

A VOR facility is a radio range station whose transmitting radiation patterns produce directional courses on “tracks” by having special characteristics in its emissions, recognizable as bearing information. These courses or “tracks” remain stationary with respect to the surface of the earth. The operation of a VOR bearing function may be compared to that of an airport beacon light. If the beacon light, rotating at a known speed, blinks each time it sweeps past magnetic north, and the time from that blink until the beam sweeps past an aircraft is measured, the magnetic bearing from the beacon can be determined. For example, if the beam

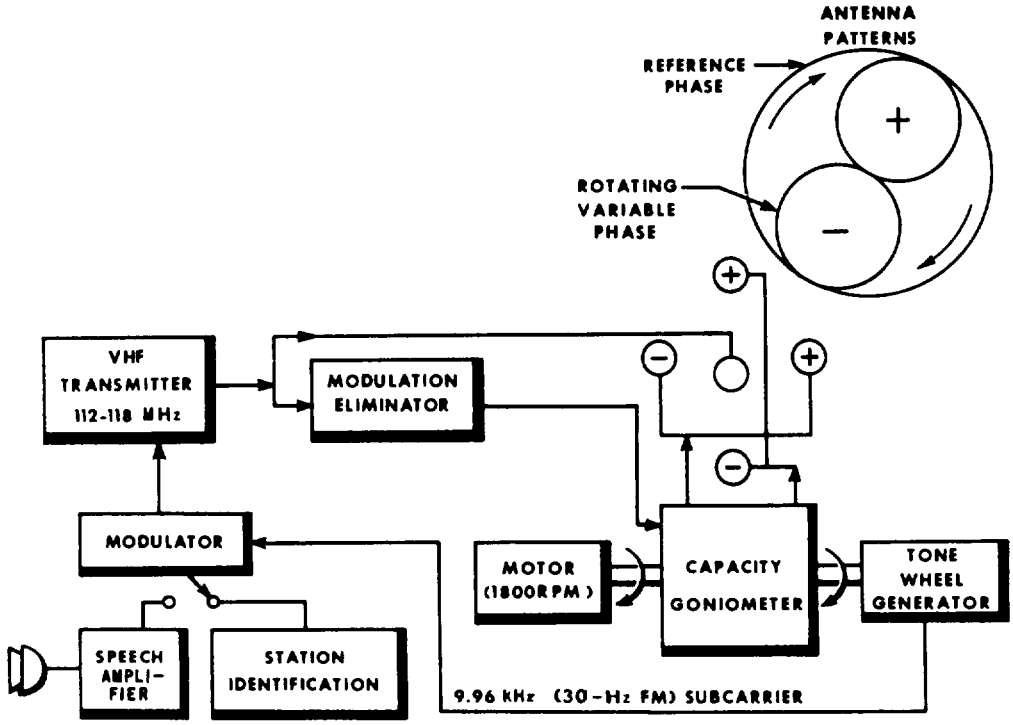


Figure 4-15.-Typical VOR transmitting station.

revolves at 1 degree per second and 120 seconds are counted between the reference blink (magnetic north) and the time the beam strikes the aircraft, the aircraft is on the 120-degree radial (track) from the beacon.

TRANSMISSION PRINCIPLE

The transmission principle of the VOR station is based on the creation of a phase difference between two simultaneously transmitted RF signals in the frequency band of 108.00 to 117.95 MHz. Magnetic north is used as the base for measuring the phase relationship. See figure 4-14. One of the two RF signals transmitted is nondirectional and has a constant phase throughout 360 degrees of azimuth. It is called the reference phase signal. This signal is transmitted on a 9.96-kHz subcarrier, frequency modulated at 30 Hz, and is applied to the center loop. The second signal is a rotating signal with a speed of 1800 RPM (30 RPS). It is called the variable phase signal.

As shown in figure 4-15, the VOR transmitter is modulated both by a 9.96-kHz subcarrier and by an additional component, either of voice or the station identification code characters. The subcarrier is frequency modulated at 30 Hz, and is generated by a notched tone wheel rotating in a magnetic field. The purpose of the subcarrier is to provide a means for separating the 30-Hz tone of the reference phase from the variable tone of the same frequency.

The modulated output of the transmitter is applied to both a modulation eliminator and to the center loop antenna of a five-element array. There it is radiated to form the reference phase signal. The modulation eliminator is a clipper that removes the amplitude modulation from the carrier. The unmodulated output is fed to a capacity goniometer, which serves as a mechanical sideband generator. The goniometer is a motor-driven, double capacitor in which one set of stator plates is displaced 90 degrees from the other set. The rotor plates to which the RF signal is applied are driven at 1800 RPM (30 RPS).

Two outputs are derived, one from each set of stator plates. These two signals contain modulation components (30 Hz) that differ in phase by 90 degrees because of the capacitor plate relationship. One output is fed to one pair of diagonally opposite loop antennas, and the other is fed to the remaining pair of loops in the square array. Each pair of corner antennas produces a figure-of-eight radiation pattern. These two patterns are displaced from each other by 90 degrees in both space and time phase. The resultant pattern is the sum of the two crossed figure-of-eight patterns, and consists of the rotating field. The transmitter is designed so that the 30-Hz, frequency-modulated component reaches its

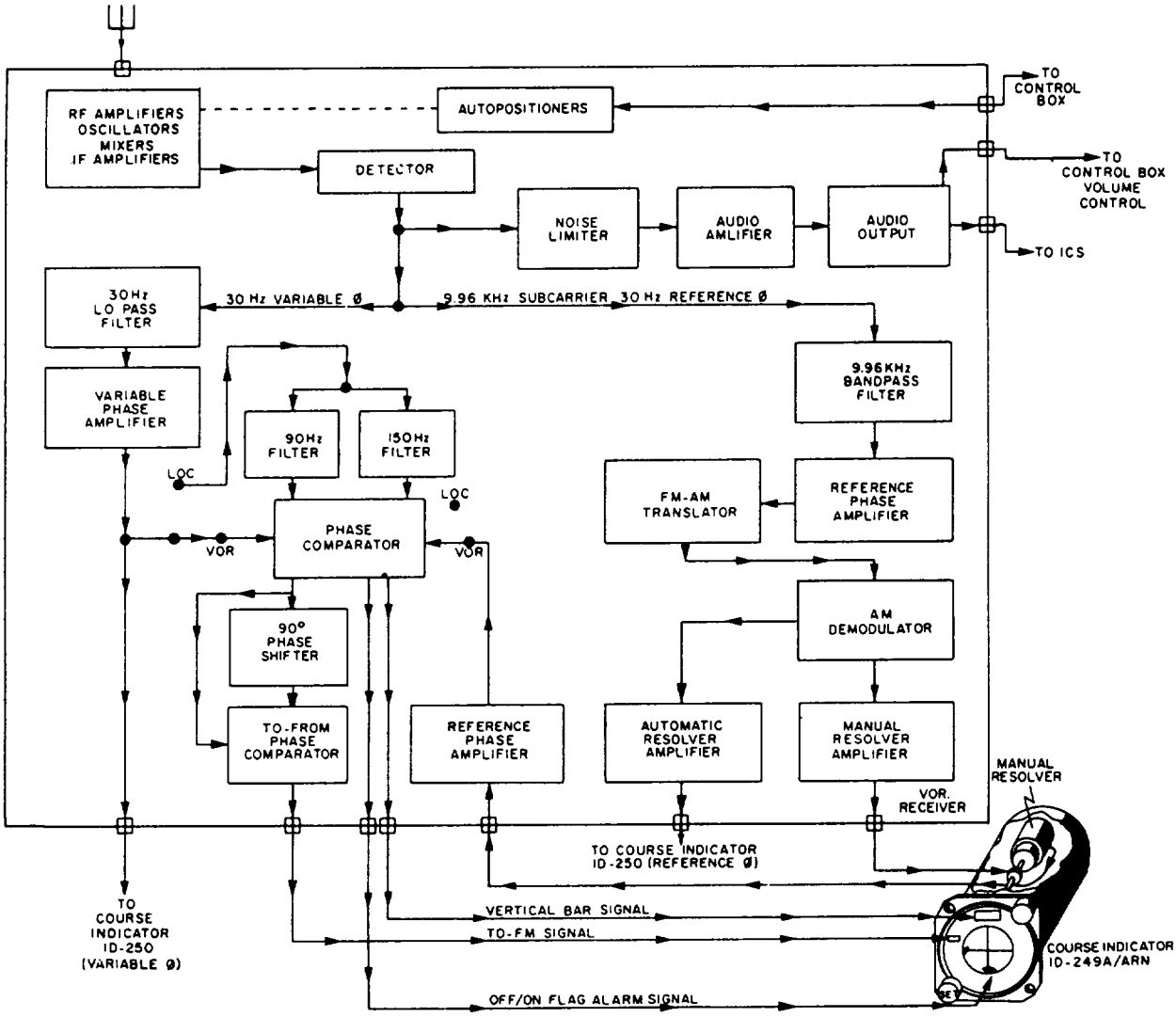


Figure 4-16.-Typical airborne VOR receiver simplified block diagram with ID-249.

positive maximum at the same time that the rotating pattern maximum passes magnetic north (in phase). As a result of the rotation of the variable phase pattern, the signal induced in the airborne VOR receiver is amplitude modulated with 30-Hz variations. The receiver develops this tone and compares it with the reference signal. The amount of phase difference between the two signals depends on the location of the aircraft with respect to the transmitting station. Although the VOR provides an infinite number of courses from the station, for simplicity it is referred to as providing 360 courses 1 degree apart. These courses are called radials. Any radial may be selected and flown, or may be used to obtain a line of position at any time.

Table 4-1.-VHF Typical Frequency Bands and Their Uses

Frequency	Use
108.00 — 117.95 MHz	VHF Omni Range (VOR)
108.1 — 111.9 MHz (odd tenths)	ILS Localizer
116.00 — 151.95 MHz	VHF Communication

TYPICAL AIRCRAFT VOR SYSTEM

Atypical aircraft VOR (omni) system consists of a receiver, control box, course deviation indicator

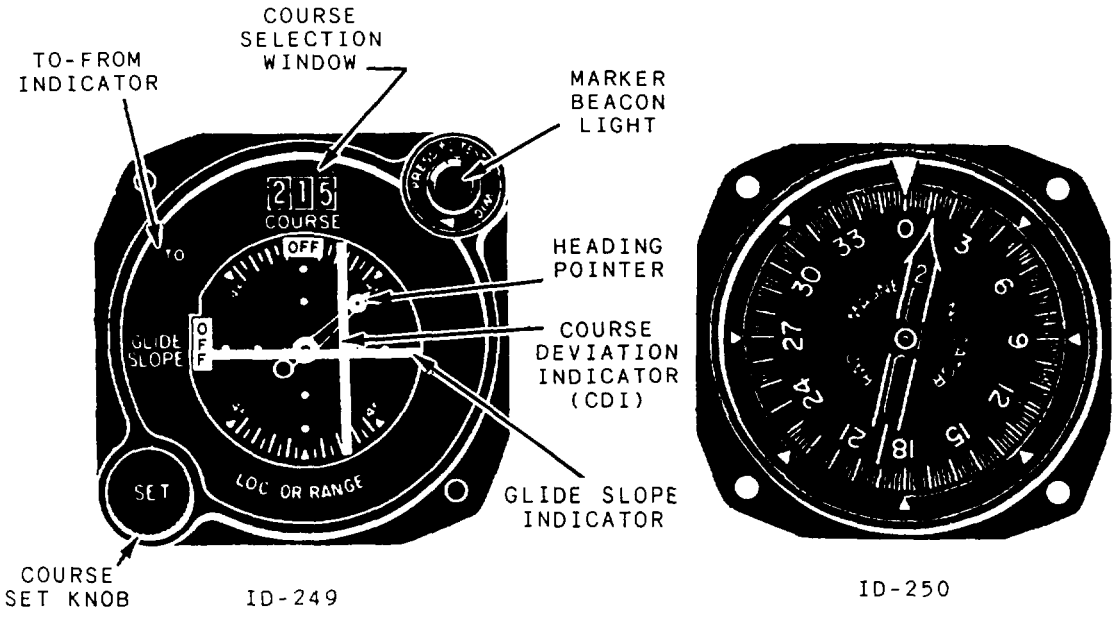


Figure 4-17.-Typical VOR Indicator.

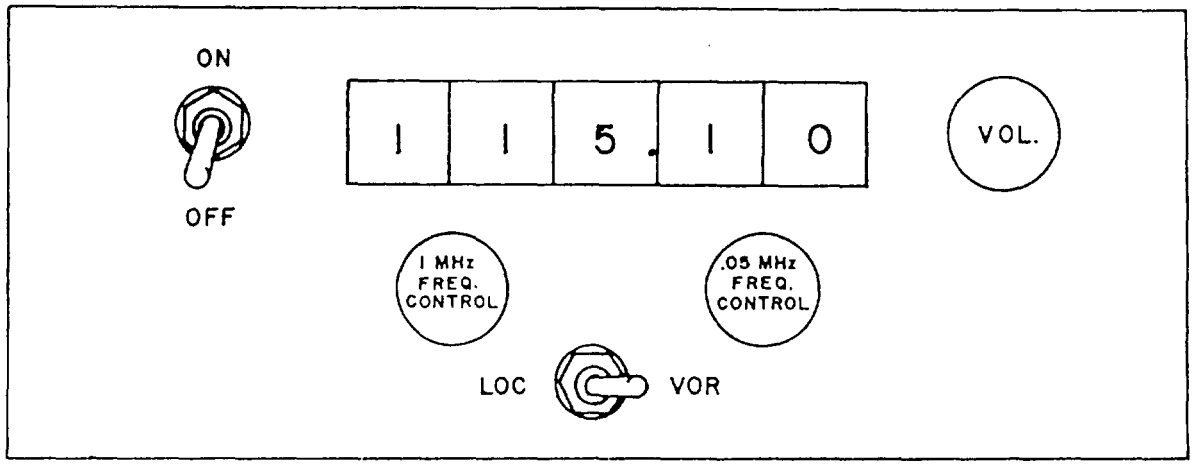


Figure 4-18.-Typical VOR control box.

(ID-249 or equivalent), course indicator (ID-250 or equivalent), and an antenna. Figure 4-16 is a simplified block diagram of a typical VOR receiver. Figure 4-17 is a picture of the course deviation indicator and course indicator used by the system, and figure 4-18 is a diagram of a typical VOR control box.

The VOR control box contains an ON-OFF power switch, which applies power to the complete system, two tuning controls (1 MHz and .05 MHz) to select the receiver's frequency, a frequency indicator, a volume control for adjusting the VOR audio fed to the ICS, and a VOR-LOCALIZER function select switch. The LOCALIZER function will not be discussed, but a representative system (AN/ARA-63) will be discussed under the ILS heading.

This receiver is tunable in increments of .3 MHz from 329.3 to 335.0 MHz, and it can be tuned simultaneously with the VOR control box with the appropriate localizer frequency. Table 4-1 shows various frequencies and their uses.

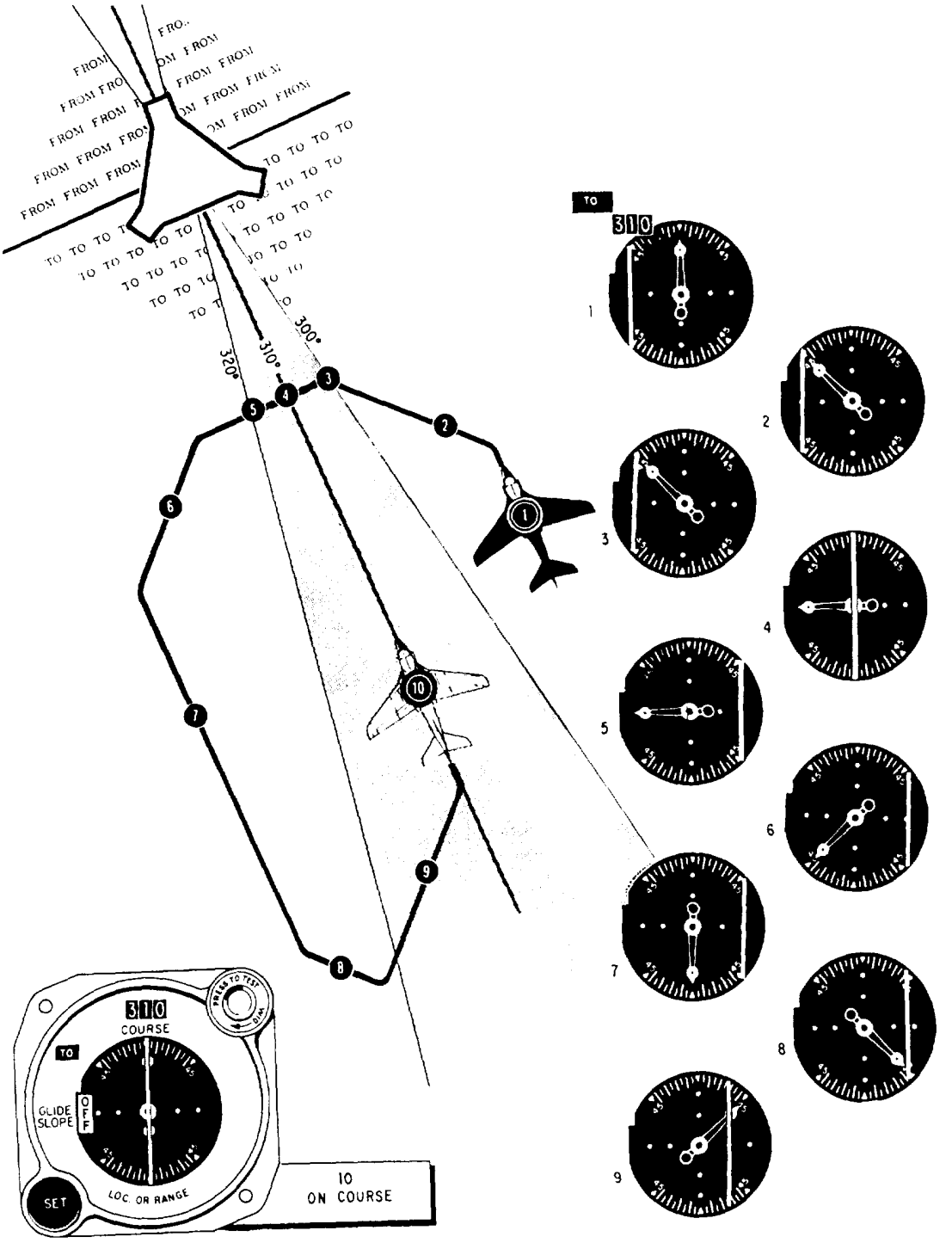


Figure 4-19.-ID-249 presentations at various positions in relation to the VOR station.

The control box tuning controls enable the operator to select any ILS or VOR station frequency. Most VOR receivers tune higher than 117.95 MHz; that is, either to 135.95 MHz or to 151.95 MHz. Due to this added capability, the VOR receiver maybe used as a secondary VHF communication receiver. As a result of antenna polarization, the signal may be degraded, but still usable.

The control box electrically operates the receiver's autopositioners, which mechanically tune the RF amplifier circuits, local oscillators, and mixer stages to the desired frequency.

As in any superheterodyne receiver, the received signals are processed by the RF amplifiers and converted to a lower IF frequency for demodulation. The portion of the received signal that contains voice or station identification is detected and fed to the audio circuits and out to the aircraft's ICS.

The portion of the received signal that contains omni-bearing information (30-Hz modulated variable phase and 9.96-kHz subcarrier, 30-Hz frequency modulated reference phase) is fed to the instrumentation circuits. The instrumentation section of the receiver is divided into two channels, which are referred to as the reference phase channel and the variable phase channel. The reference phase channel is concerned with the 9.96-kHz subcarrier frequency that is modulated by the 30-Hz reference signal. The subcarrier signal passes through a 9.96-kHz bandpass filter and is amplified and applied to an FM-AM translator. The output of the translator is an AM signal modulated with a 30-Hz component. It also still contains the original FM signal. This signal is fed to an AM demodulator bridge circuit, and the resulting 30-Hz output is amplified and applied to a manual resolver located in the course deviation indicator. It is also amplified by an automatic resolver amplifier, and is applied to windings of a resolver located in the course indicator. This signal will be mixed with a 30-Hz signal from the variable phase channel to drive the pointer to the correct bearing to the station.

The reference phase signal, applied across the course deviation indicator manual resolver, re-enters the receiver, and is again amplified and applied as one input to a phase comparator. The variable phase channel accepts the 30-Hz variable phase signal obtained at the output of the detector, amplifies it, and applies the signal as the second input to the phase comparator. It also applies the signal to the course indicator phase signal and drives the pointer to the correct bearing of the received station.

The output of the phase comparator is a vector summation voltage (at), which is applied to three different places. First, it is applied to the course deviation indicator flag alarm circuit to drive the ON-OFF flag to indicate adequate station signal reception. Second, it is also applied to resolver windings in the course deviation indicator. The reference phase signal in the course deviation indicator manual resolver is compared with the variable phase signal, and the resulting vector summation voltage will drive the vertical bar either left or right of center, depending on whether the course to the station is left or right of the course set in the ID-249 course selection window. By turning the Course Set knob, the vertical bar can be centered, and the reading in the course selection window should be the same as that bearing indicated by the pointer of the ID-250. Another use of the course deviation indicator is to select a desired course to the station and fly the aircraft in a direction so as to center the vertical bar, thus ensuring that the aircraft is on the desired course. See figure 4-19.

The third output of the phase comparator is fed to a TO-FROM phase comparator and a 90-degree phase shifter circuit, and to the TO-FROM phase comparator. The vector summation voltage output of the TO-FROM phase comparator is applied to the ID-249 TO-FROM to drive the TO-FROM indicator flag. If a received signal is “from” the course selected in the course selection window, the TO-FROM flag will read “FROM.” If the course set is to the station, the TO-FROM flag will read “TO.” As shown in figure 4-19, the combination of the ID-249 vertical crossbar position, the course selected in the course selection window, and the TO-FROM indication will inform the operator just where the aircraft is located with respect to the received VOR station.

INSTRUMENT LANDING SYSTEM (ILS)

Learning Objective: *Recognize a typical ILS to include the detail theory of operation for the AN/ARA-63 system.*

The ILS is a radio and radar system that enables aircraft to land in low visibility. In this section we will discuss the AN/ARA-63 as a representative system.

PRINCIPLES OF OPERATION

The AN/ARA-63 is an all-weather aircraft approach guidance system, which consists of ground and airborne

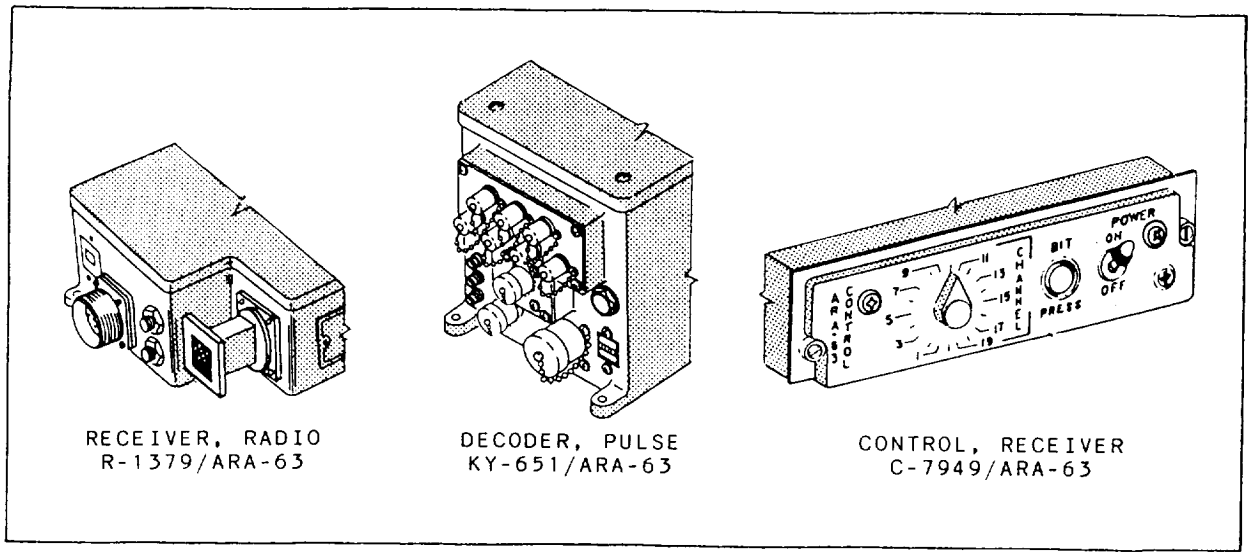


Figure 4-20.-Recelving-Decoding Group AN/ARA-63.

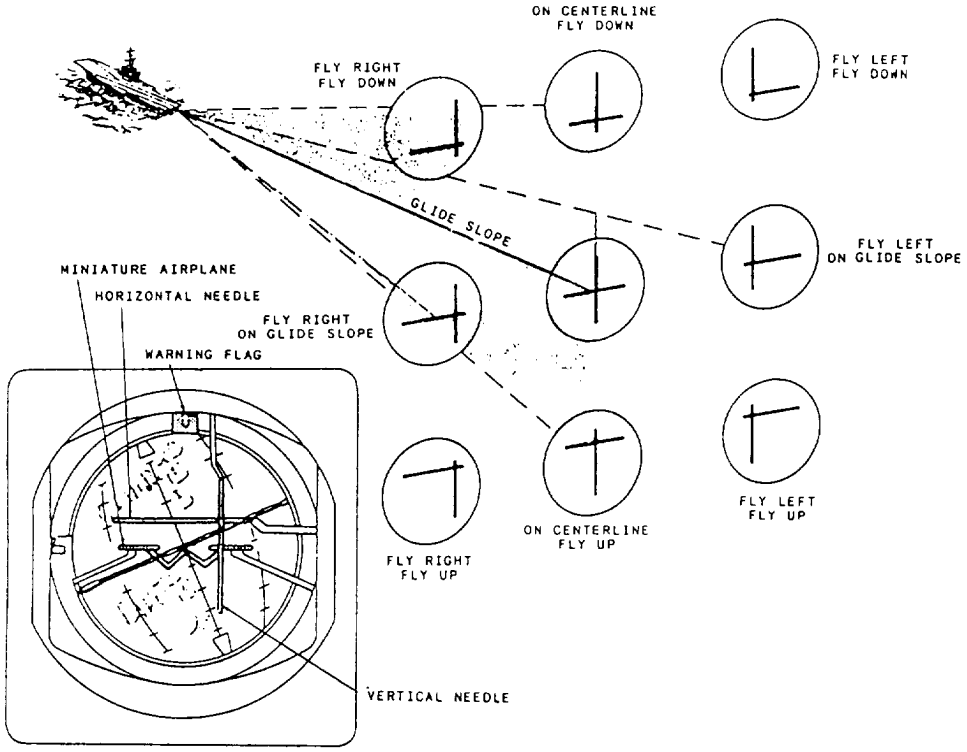


Figure 4-21.-Crosspointer indicator.

equipment. See figure 4-20. Receiving-Decoding Group AN/ARA-63 is the airborne portion of the AACS. This equipment receives coded microwave transmissions from ground or carrier-based azimuth and elevation transmitters, and decodes these signals for display on a crosspointer indicator in the aircraft cockpit. See figure 4-21. A centerline display of both elevation and azimuth on the crosspointer indicator depicts the flight path the pilot must follow to line up accurately with the airport runway or carrier deck. By consecutively scanning through azimuth and elevation, the system provides continuous measurement of the lateral and vertical aircraft deviations from the optimum approach line in space.

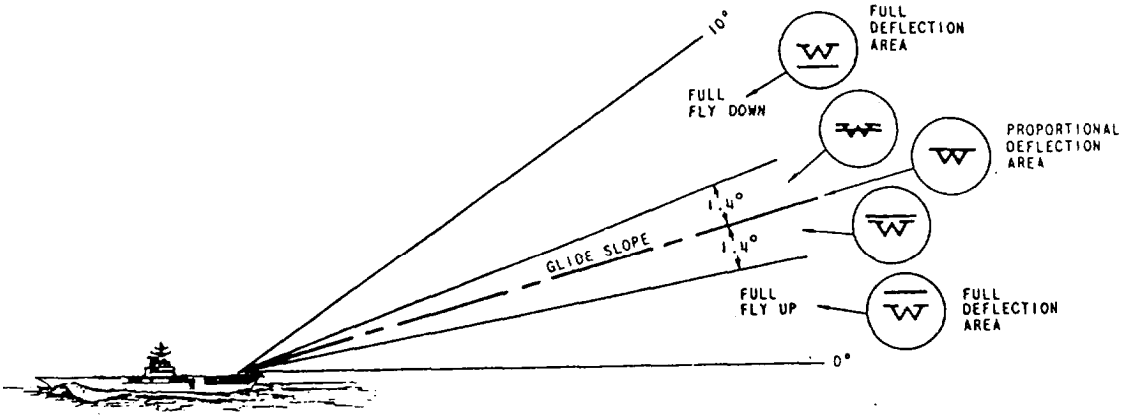


Figure 4-22.-Elevation angle.

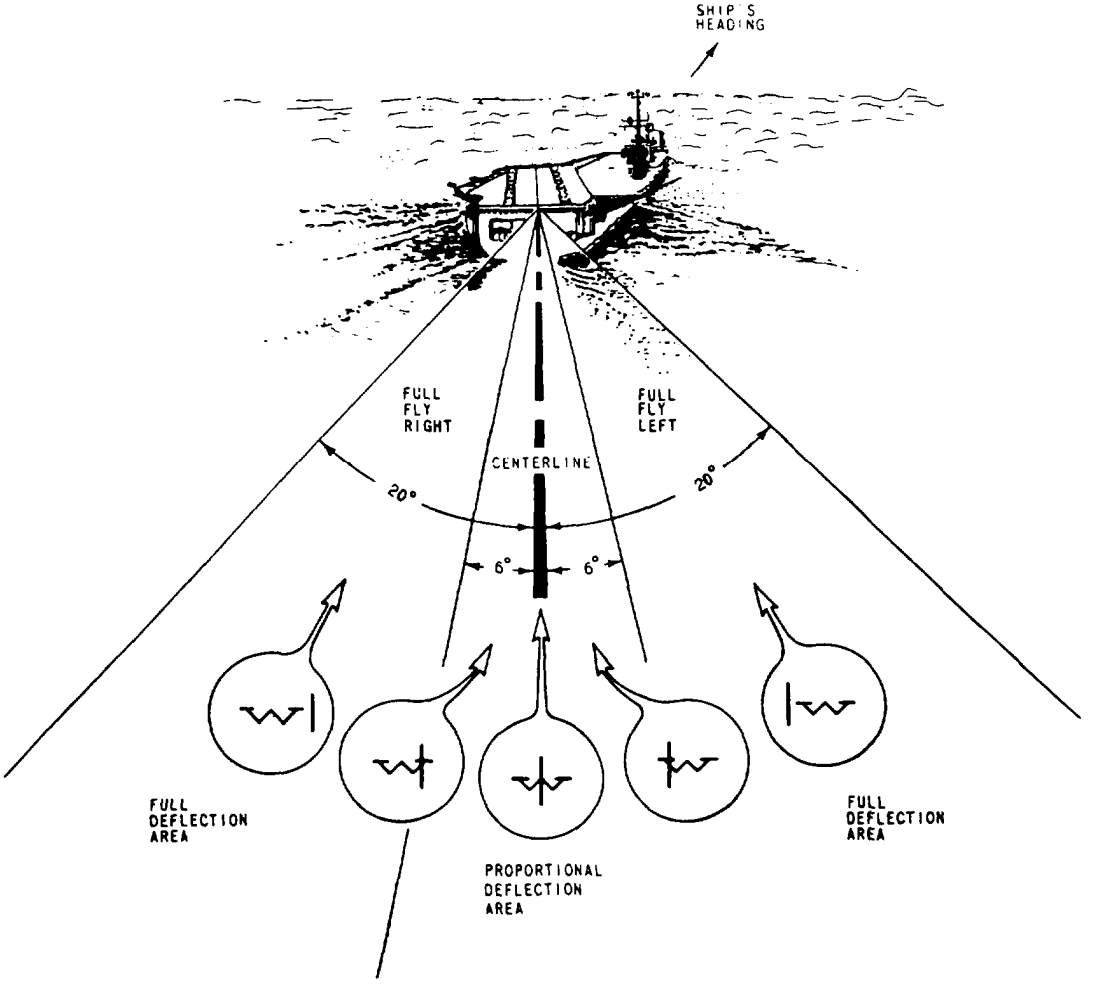


Figure 4-23.-Azimuth angle.

As you can see from figure 4-22, proportional angle information in elevation is displayed ±1.4 degrees from the glide slope. For azimuth, proportional angle information is displayed between ±6 degrees. See figure 4-23. The normal azimuth course is along the runway centerline. The normal glide slope is 3 degrees, but, by means of jumper wires in the airborne decoder, the glide scope can be adjusted from 2 to 5 degrees in 0.5-degree increments. (The glide slope can also be adjusted on the ship's transmitter.)

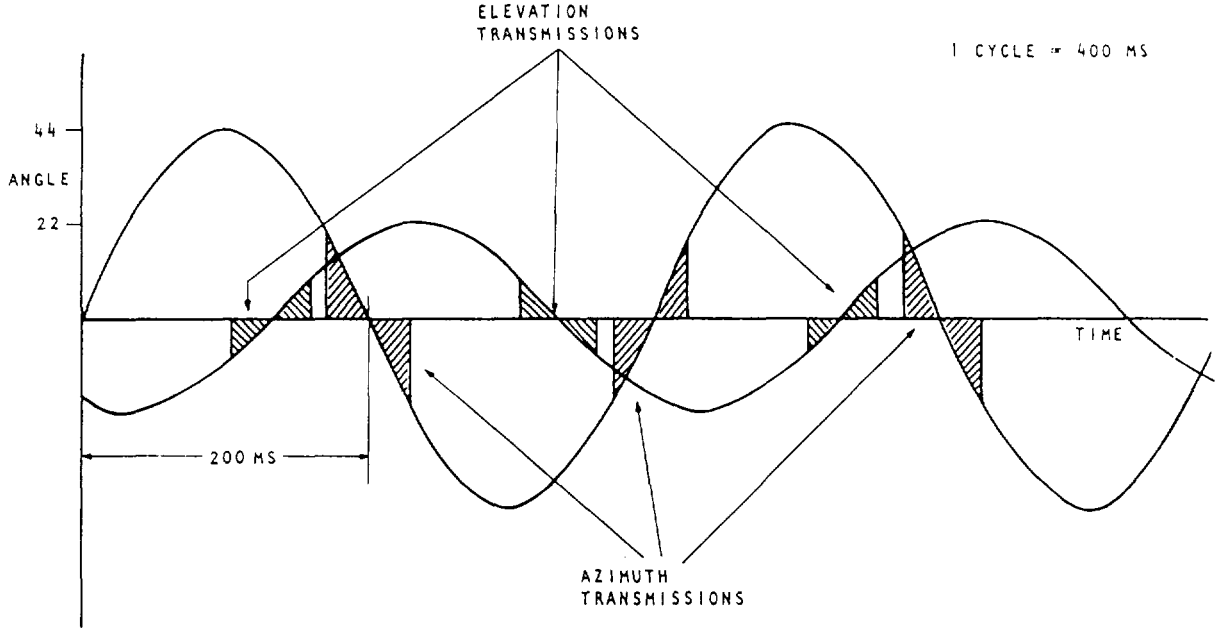


Figure 4-24.-Antenna scan diagram.

Input Signal Characteristics

The ground equipment transmits a complete series of elevation and azimuth signals five times per second using sector scanners in simple harmonic motion. See figure 4-24. Since the mechanical scan rate is 2.5 Hz, the 5-Hz signal rate is achieved by transmitting during a portion of the left-to-right azimuth scanning time, and transmitting again during a portion of the right-to-left scanning time. A similar but interlaced two-way scan is used for elevation.

The azimuth and elevation transmissions share time on the same frequency; that is, the on-and-off periods are governed by the synchronized scans of the two antennas. Each transmission consists of a series of paired pulses, by means of which the transmission is encoded with either azimuth or elevation identification and with angle information. The signal pulses are 0.3 μsec wide.

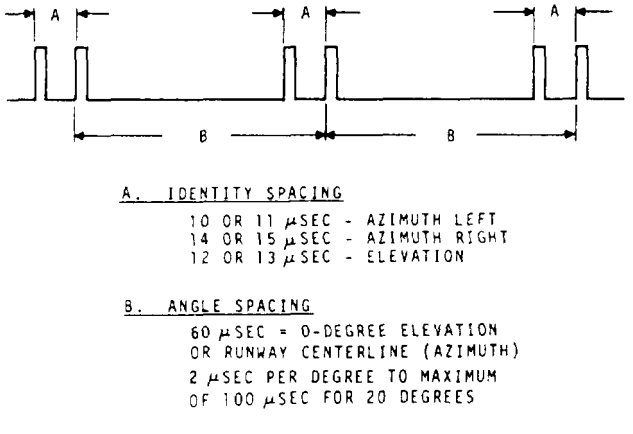


Figure 4-25.-Pulse coding.

Figure 4-25 shows the pulse coding. The time between the first two pulses in the data group identifies the group as being from the elevation or azimuth transmitter. The two pulses in each pair are spaced 12 μsec apart during elevation transmission, and either 10 or 14 μsec apart during azimuth transmission. These spacings can be increased by 1 μsec to provide an additional channel at each frequency.

The time between pulse pairs represents a value of angle data. As the antenna scans, the angle data changes at 1/8-degree intervals (0.25 μsec). Figure 4-26 shows the variation of angle data spacing with time.

Output Signals

Outputs from the AN/ARA-63 drive the crosspointer indicator and are as follows:

Azimuth Deviation	±2.2 V Linear	>7.0 V Pegged Off scale
Elevation Deviation	±2.2 V Linear	>7.0 V Pegged/Off scale
Azimuth Flag	0.5 V	
Elevation Flag	0.5 V	

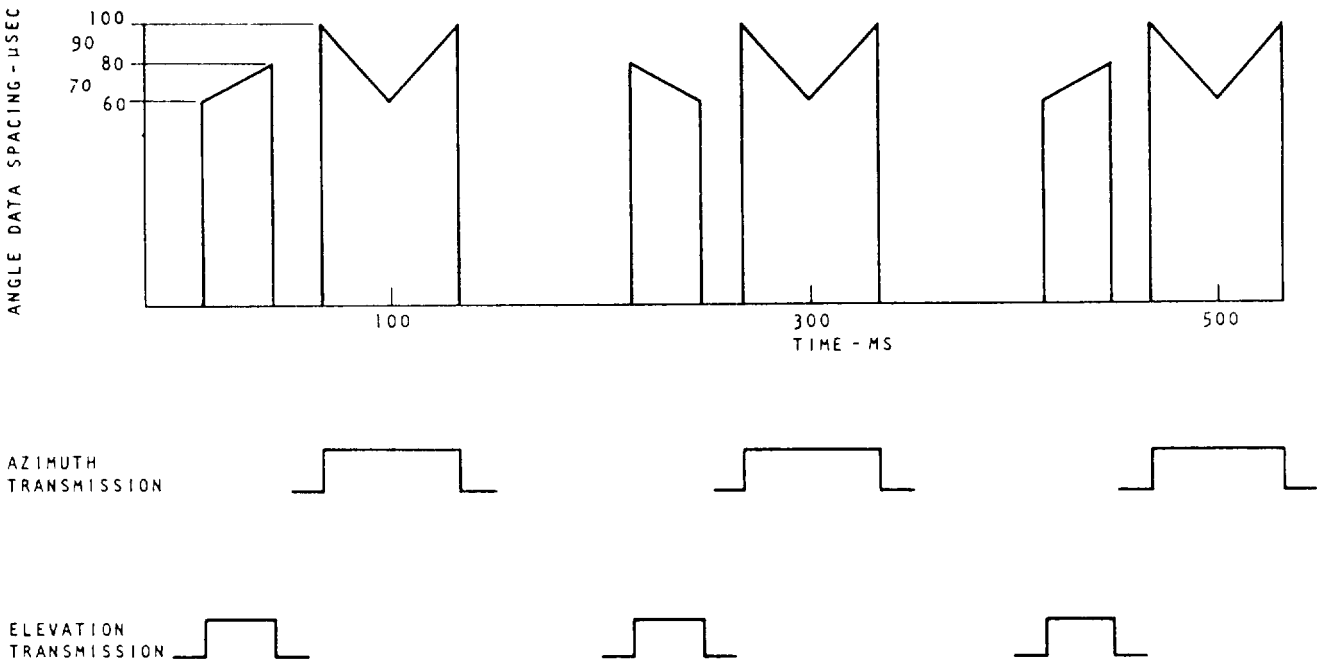


Figure 4-26.-Angle data spacing.

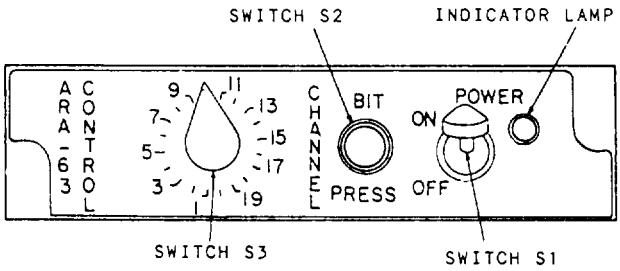


Figure 4-27.-Controls and indicators for Receiver Control C-7949/ARA-63.

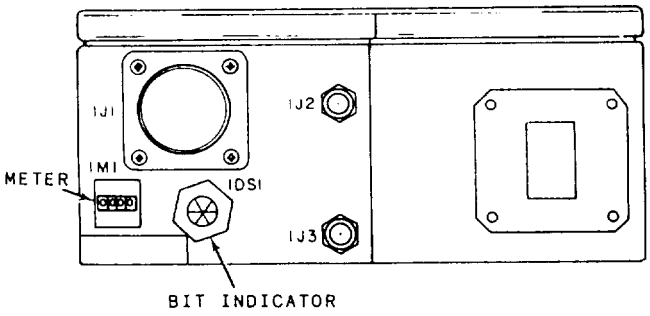


Figure 4-28-Controls and indicators for Radio Receiver R-1379/ARA-63.

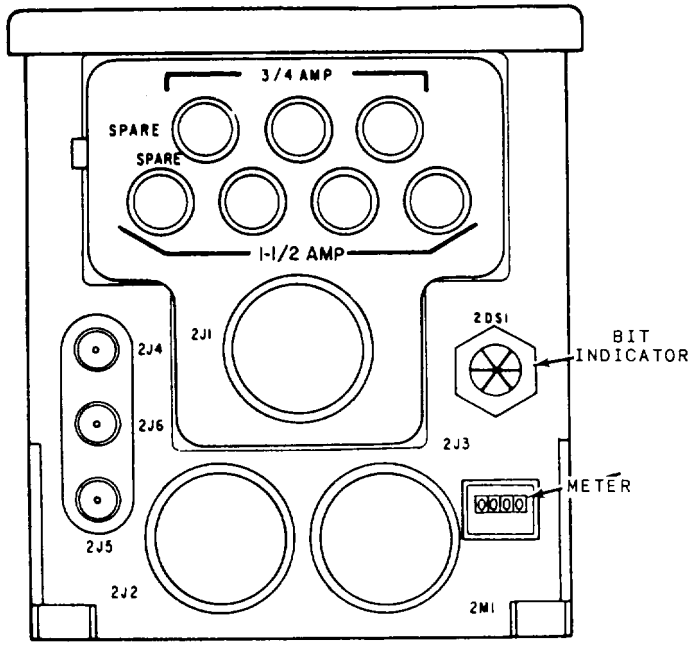


Figure 4-29.-Controls and indicators for Pulse Decoder KY-651/ARA-63.

NOTE: Crosspointer indicators ID-1144, ID-811, and ID-1329 are designed for 2.2-V full-scale operation into 1000 ohms or 2.2 mA. Crosspointer indicators ID-249, ID-351, ID-48, and ID-387 are designed for 150-μA full-scale operation, and a 15,000-ohm resistor is required in series with each deviation output.

CONTROLS AND INDICATORS

See figures 4-27,4-28, and 4-29 for the controls and indicators. These illustrations show the radio receiver R-1379/ARA-63, the pulse decoder KY-651/ARA-63, and the receiver control C-7949/ARA-63.

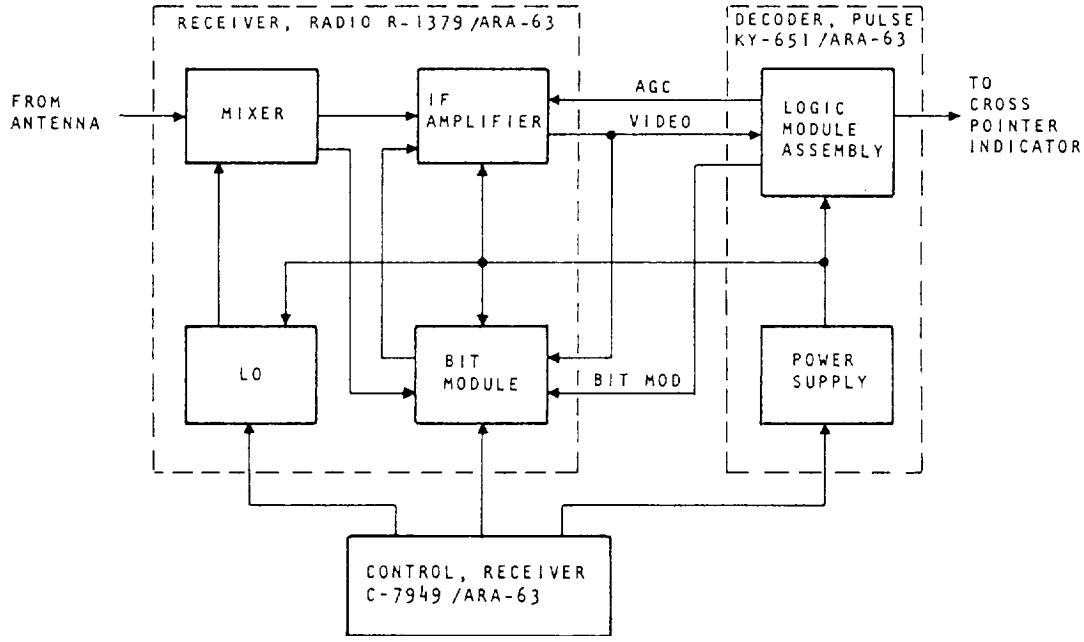


Figure 4-30.-AN/ARA-63 simplified block diagram.

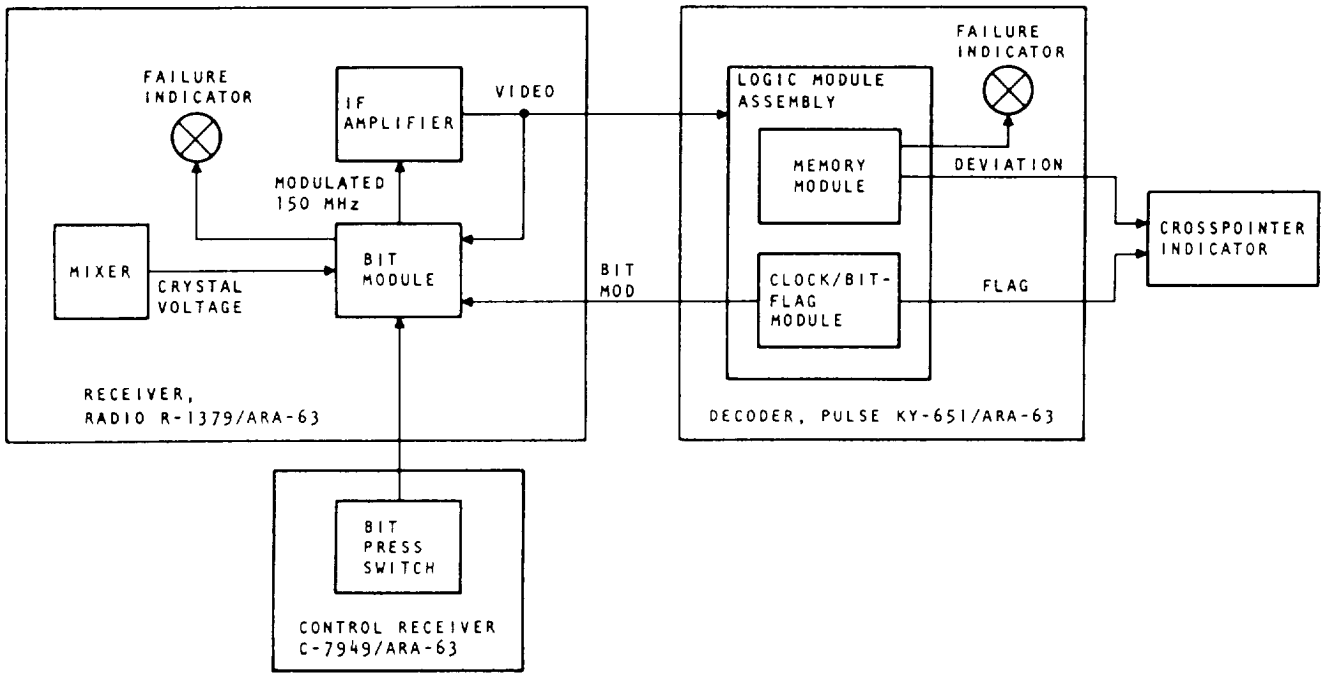


Figure 4-31.-BIT functiona1 block diagram.

Radio Receiver R-1379/ARA-63

RF input signals intercepted by the antenna are mixed with the LO to produce an IF output of 150 MHz. The mixer is a conventional balanced wave guide mixer. The LO is a crystal controlled, solid-state unit employing multipliers, amplifiers, and falters to obtain the required output. The outputs from the mixer are amplified by the IF amplifier. The input of the amplifier is filtered to reject spurious signals. The detector is a single diode, with the video output faltered to remove the IF component. The detector output is amplified by a video amplifier. The stages are direct-coupled and designed for preservation of the pulse signal. The receiver also contains a BIT module. See figure 4-30.

Pulse Decoder KY-651/ARA-63

The decoder is composed of two basic units. They are the logic module assembly and the power supply assembly.

The power supply furnishes all dc voltages necessary for the operation of the AN/ARA-63 (both receiver and decoder), except for 28 Vdc aircraft power.

The logic module assembly converts the video pulses from the receiver to dc voltages to drive the indicator. It also generates an automatic gain control (AGC) voltage for the IF amplifier. See figure 4-30.

Receiver Control C-7949/ARA-63

The receiver control controls system power, channel selection, and the BIT. See figure 4-31.

The function of the BIT is to provide the pilot with a degree of confidence of the system's operation. The pilot activates the BIT by pressing and holding the BIT PRESS switch on the receiver control. Correct functioning of the system gives the following readings on the crosspointer indicator.

- Vertical needle—oscillating 1/3 scale fly right and 1/3 scale fly left at a 4-second-per-cycle rate.
- Horizontal needle-on the glide slope in elevation (center scale).

A malfunction detected by the BIT turns on failure indicators in the receiver or the decoder, or in both. When the BIT is activated, a video train of pulses is generated from the clock/BIT-flag module in the decoder. These pulses are used to modulate the 150-MHz BIT oscillator in the BIT module in the receiver. The output of the oscillator is connected to input jack J3 of the IF amplifier. The output of the IF amplifier is fed back to the BIT module. If no output is detected by the BIT module because of IF amplifier failure, both failure indicators (on the receiver and decoder) will appear.

The BIT module also checks the mixer crystal voltage. Inadequate mixer crystal voltage inhibits the 150-MHz BIT oscillator so that the flags on the crosspointer indicator are in view, and both failure indicators are on. The BIT monitor in the memory module evaluates the decoder response to the BIT simulated input and generates a decoder failure indicator signal when a malfunction is detected. Failure of the power supply also results in a decoder failure indicator signal.

THEORY OF OPERATION

The front end of the radio receiver consists of the LO, mixer, and IF amplifier subassemblies. The subassemblies are repairable only at the factory. Therefore, the repair of these subassemblies is not covered in this chapter. For reference, characteristics of these units are shown in figures 4-30, 4-31, and 4-32.

The BIT module contains a 150-MHz oscillator and logic circuitry. This circuitry is used to detect the video output, evaluate the mixer crystal voltage, and to drive the receiver failure indicator. The decoder consists of the logic module assembly and the power supply assembly.

Logic Module Assembly (General Description)

The logic module assembly consists of five two-card modules. The five modules are the video-identity, the error, the memory, the AGC, and clock/BIT-flag modules.

The video-identity module receives video from the IF amplifier and performs a three-level threshold detecting process. It also identifies the intrapair (identity) pulse spacing being received. This module consists of a video decoder board and an identity shift register board. Each board is described separately later in this chapter.

The error module receives the quantized video from the video-identity module and compares the spacing of the angle data pulses with reference timing signals. This allows the error module to produce an error pulse that is proportional to the angular error of the aircraft from the glide path.

The memory module converts the error pulse from the error module into a dc voltage. This dc voltage is proportional to the error pulse width. The memory module averages all valid error pulses received during a beam. It then averages the result of the last beam with the present beam. The output signal is a dc voltage proportional to the average angular offset of the aircraft from the glide path.

The AGC module develops a dc voltage that is a logarithmic function of the received signal strength. This is used to control the gain of the IF amplifier.

The clock/BIT-flag module supplies the basic clock frequencies needed throughout the decoder for timing purposes, and also produces the pulse signal to modulate the 150-MHz BIT oscillator. The flag board determines if sufficient video information is being received for proper tracking. It also senses if unwanted information

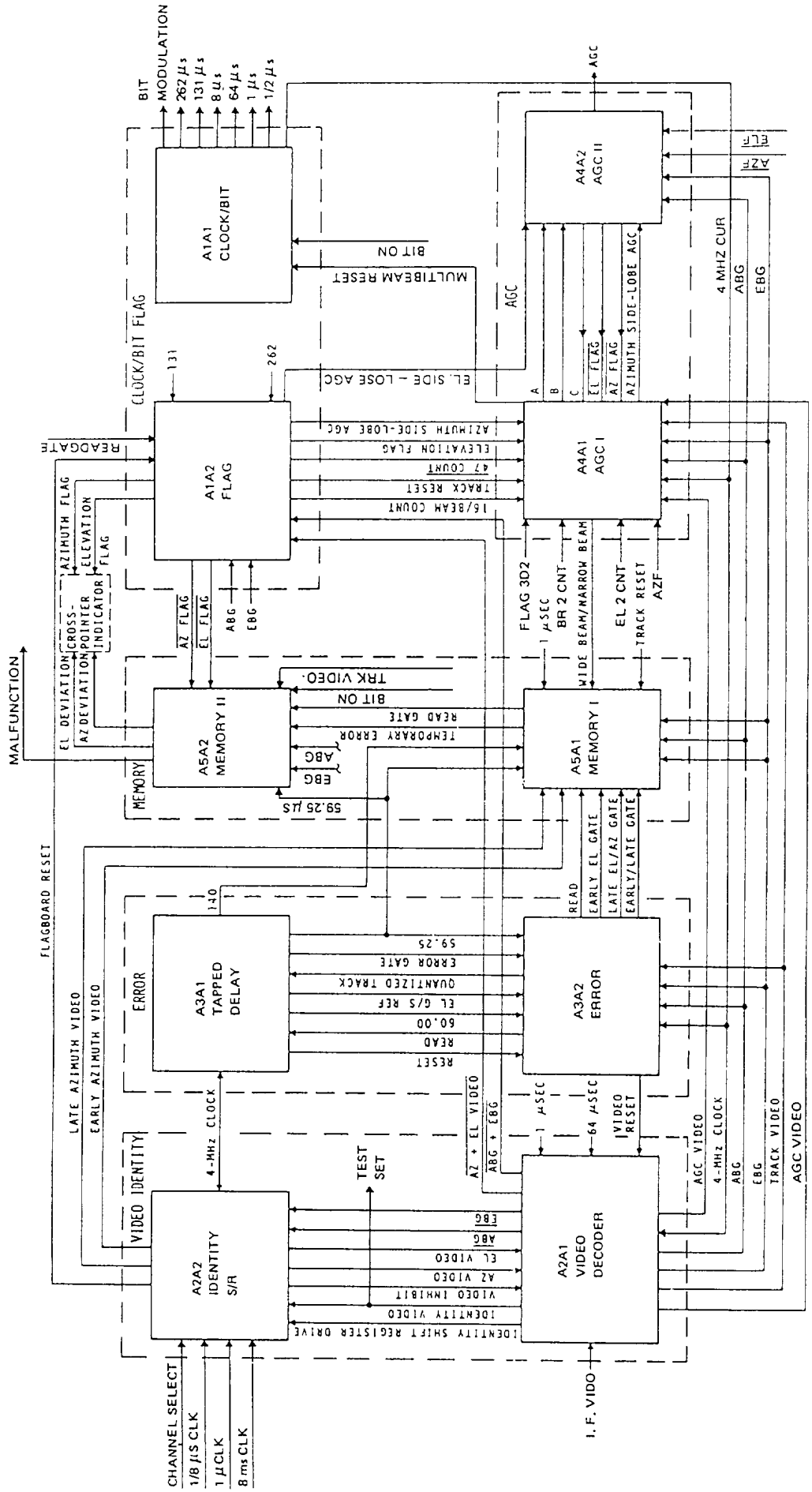


Figure 4-32.–Logic module assembly.

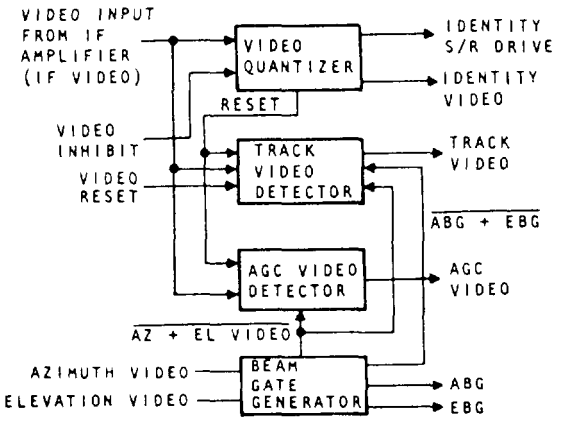


Figure 4-33.-Video decoder board.

is being decoded. The output signals are go/no-go indications.

Video Decoder Board

The video decoder board receives the IF video pulses from the IF amplifier. It can be subdivided into four basic sections. They are the video quantizer, the track video detector, the AGC video detector, and the beam gate generator sections. See figure 4-33.

If the IF video pulses exceed 0.57 V, the video quantizer will produce two output signals–identity shift register drive and identity video. These signals are used in the identity shift register board to decode the identity pulses.

If the IF video pulses exceed 1.2 V and an azimuth or elevation beam gate is present, the track video detector will produce an output pulse track video that is used to decode the angle data information.

If the IF video pulses exceed 3.2 V and Az + E1 video is present, the AGC video detector will produce an output pulse (AGC video) that is used to determine the respective AGC voltage.

The beam gate generator receives azimuth video and elevation video pulses from the identity shift register unit. These pulses have been identified as having the correct identity spacing for either azimuth video or elevation video. If three azimuth video pulses are received, none of which are spaced more than 512 sec apart, an azimuth beam gate (ABG) signal is generated. If three elevation video pulses are received, none of which are spaced more than 512 sec apart, an elevation beam gate (EBG) signal is generated.

Identity Shift Register Board

The identity shift register is the digital equivalent of a delay line. Its function is to decode properly spaced elevation and azimuth identity pulse pairs. The identity shift register can be divided into six basic sections. The sections are the high-low channel select circuit, the 1-sec delay, the 4-sec video inhibit circuit, the shift register, the identity decoder, and the side-lobe counter reset circuit sections.

The function of the high-low channel circuit is to allow the pulse going down the shift register to pass through undelayed or to delay the pulse by 1 sec. The pulse is controlled by the channel selector signal, which doubles the number of operating channels without adding RF channels. This is accomplished by using two different identity spacings with the same frequency. Thus, one channel can use 15.412-GHz and x-μsec spaced identity pulses, and a second channel can be 15.412-GHz and (x + 1) μsec spaced identity pulses.

The 4-see video inhibit circuit generates a 4-sec video-inhibit pulse. The video inhibit pulse inhibits the video quantizer (on the video decoder) from generating identity video, to ensure that no reflected signals will be picked up within 4 μsec of a valid signal.

The shift register is a series of 8-bit shift registers connected in series. The output taps of the shift register are shown (fig. 4-34) with the high-low channel in the

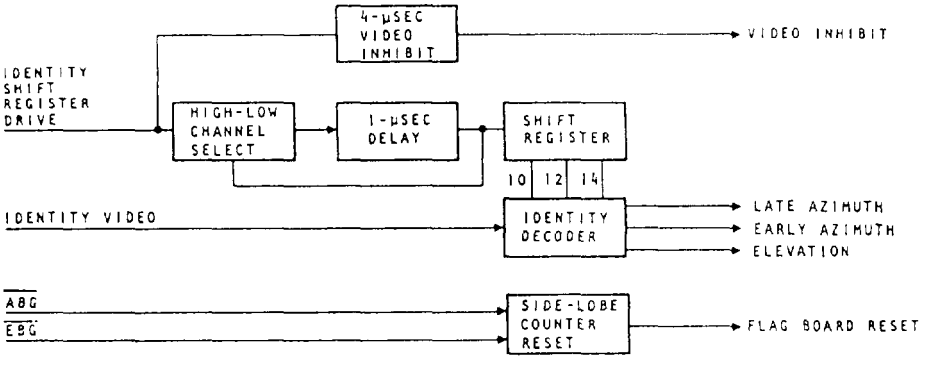


Figure 4-34.-Identity shift register board.

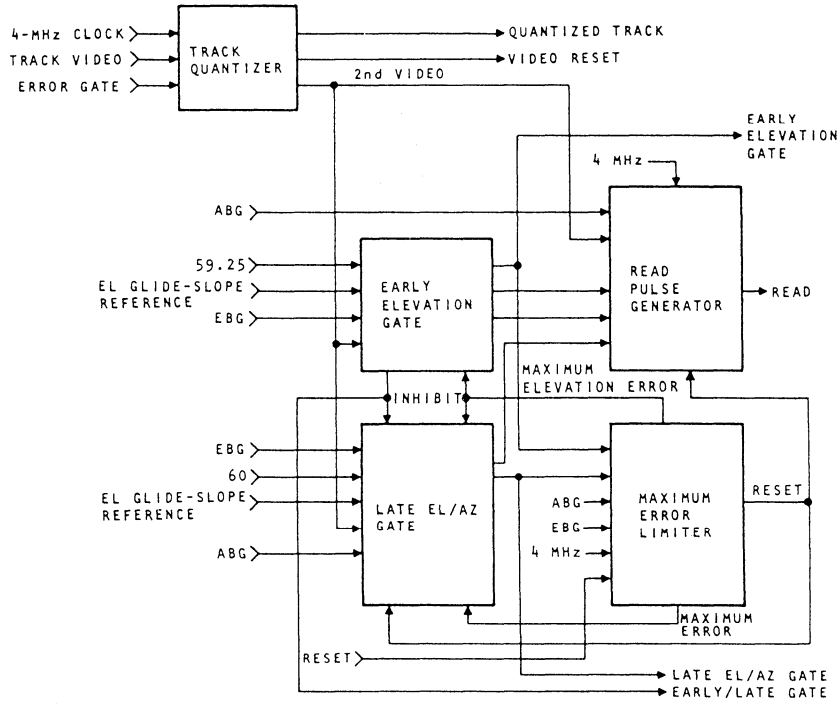


Figure 4-35.—Error board.

undelayed (channels 1 through 10) position. The information is clocked through the shift register by the 4-MHz clock. See figure 4-34.

The basic decoding principle used in the shift register is to delay the first pulse of an identity pulse pair a period of time equal to the time expected between the pulses. When the delayed pulse and the second pulse of the pulse pair are compared at a NAND gate,an output will be produced if they are in time coincidence. For each tap on the shift register, a pair of identified pulses can be decoded. There are taps at 10, 12, and 14 sec corresponding to left azimuth, elevation, and right azimuth.

A side-lobe counter reset signal will be sent to the video counter on the flag board 64 MS after the start of abeam gate. The counter is also reset if the type of beam gate changes.

Error Board

The function of the error board is to produce a pulse output (late EL/AZ gate, early/late gate), the width of which is proportional to the spacing between the angle data pulses. The maximum width of this pulse is limited to prevent the crosspointer needles from being driven off scale. The error board also produces three other signals. These signals are the track quantizer, the read, and the video reset signal. See figure 4-35.

TRACK QUANTIZER SIGNAL.— The function of the track quantizer is to generate a pulse that has a fixed width of two clock pulses. This signal is called quantized track. It also produces a reset pulse (video reset), which goes to the video decoder board, and a pulse called 2d video (if error gate is present). If a series of angle data pulses is received, only the second and succeeding pulses are allowed to pass.

The early elevation gate circuitry will produce a pulse output that is proportional to the angle data pulse spacing if the following conditions are present:

EGB signal present.

59.25 -µsec signal has been received but an elevation glide-slope reference signal has not been received.

The pulse is initiated by the 2d video pulse, and it is terminated by elevation glide slope reference or by a maximum elevation error signal, whichever occurs first.

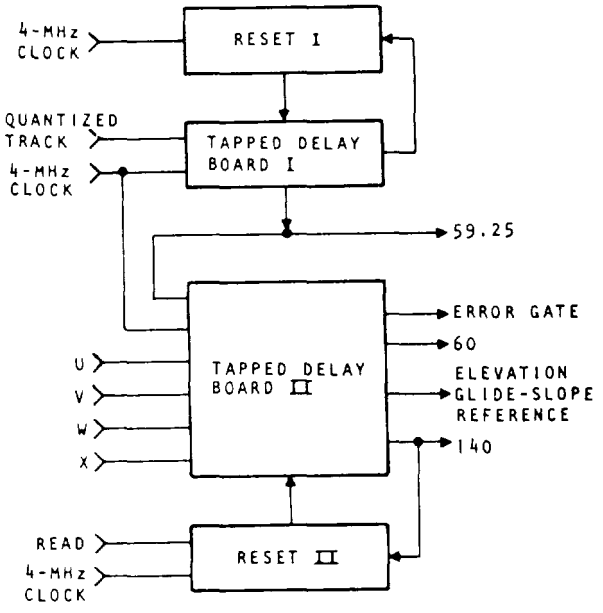


Figure 4-36.-Tapped delay board.

- This circuit also generates an inhibit signal, which prevents a late elevation/azimuth gate from being generated during an early elevation gate.

The late elevation/azimuth gate will produce a pulse output that is proportional to the angle data pulse spacing if the following conditions are present:

1. ABG present:

   If an ABG signal is present, the gate is initiated by the 60-sec delayed pulse and terminated by a 2d video pulse or a maximum azimuth error pulse, whichever occurs first.

2. EBG present with the following conditions:

   - No early elevation gate was generated.
   - Elevation glide-slope reference.

If an EBG signal is present, the gate is initiated by elevation glide-slope reference and terminated by the 2d video pulse, or by maximum elevation error signal, whichever occurs first.

The maximum error limiter sets a limit to the pulse width of the early elevation gate and the late elevation/azimuth gate. The maximum gate widths are as follows:

- Late elevation/azimuth-14.25 μsec
- Early elevation-4.25 μsec

READ SIGNAL.– The read pulse generator circuit generates a pulse whenever a 2d video pulse is received during a late elevation/azimuth error gate, or when a glide-slope reference is generated in the case of an early elevation gate. The read pulse goes to the memory I board in the memory module and to the tapped delay board to reset the tapped delay board II counter.

The function of the tapped delay board is to generate timing signals referenced to the quantized track signals. See figure 4-36. The tapped delay board I circuit receives the quantized track signal from the error board. The tapped delay board I circuit is a ripple-through counter used as a timer. An output pulse called 59.25 is produced 59.25 μsec after a quantized pulse is received. This pulse goes to the error board and to the tapped delay board II circuit.

VIDEO RESET SIGNAL.– The reset I circuit resets the tapped delay board I circuit whenever a 59.25-psec signal is generated.

The tapped delay board II is another ripple-through counter used as a timer circuit. The tapped delay board II circuit produces four output signals:

- Error gate starting 59.25 μsec after quantized track and terminated by the read signal or by the 140-sec count.
- A signal called 60, which occurs 60 μsec after the tapped delay board I receives a quantized track signal.
- A signal called elevation glide-slope reference. This signal can be adjusted to occur between 64 and 70 μsec (in 1-μsec intervals) after the quantized track signal is received.
- A signal called 140, which occurs 140 μsec after a quantized track signal is received.

The reset II circuit resets the tapped delay board II whenever a read signal is generated or a 140 signal is generated. It also produces a reset signal that goes to the error board to reset the maximum error counter.

Memory Module

The memory module consists of two boards called memory I and memory II.

MEMORY I BOARD.– The function of the memory I board is to convert the pulse width of the early/late gate (generated on the error board) into a dc signal that is proportional to the early/late gate pulse width, starting from zero volt each time a new gate is

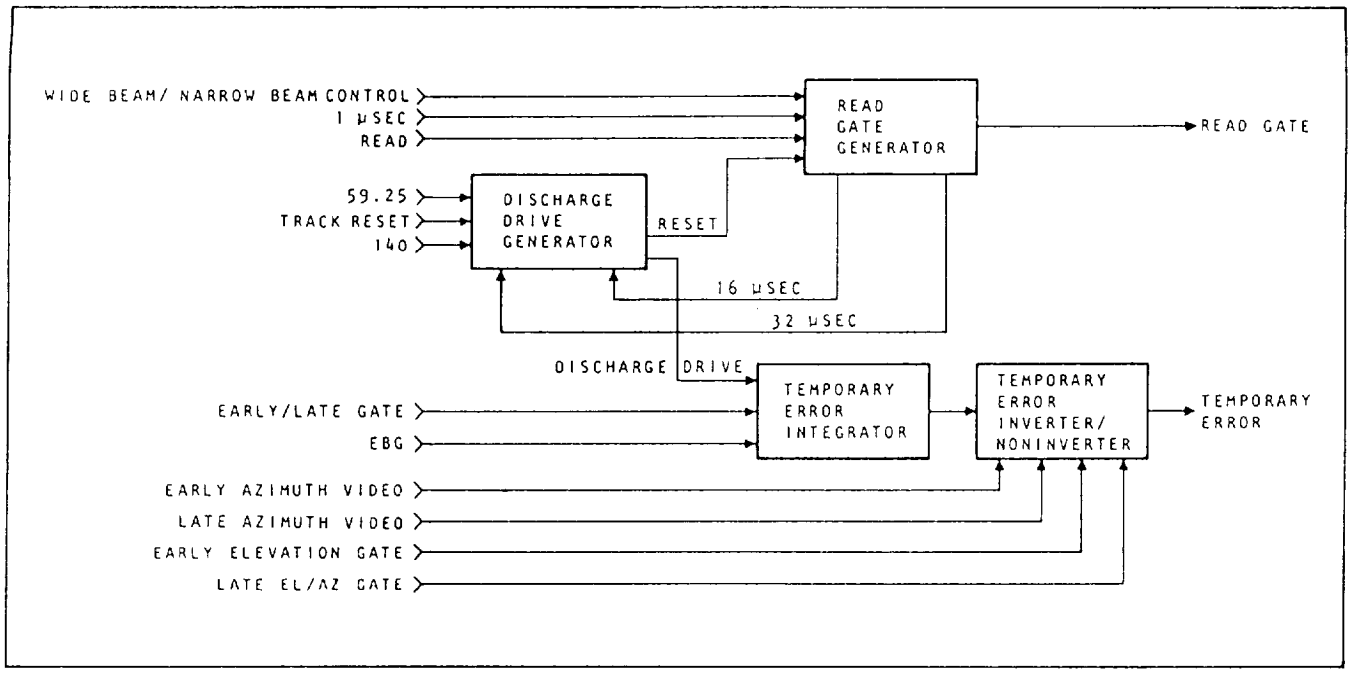


Figure 4-37.-Memory I board.

received. Memory l also generates a read gate signal and a discharge drive signal. See figure 4-37.

The temporary error integrator generates the temporary error signal, which is proportional to the pulse width of the early/late gate each time a valid error signal is received. This signal starts from ground each time a new early/late gate is received.

The temporary error inverter inverts the temporary error signal if a late azimuth video (fly right) signal is present or if an early elevation gate (fly up) is generated.

The read gate generator produces two output signals–the read gate and a control signal to the discharge drive generator.

The read gate is either 16 μsec wide or 32 μsec wide, depending on the level of the wide beam/narrow beam control. If the wide beam/narrow beam signal is high (47 count signal present-see AGC I), the read gate is 16 μsec wide. If the wide beam/narrow beam signal is low (narrow beam being received), the read gate is 32 μsec wide. This signal goes to memory II and gates the temporary error signal into memory.

The discharge drive generator produces a signal called discharge drive. Its purpose is to discharge the temporary error integrating capacitor. The discharge drive signal is generated whenever any one of the following signals is present.

- End of read gate
- End of either azimuth or elevation beam
- 140 μsec after quantized track

The temporary error integrating capacitor is allowed to charge 59.25 μsec after quantized track.

MEMORY II BOARD.– The function of memory II board is to receive the temporary error signal from memory I to produce output signals called azimuth deviation and elevation deviation, which drive the crosspointer indicator. The amplitude of the output signal is the average of the latest computed deviation and the computed deviation of the previous beam. See figure 4-38.

The temporary error signal goes to the feedback network. This is multiplexed between azimuth and elevation to permit the memories to be updated on a beam-to-beam basis so that an average correction may be made during beam scan time.

The output selector switch selects the proper signal to be fed into the feedback network. This is determined by either ABG´ or EBG´.

The input selector switch connects the output of the feedback network to the proper integrator. This is determined by ABG´ and EBG´.

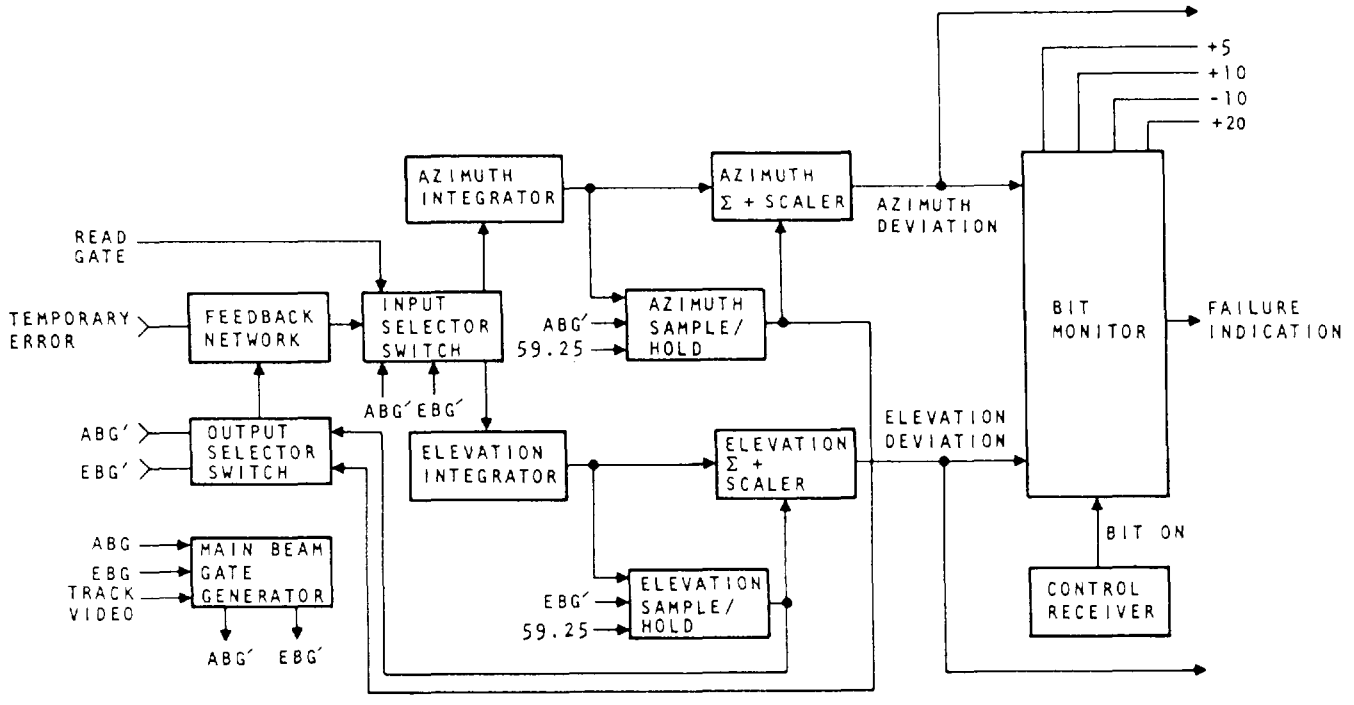


Figure 4-38.-Memory II board.

The azimuth integrator integrates the signal from the feedback network with stored de voltage. The signal from the feedback network is the sum of the temporary error signal and the signal from the azimuth sample/hold network. The feedback ratio is determined by the feedback network and is integrated for a time equal to the read gate signal. Thus, the azimuth integrator is updated each time a read gate is generated. At the end of a beam, it should contain the average error decoded in that beam.

At the beginning of an azimuth beam, the voltage stored by the azimuth integrator is stored in the azimuth sample/hold network.

The azimuth summation and scaler network has two functions: it averages the output of the azimuth integrator and the azimuth sample/hold. The elevation integrator is identical to the azimuth integrator. The elevation sample/hold is identical to the azimuth sample/hold. The elevation summation and scaler is identical to the azimuth summation and scaler. The function of the main beam gate generator circuit is to produce an ABG´ or an EBG´ to prevent the circuits in memory II from responding to beam gates generated from side lobes.

The BIT monitor evaluates the decoder response to the BIT simulated input and generates a decoder failure indicator signal when a malfunction is detected. Failure of the power supply also results in a decoder failure indicator signal. The AGC module develops AGC bias signals, depending on signal strength for both elevation and azimuth, and generates a wide-beam/narrow-beam signal. The eight count circuit counts the AGC video pulses. When eight pulses are received, a signal is sent to the azimuth AGC circuit and the elevation AGC circuit. The azimuth AGC circuit generates a dc signal, which is dependent upon the received signal. The AGC circuit will raise the dc level (desensitize the IF amplifier) if a pulse is received from the eight-count circuit, or lower the dc level if a two-count pulse is received. The eight- and two-count circuits, by varying the dc levels, ensure that pulses are being received in the top center portion of the beam (not side lobe or less than

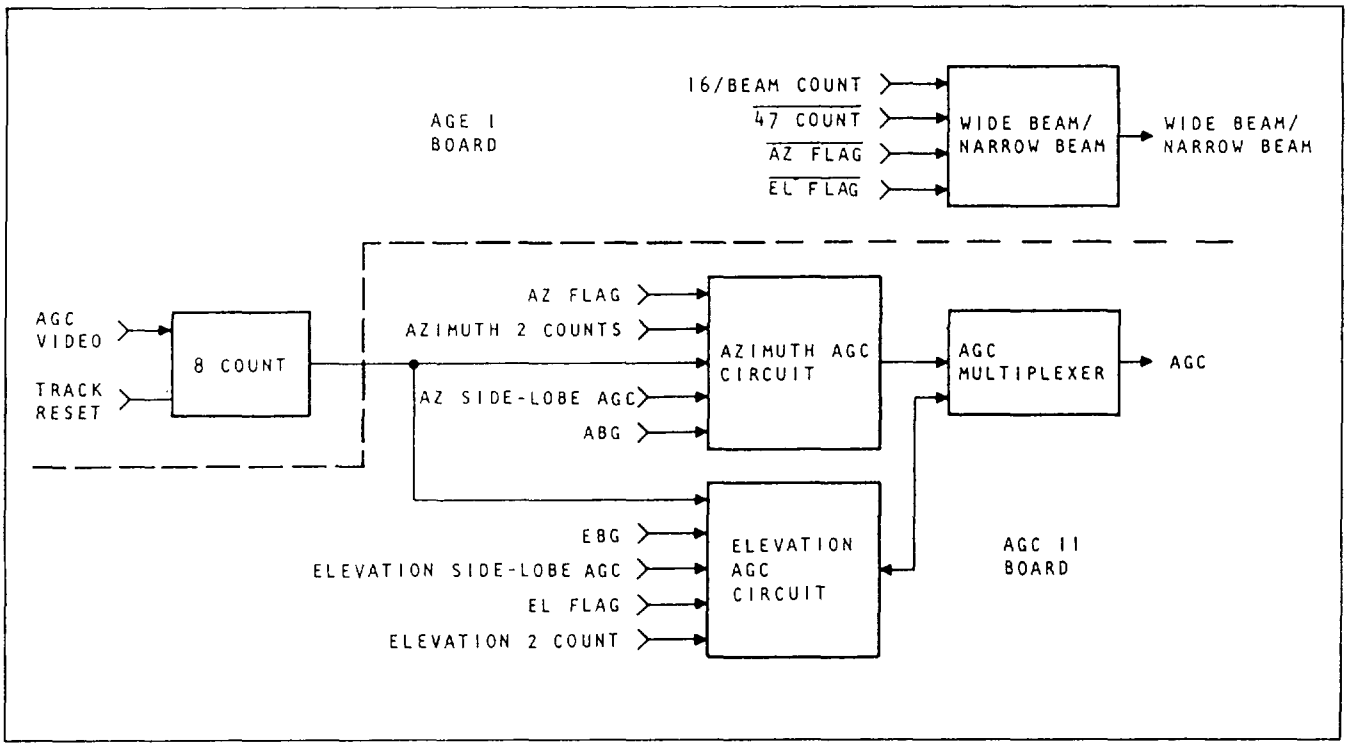


Figure 4-39.-AGC module.

16 pulses).The elevation AGC circuit is identical to the azimuth AGC circuit. See figure 4-39.

The function of the AGC multiplexer is to select the proper AGC voltage. If an ABG signal is present, the azimuth AGC voltage or the noise level AGC is selected, depending on which is more positive. Similarly, if an EBG signal is present, either the elevation AGC or the noise level AGC is selected, depending on which is more positive. The wide beam/narrow beam control determines whether more than 47 or less than 47 pulses within the beam are being received. Since this receiver is able to fly against 1- and 2-degree wide beams, a method of detecting which beam width is being received is necessary. This allows the memory to make unity correction in one beam passage. Unity correction corresponds to critical damping in a servo system.

Clock/BIT-Flag Module

The clock/BIT-flag module consists of the clock/BIT board and the flag board.

The function of the clock/BIT board is to produce clock pulses of various frequencies, which are used as timing pulses throughout the decoder. It consists of a 4-MHz crystal oscillator, which produces the basic 4-MHz clock pulse, and a ripple-through counter, which divides down the 4-MHz clock to lower clock frequencies that are used throughout the logic module assembly. The board also produces the pulse train used to modulate the 150-MHz BIT oscillator. See figures 4-40 and 4-41.

The function of the flag board is to determine if sufficient information is being received to permit tracking and to supply the following output signals:

- Azimuth flag
- Elevation flag

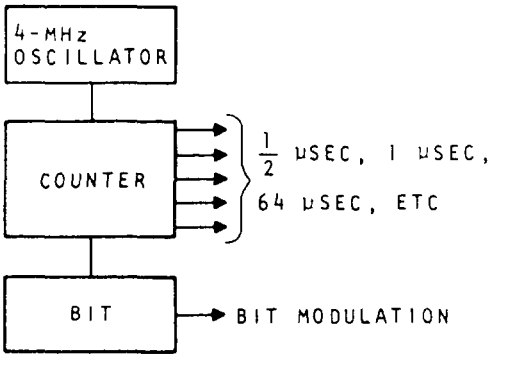


Figure 4-40.-Clock/BIT board.

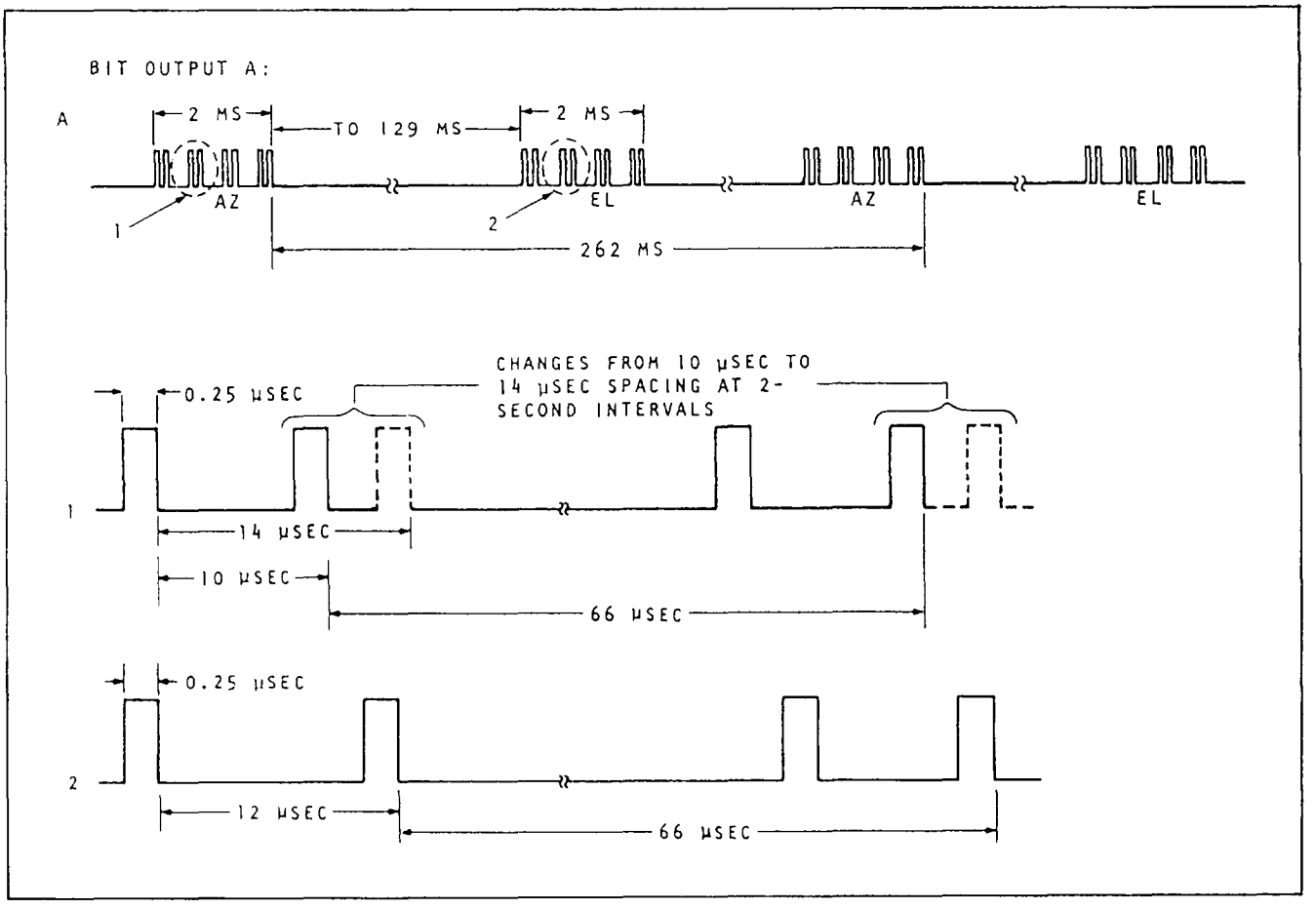


Figure 4-41.-BIT pulse spacing.

- Azimuth side-lobe AGC
- Elevation side-lobe AGC
- 16/beam count (a count of 16 pulses per beam)
- Azimuth 2 count (a count of two azimuth video pulses)
- Elevation 2 count (a count of two elevation video pulses)
- 47 count (a count of 47 azimuth/elevation video pulses)

The 16/beam circuit counts the read gate pulses. If 16 read gate pulses in one beam are received, the information within that beam is considered valid and a signal is sent to the azimuth flag circuit or to the elevation flag circuit, depending on which beam gate has been generated. If three beams having more than 16 pulses per beam are counted in the azimuth flag circuit, the 3-counter in the azimuth flag circuit sets the azimuth flag signal to the high level. The azimuth flag circuit also contains a counter that serves two functions. The first is to determine if the 16 pulses per beam received by the azimuth flag circuit are within 1 sec of each other. If they are not, the 3-counter is reset, and counting starts again from zero. The other function is to change the azimuth flag signal from a high to a low level, if no valid beams are received within 2.1 seconds. The elevation flag circuit is identical to the azimuth flag circuit. The video counter counts $\overline{Az + El\ VIDEO}$ pulses. If two pulses are counted, an output pulse is generated to put the AGC buss into a discharge mode. If 47 pulses are counted, an output pulse is generated. This pulse goes to the wide beam/narrow beam control located on the AGC I board. If 144 pulses are counted (too many for normal tracking), an azimuth side-lobe AGC or an elevation side-lobe AGC signal is generated. This signal is sent to the AGC board and is used to raise the azimuth or elevation AGC voltage, thereby reducing receiver sensitivity. The 144 count also makes the flags on the crosspointer indicator visible. The video counter is

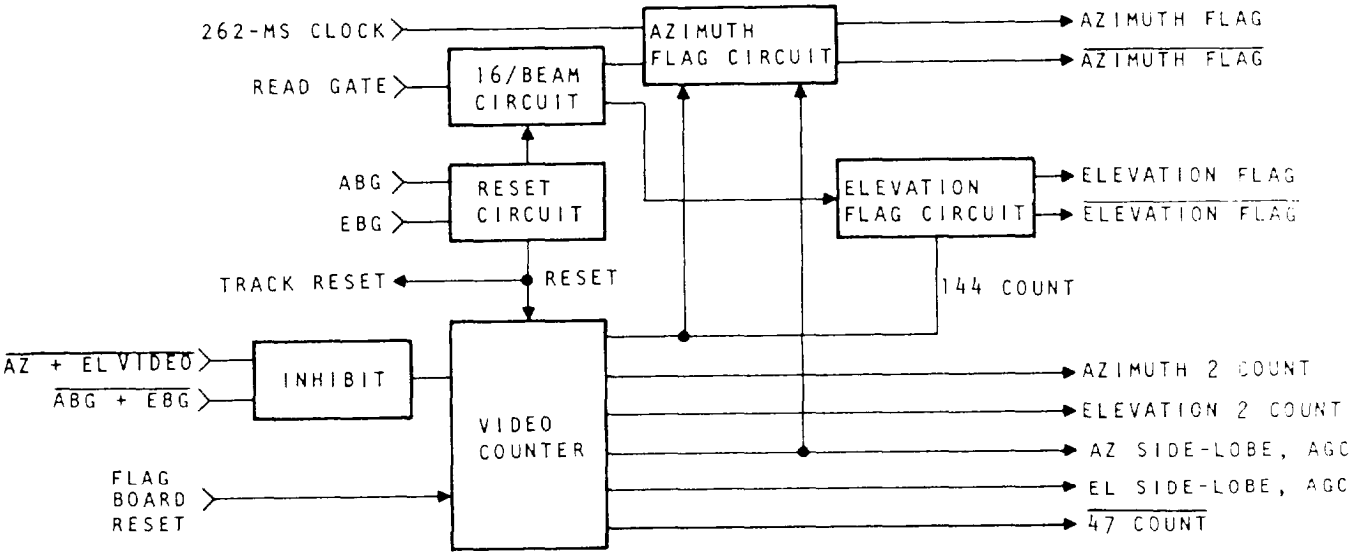


Figure 4-42.-Flag board.

enabled by the presence of an $\overline{ABG + EBG}$ signal. See figure 4-42.

TACTICAL AIR NAVIGATION (TACAN) SYSTEM

Learning Objective: *Explain a TACAN system to include receiver and transmitter circuits.*

TACAN is a polar coordinate navigation system. It is used to determine the relative bearing and/or slant-range distance to a TACAN ground station or cooperating aircraft. Therefore, only distance data is supplied to the interrogating aircraft. The TACAN system operating range limit is line-of-sight and depends on aircraft altitude and type of terrain.

The TACAN system operates on a selected channel from the 252 TACAN channels available. The 252 channels are equally divided: 126 X-channels and 126 Y-channels. Both X- and Y-channels are spaced at 1-MHz intervals. The TACAN channels provide airborne transmit (interrogation) frequencies from 1025 to 1150 MHz. The TACAN airborne receive frequencies are from 962 to 1213 MHz.

The AN/ARN-118(V) TACAN system operates in the following ground-to-air modes: receive (REC) and transmit-receive (T/R). It also operates in air-to-air modes: air-to-air transmit-receive (A/A T/R), air-to-air receive (A/A REC), and self-test (in-flight confidence test). The system produces both digital and analog data that can drive digital HSI, analog HSI, and computer systems. It can also drive bearing distance-heading indicators (BDHIs), course deviation indicators (CDIs), distance indicators, and radio magnetic indicators (RMIs). A TACAN beacon, either surface or airborne, is required for operation of a TACAN system.

RECEIVER SECTION

The binary-coded decimal (BCD) channel data and the high/low band (low level = low band) signal from the input control circuit is supplied to synthesizer A15A2. See figure 4-43. The BCD channel data is used to produce the voltage-controlled oscillator (VCO) voltage that is supplied to VVO A15A3. The high/low-band signal and the BCD channel data are used to produce a digital-to-analog output that is supplied to curve shaper A14A4. The synthesizer produces a specific control voltage and digital-to-analog output for the selected channel.

The control voltage supplied to VCO A15A3 causes the VCO to produce a fundamental RF frequency. The frequency is for the selected channel that is between 256.25 and 287.50 MHz. The fundamental RF frequency is supplied to RF driver A15A4.

RF driver A15A4 multiplies and amplifies the fundamental frequency to produce a local oscillator frequency between 1025 and 1150 MHz. The specific frequency depends on the selected TACAN channel. The local oscillator signal is supplied to RF amplifier/mixer A14A2. Curve shaper A14A4 uses the digital-to-analog output from A15A2 to produce a tuning voltage that is supplied to preselector A14A1. The tuning voltage produced by A14A4 varies nonlinearly with channel selection, but is a specific voltage for each selected channel.

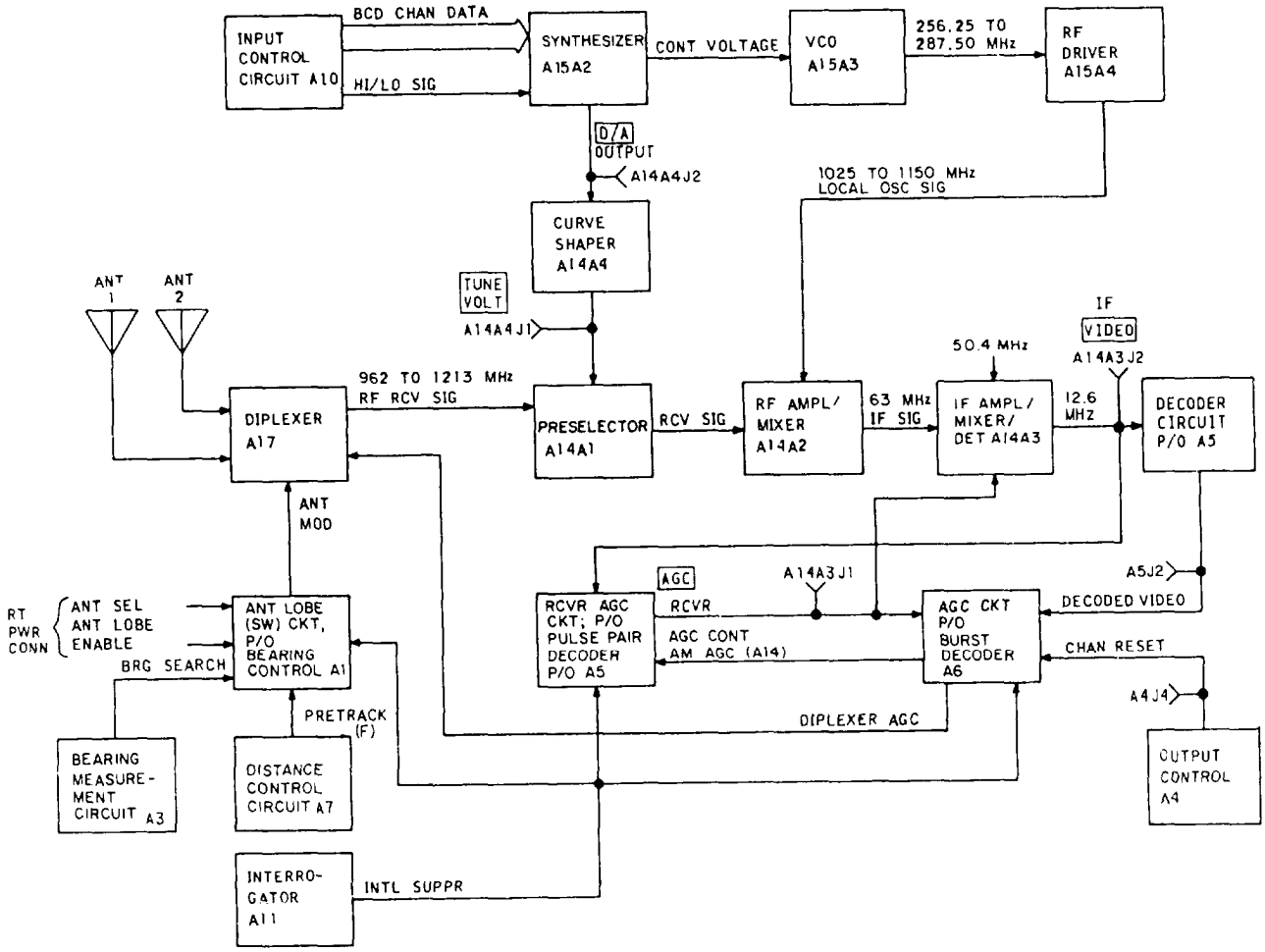


Figure 4-43.-Receiver section operation.

The external antenna select and antenna lobe enable signals are supplied to the antenna lobe (switching) circuit on bearing control Al. The bearing search signal and the pretrack signal are supplied to the antenna switching circuit. The antenna select signal allows either antenna 1 or antenna 2 to be manually selected. The antenna lobe enable signal allows the transmit and receive signals to be switched between antennas 1 and 2.

The antenna switching takes place at 5-second intervals until a usable signal is located. The bearing search and/or pretrack signals inhibit the switching when the bearing circuits are receiving a usable bearing signal. They also inhibit the switching when the distance circuits switch to the pretrack submode.

When switching is inhibited, the receive and transmit signals arc received from the antenna that provides a usable receive signal. The antenna modulation (switching) signal is applied from the antenna switching circuit to diplexer A17. The internal suppressor signal from A11 is supplied to the antenna switching circuit. The suppressor inhibits antenna switching during the transmission of the interrogation pulses.

Diplexer A17 receives the TACAN station signal from either antenna 1 or antenna 2, depending on the antenna modulation signal from the antenna switching circuit. The 962- to 12 13-MHz receive signal is supplied from diplexer A17 to preselector A14A1. The preselector circuit is composed of one 2-pole preselector and one 4-pole preselector. The preselectors are broadband filters that pass frequencies in the spectrum of the selected TACAN station channel. The tuning voltage from curve shaper A14A4 controls the selection of the frequency spectrum.

The receive signal is applied from the preselector circuit to RF amplifier mixer A14A2. The RF amplifier is actually located between the two preselectors, and is

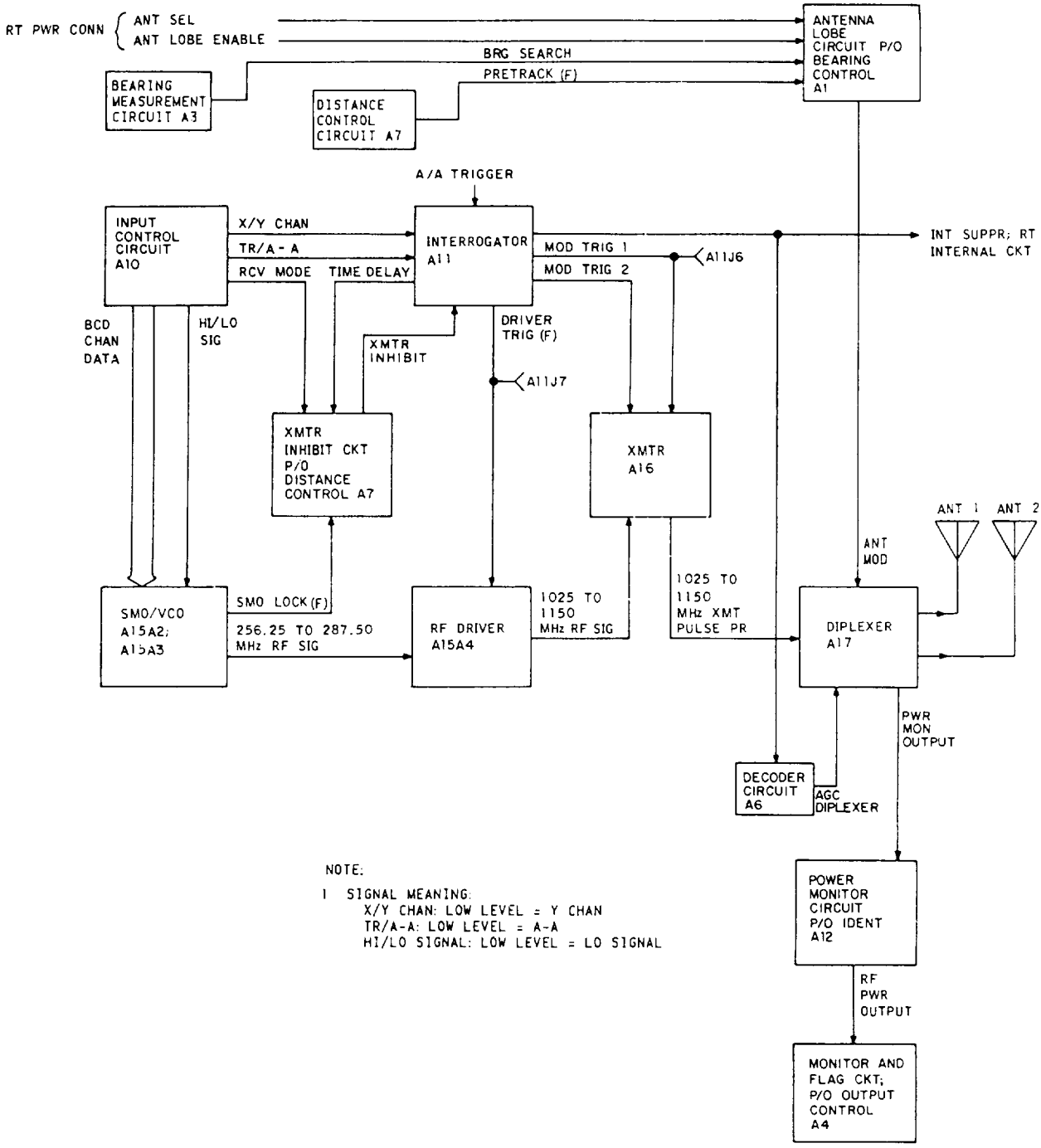


Figure 4-44.-Transmitter section operation.

used to amplify the output of the 2-pole preselect. The amplification takes place before the signal is supplied to the 4-pole preselector. The mixer portion of A14A2 heterodynes the receive signal with the 1025- to 1150-MHz local oscillator signal from RF driver A15A1.

A 63-MHz IF signal is then produced. The 63-MHz signal is supplied to IF amplifier A14A3. After amplification and application of AGC, the 63-MHz IF is converted to a 12.6-MHz second IF. Then, the 12.6-MHz IF is amplified and detected to form the IF video signal that is supplied to the decoder circuit and to the AGC receiver circuit on pulse pair decoder A5.

The decode circuits and the receiver AGC circuit on A5 process the IF signal. This produces an AGC receiver signal that is supplied to the diplexer AGC circuit on

burst decoder A6 and to IF amplifier A14A3. The decoded video signal from the decoder circuit is supplied to the AGC circuits on burst decoder A6. The AGC circuits on A6 determine the number of decoded video pulses received per second. The circuits also determine when internal suppressor pulses occur and when the channel is reset. The AGC circuit produces an AGC control voltage that is supplied to the receiver AGC circuit.

When over 680 decoded video pulses are received per second or an internal suppressor pulse occurs, the control voltage changes. The voltage changes the AGC receiver signal so that the IF amplifier is reduced to minimum gain. When a channel reset signal is received from output control A4 (new TACAN channel), the AGC circuit on A6 produces an AGC control voltage. This causes the AGC receiver signal to increase the IF amplifier gain to maximum for acquiring the new TACAN station receive signal. The AGC circuits on A6 produce a diplexer AGC signal that is supplied to A17, and controls the diplexer output. When an internal suppressor pulse occurs, the diplexer AGC signal inhibits the receive signal from being supplied to the preselector circuit.

TRANSMITTER SECTION

The transmitter is used only in the transmit-receive and air-to-air transmit-receive modes of operation. See figure 4-44.

The BCD channel data and high/low signal from the input control circuit contains selected channel data supplied to stabilized master oscillator/voltage-controlled oscillator (SMO/VCO) A15A2 and A15A3. The channel data tunes the SMO/VCO so that the fundamental frequency is produced for the selected channel. The SMO/VCO produces a 256.25- to 287.50-MHz fundamental frequency that is supplied to RF driver A15A4.

The X/Y channel data from the input control circuit and the transmit-receive/air-to-air mode signal are supplied to interrogator All. The X/Y and transmit-receive/air-to-air signals cause A11 to produce driver trigger and modulator trigger pulse pairs. The pulse pairs have the correct spacing between pulses for the selected channel and mode of operation.

The driver trigger pulse pairs are supplied to RF driver A15A4, and the modulator trigger pulses are supplied to transmitter A16. Interrogator A11 produces an internal suppression pulse during the time the driver and modulator trigger pulses are produced. The internal suppression pulse is supplied to the internal receiver-transmitter (RT) circuits. This is done to prevent the transmit interrogation pulse pairs from being applied to the receiver section and the decoder circuits.

RF driver A15A4 multiplies and amplifies the fundamental frequency from the SMO/VCO circuit. This produces a 1025- to 1150-MHz pulse pair during the time that the driver trigger pulse pair occurs. The specific frequency of the pulse pair depends on the channel selected. The 1025- to 1150-MHz pulse pair is supplied to transmitter A16 as the RF driver signal.

Transmitter A16 amplifies and modulates the 1025- to 1150-MHz RF driver signal. This produces a 1025- to 1150-MHz transmit pulse pair that is supplied to diplexer A17.

The antenna lobe circuit operates the same as described in the receiver section discussion. The antenna select signal manually selects antenna 1 or 2. Antenna lobing circuits switch between antennas 1 and 2 at 5-second intervals until a usable signal is located. The diplexer directs the transmit pulse pairs to the correct antenna. The internal suppression signal from A11 is applied to the decode circuits. This causes the decode circuits to produce an AGC diplexer signal that is supplied to A17. The AGC diplexer signal switches A17 to prevent the transmit pulse pair from being applied to the receiver section.

The transmit pulse pair power is sampled and supplied to the power monitor circuit on A12 as a power monitor output signal. The monitor circuit checks that the transmit pulse pair power is above a predetermined minimum. It also supplies the RF power output signal to the monitor and flag circuit on output control A4. If the transmitter power is insufficient, the RF power output signal from A12 causes A4 to produce a flag warning. When the RT is initially turned on, a nominal 90-second time-delay signal is supplied to the transmitter inhibit circuit on distance control A7.

The transmitter inhibit circuit applies a transmitter inhibit signal to A11 that causes the modulator and driver trigger pulses to be inhibited. The transmitter pulse pair is inhibited for the 90-second period to allow the transmitter to warm up. When the receive mode or air-to-air receive mode is selected, a receive mode signal is applied to the transmitter inhibit circuit.

The inhibit circuit produces an inhibit signal that keeps the modulator trigger pulses inhibited on A11. If the SMO does not lock on the selected channel frequency, the SMO lock signal causes an inhibit signal from A7. The modulator trigger pulses are inhibited until

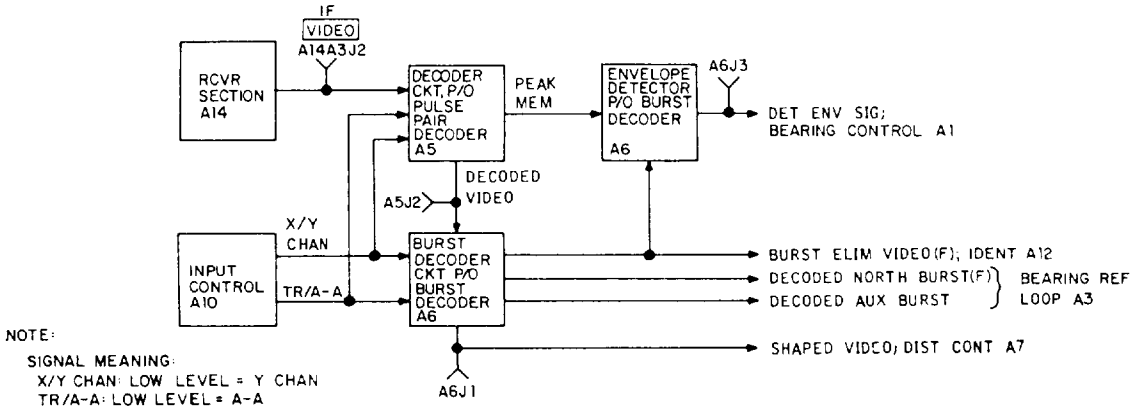


Figure 4-45.-Decoder circuit operation.

the SMO locks on the correct frequency. The SMO lock signal prevents the transmitter from transmitting on an incorrect frequency while the SMO is locking on the correct frequency.

DECODER CIRCUIT

The decoder circuits decode the IF video signal to produce a shaped video signal and a detected envelope signal. They also produce a decoded north burst pulse, a decoded auxiliary burst pulse, and a burst eliminated video signal. See figure 4-45.

The IF video signal from the receiver section is supplied to the decode circuits on pulse pair decoder A5. The X/Y channel and the transmit-receive/air-to-air mode signals from the input control circuit are supplied to A5. They are also supplied to burst decoder A6.

The signals from the input control circuit switch the decode circuits to decode only the data for the selected channel and mode of operation.

The decode circuits on A5 produce a decoded video signal. The signal is a pulse for each received pulse pair with the correct spacing for the selected channel and mode of operation. The decoded video signal is supplied to the burst decoder circuits on A6. The decode circuits also produce a peak memory signal that is the amplitude of each pulse received. The memory signal is supplied to the envelope detector circuit on A6.

The burst decode circuits on A6 process the decoded video to provide a shaped video signal with uniform width pulses. The shaped video signal is supplied to distance control A7. The decode circuit produces a decoded north burst pulse that is a bearing pulse that represents magnetic north. It also produces a decoded auxiliary burst pulse that is the bearing pulse that divides the 360-degree bearing circle into 40-degree sectors.

The decoded burst pulses are supplied to bearing reference loop A3 for calculating the aircraft relative bearing with respect to the TACAN station. A burst eliminated video signal is produced by the burst decode circuit and supplied to the envelope detector circuit. The signal is also supplied to the identification circuits on A12. The envelope detector circuit uses the peak memory signal and the burst eliminated video signal to produce a detected envelope signal that is applied to bearing control A1.

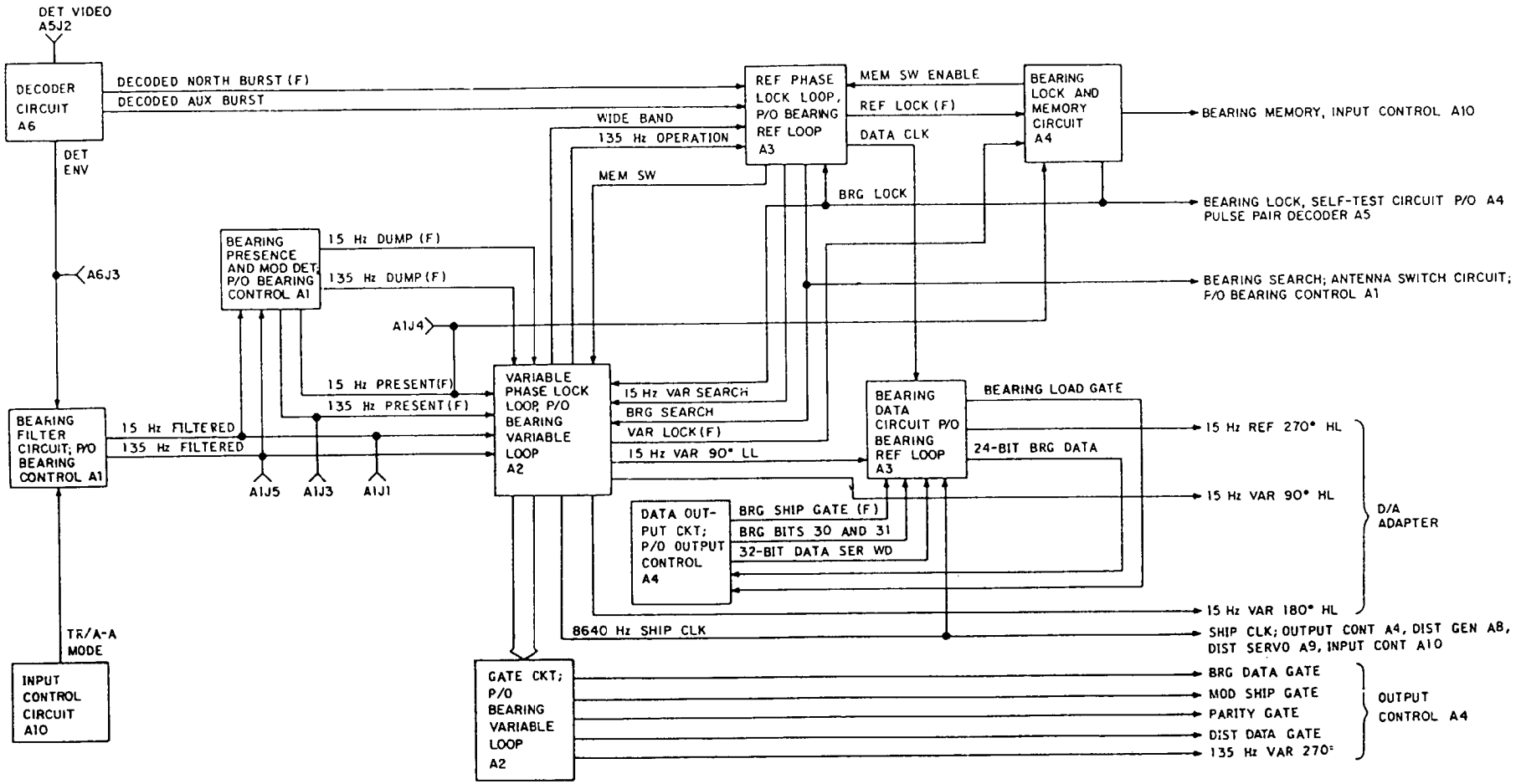
The detected envelope signal is a 135-Hz signal modulated with the 15-Hz signal. The 15-Hz portion of the signal is the coarse bearing data. The 135-Hz portion of the signal is the tine bearing signal. The burst eliminated video signal is used to produce the detected envelope signal, since the bursts would distort the envelope detector.

BEARING MEASUREMENT CIRCUIT

The bearing measurement circuit calculates the aircraft relative bearing with respect to the selected TACAN station. The data is calculated from the detected envelope signal and the north and auxiliary burst pulses from the decode circuit. The bearing data is supplied to the output control circuit as digital data. See figure 4-46.

The detected envelope signal from the decode circuit is supplied to the bearing filter circuit on bearing control A1. The transmit-receive/air-to-air mode signal from the input control circuit is also supplied to the bearing filter circuit.

The filter circuit separates and filters the detected envelope signal. This produces 15- and 135-Hz filtered



NOTE:

(1) LL MEANS LOW LEVEL (+3.5 -5 V DC FOR LOGIC 1)
HL MEANS HIGH LEVEL (> +10.0 V DC FOR LOGIC 1)
TR/A-A LOW LEVEL = A-A

Figure 4-46.–Bearing measurement circuit operation.

signals that are supplied to the bearing presence on A1. They are also supplied to the modulation detector circuit on A1 and to the variable phase-lock loop circuit on bearing variable loop AZ. The 135-Hz signal is inhibited in the air-to-air mode.

The bearing presence and modulation detector circuit checks the 15- and 135-Hz faltered bearing signals. This determines if the modulation of the signals exceeds a predetermined minimum. It also determines if a sufficient signal is present for correct bearing calculations. The bearing presence and modulation detector circuit supplies 15- and 135-Hz present signals and 15- and 135-Hz dump signals to AZ.

The present signals are supplied to A2 when the bearing signals are usable. When the signals are not usable, the detector circuit applies 15- and 135-Hz dump signals to A2. This inhibits the 15- and 135-Hz filtered bearing signals from being applied to the phase-lock loop circuit on A2. When the air-to-air mode of operation is selected, the 135-Hz presence detector is forced to indicate that 135-Hz data is not available. The 135-Hz dump signal is then produced.

The bearing variable phase-lock loop circuit on A2 is a 90-degree phase-lock loop that is locked when the output of the loop is 90 degrees out of phase with the input. The loop locks with three different inputs. Before a usable bearing signal is received, a bearing search signal from A3 is applied to the variable phase-leek loop. The 15-Hz variable search signal from the reference phase-lock loop circuit is used to lock the variable phase-lock loop.

The bearing search signal is supplied to the antenna switching circuit on A1. This produces antenna lobing until a usable signal is obtained. When a usable bearing signal is received, the variable phase-lock loop locks on the 15-Hz coarse bearing signal first. During the 15-Hz lock-on, the variable phase-lock loop applies a wide-band signal to the reference phase-lock loop. This switches the loop to a wide passband for locking on the decoded north burst from the decode circuit. The bearing data during and after lock-on is supplied to the bearing data circuit on A3 as a 15-Hz variable 90-degree signal.

When the variable loop is locked on the 15-Hz signal, a variable lock signal is supplied to the bearing lock. The lock is also supplied to a memory circuit on output control A4. When the reference loop locks on the decoded north burst, a reference lock signal is supplied. The lock signal is supplied to the bearing lock and memory circuit. After both loops are locked, the bearing lock and memory circuit supplies a bearing lock signal to the variable and reference loops. The lock signal is also supplied to the self-test circuit on A4.

The bearing lock signal switches the variable loop to lock on to the 135-Hz (fine) bearing signal. The variable loop supplies a 135-Hz operation signal to the reference loop. This signal causes the reference loop to lock on the decoded auxiliary burst. The lock-onto the 135-Hz bearing signal and the decoded auxiliary burst finely adjusts the phase-lock loop circuit. This adjustment provides an accurate bearing calculation. The bearing data is still supplied to the bearing data circuit by the 15-Hz variable 90-degree signal.

The bearing lock and memory circuit uses the variable lock signal, the reference lock signal, and the 15-Hz present signal to control the memory circuit. If the 15-Hz present signal is lost or if either of the phase-lock loops unlock the memory circuit applies a memory switch enable signal to the reference phase-lock loop. The reference loop supplies a memory switch signal to the variable phase-lock loop.

The memory signals cause both loops to remain locked on the last valid bearing data. The memory signals last for a nominal 3 seconds after a lock or present signal is lost. The memory circuit supplies a bearing memory signal that indicates when the bearing circuits are in bearing memory. The input control circuits on A10 receive this indication. The input control circuit incorporates the bearing memory data in the frequency word, which provides the area navigation system with bearing memory data.

The variable phase-lock loop contains a frequency divider that provides specific frequency signals to the gate circuit. The gate circuit is on A2 and produces an 8640-Hz ship clock that is used throughout the RT. The 8640-Hz ship clock is used to ship serial data between circuits and to ship the data out of the RT. The gate circuit produces a bearing data gate, a modified ship gate, a parity gate, a distance data gate, and a 135-Hz, variable, 270-degree signal. All these signals go to output control A4.

The bearing data and distance data gates are used to determine the status of the bearing and distance data. This information is shipped out of the RT. The parity gate is used to check the parity of the bearing and velocity serial words. The modified ship gate is used as the word synchronizer pulse for the 6-wire serial data shipped out of the RT. The 135-Hz, variable, 270-degree signal is used to clock a label circuit. The circuit produces labels for the serial words shipped out of the RT.

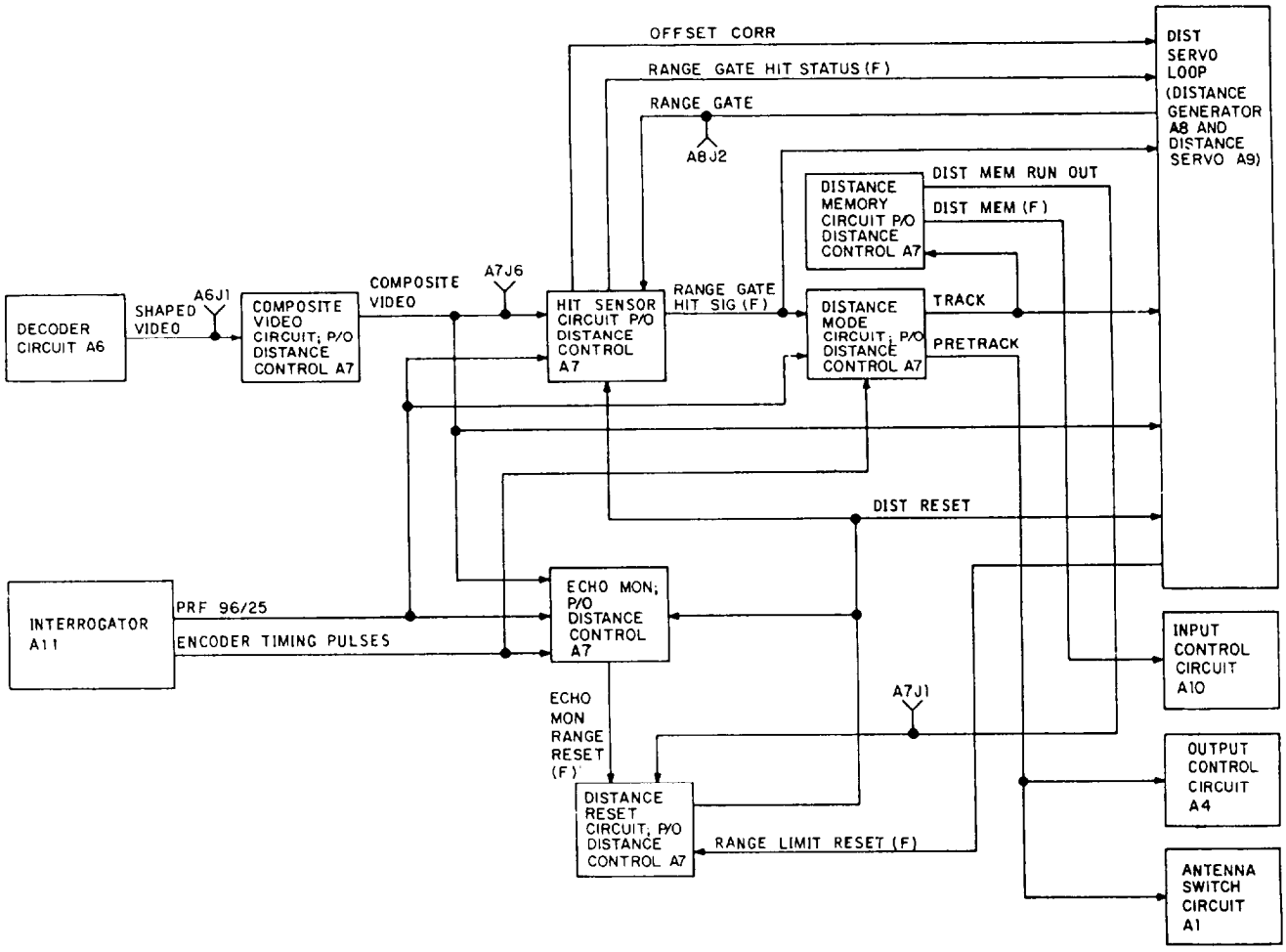


Figure 4-47.-Distance control circuit operation.

When the 15-Hz variable 90-degree signal makes a positive zero crossing, the bearing data circuit loads a shift register. This is done by a continuous running counter that is clocked by a data clock signal from the reference phase-lock loop. Bearing status bits 30 and 31 are applied from output control A4 to the shift register in the bearing data circuit. Bits 30 and 31 are added to bearing data.

The shift register stores 24 bits of data. Bearing data is in bits 18 through 29, status data in bits 30 and 31, and bit 32 is a logic 0 parity bit. The parity bit logic can be changed when the parity is checked in the data output circuits.

When the bearing ship gate is supplied to the bearing data circuit, the 24-bit serial bearing data is sent to A4 at the 8640-Hz ship clock rate. When the 32-bit bearing word is sent out of the RT, the word is supplied back to the shift register in A4. The word reloads the shift register with the same data that was sent out of the register. The register must be reloaded with the same data because it is sent out faster than it is updated.

The bearing data circuit produces a bearing load gate when the bearing data is loaded into the shift register. The load gate is supplied to output control A4 for use in the bearing status circuit.

DISTANCE CONTROL CIRCUIT

The distance control circuit is part of the distance measurement circuits. Due to the complexity of the distance measurement circuits, the distance control circuit is discussed separately. The distance control circuit determines the mode of operation for the distance measurement circuits. It monitors the distance reply pulse to determine if the distance circuits are locked on the reply pulse or an echo pulse. See figure 4-47.

The shaped video signal from the decoder circuit is supplied to the composite video circuit on A7. The composite video circuit supplies the composite video

pulses to the hit sensor circuit, the echo monitor circuit, and the distance servo loop.

The range gate from the distance servo loop is supplied to the hit sensor circuit. The hit sensor circuit determines when hits occur within the range gate. This indicates that the distance circuits have located the distance reply target. When a hit occurs in the range gate, the hit sensor circuit supplies a range hit signal. This hit signal is supplied to the distance servo loop and to the distance mode circuit.

When hits do not occur in the range gate, the hit sensor circuit applies an offset correction signal to the distance servo loop. This causes the servo loop to move the range gate in search of the distance reply target. The hit sensor circuit is timed by a PRF pulse from interrogator A11. The pulse occurs at the beginning of each interrogation period. The hit sensor circuit looks for a hit in the range gate during each interrogation period.

The PRF pulse and encoder timing pulses are supplied to the distance mode circuit from A11. The distance mode circuit keeps the distance circuits in the search submode. The circuits remain in this mode until at least seven hits occur in the range gate within 15 consecutive PRF periods. When the seven hits occur, the distance mode circuit switches the distance circuits to the pretrack submode. A pretrack signal is then supplied to the output control circuits and to the antenna switching circuit.

The pretrack signal applied to the antenna switching circuit inhibits antenna lobing. This keeps the RT signals connected to the last antenna used. The pretrack signal supplied to the output control circuit enables the automatic self-test function in the self-test circuits.

After a nominal 7 seconds, if hits continue to occur in the range gate, the distance mode circuit switches to track submode. A track signal is supplied to the distance servo loop circuit and to the distance memory circuit. The track signal enables the memory circuit to keep the distance circuits in memory if the distance signal is lost.

If hits stop occurring in the range gate and the distance circuits are in the track submode, the distance memory circuit locks the distance circuits. The circuits are locked with the last valid distance data, but the memory circuit updates this with the last known velocity data for a nominal 15 seconds. At the end of the 15-second period, a distance memory runout signal is applied to the distance reset circuit. This causes the reset circuit to supply a distance reset signal to the distance servo loop circuit. The reset signal is also supplied to the hit sensor circuit and to the echo monitor circuit.

The distance reset signal resets the distance circuits to the search submode. A distance memory signal is supplied to the input control circuit. The memory signal provides the memory data used in the frequency word. The frequency word is supplied to the area navigation system and indicates the distance memory status.

The echo monitor circuit determines if the distance circuits are locked on an echo pulse instead of the distance reply target. The circuit searches outbound from 0 to 300 nmi or the range gate, whichever occurs first. An echo pulse occurs outbound of a distance reply pulse. This occurs because of delayed interrogation of the TACAN station by the reflected interrogation pulse.

If the echo monitor locates a distance reply pulse inbound of the pulse that the distance circuits are locked on and tracking, the distance circuits are locked on an echo pulse. When this occurs, an echo monitor reset pulse is supplied to the distance reset circuit. A distance reset signal is produced and resets the distance circuits to the search mode to look for the distance reply target.

The maximum range of the distance circuit is 389.9 nmi. When the maximum range is reached and exceeded, a range limit reset signal is supplied from the distance servo loop circuit to the distance reset circuit. The distance reset circuit resets the distance circuits to the search submode of operation.

DISTANCE SERVO LOOP

The distance servo loop uses the range reply (target) pulse contained in the composite video signal. The video signal from the distance control circuit is used to calculate the distance (range) to or from the selected TACAN station. It also calculates the velocity (range rate) at which the aircraft is approaching or flying away from the station. The servo loop supplies the range and range rate data to the output circuit as 24-bit serial words. The range data is also supplied to the output circuit as an analog distance pulse pair. See figure 4-48.

The composite video signal consists of squitter and target pulses. The offset correction signal indicates when hits are not occurring in the range gate. The range gate hit status signal indicates when hits are occurring in the range gate. These three signals are supplied to the

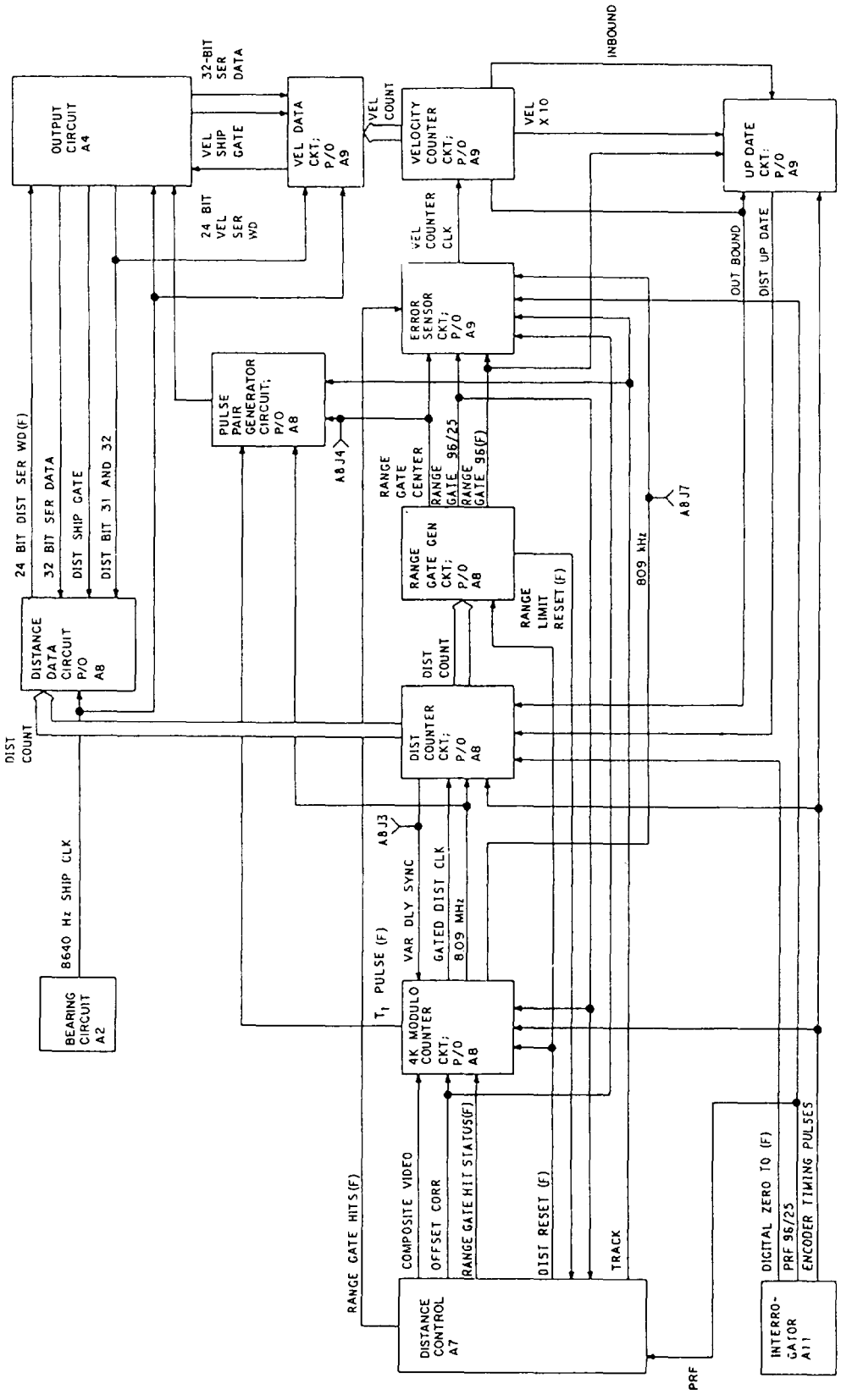


Figure 4-48.–Distance servo loop operation.

4K modulo counter circuit. Encoder timing pulses from interrogator A11 and the range gate pulse from the range gate generator circuit are supplied to the counter circuit.

At the beginning of each interrogation period, an encoder timing pulse presets the 4K modulo counter to a count of 3958. The counter is clocked at a 0.1-nmi rate and produces a gated distance clock signal that is supplied to the distance counter circuit. The 4K counter operates in the search, pretrack and track submodes. The first 42 counts, or from 3958 to 4000 (0), are used to produce gate pulses. These pulses are for distance circuit operation and a T1 pulse that is supplied to the pulse pair generator circuit.

The offset correction and range gate hit status signals control the operation of the gated distance clock signal. When the distance circuits are in the search submode, the distance circuits search outbound. The search starts from 0 nmi to locate a range reply target pulse.

The offset correction signal causes the 4K modulo counter circuit to produce four up-distance clocks and a 4K modulo down-distance clock. The down-distance clock is inhibited from the trailing edge of the range gate to the next composite video pulse. This can be either a squitter or range reply composite video pulse. This causes the range gate to be moved outbound in search of the range reply target pulse.

When a hit occurs in the range gate, the range gate hit status signal from the distance control circuit inhibits the four up-distance clock pulses. During the prior interrogation period, the four up-distance clock pulses were used to center the range reply pulse in the range gate. After the range gate hit status signal occurs, only the down-distance clock is supplied to the distance counter circuit. If the hit in the range gate is a squitter pulse and not a range reply pulse, the pulse does not occur in the same position. The distance circuit again supplies the offset correction signal to the 4K modulo counter circuit. Finally, the distance circuits search for another target pulse outbound.

The distance reset signal occurs when the distance memory runs out or when the 389.9-nmi distance range limit is reached. It also occurs when the distance circuits are locked on an echo pulse. The reset signal is then supplied to the distance counter circuit. The distance reset signal presets the distance counter to 399.00 nmi. On initial turn on, the distance counter is preset to 399.00 nmi. The counter is preset by a timing pulse from A11, which occurs at the beginning of an interrogation period.

Either the distance reset or the timing pulse causes the distance circuits to be placed in the search submode. The search submode requires the distance counter to be set to 399.00 nmi. The distance counter is counted up four 0.1 counts and down 4000 counts each interrogation period. The only time it does not count is the period from the trailing edge of the range gate to the next composite video pulse. The 4K modulo down-distance clock is inhibited. The inhibiting of the down-distance clock causes the range gate to move outbound from the previous position to the next composite video pulse.

The distance count produced by the distance counter circuit is supplied to the distance data circuit and to the range gate generator circuit. When the range reply hit occurs in the range gate and the distance circuits are in the track submode, the distance counter is counted down 4000 clock pulses. This causes the counter to count down from the number stored in the distance counter through zero to the same number again each interrogation period.

The digital zero signal from interrogator A 11 is applied to the distance counter circuit. The digital zero signal synchronizes the counter to digital zero, which is adjustable to allow for internal delays of the receive signal. The outbound from velocity counter and distance update signals from the update circuit are supplied to the distance counter circuit. The distance update signal corrects the counter in 0.01-nmi increments for the range rate.

The outbound signal is used to count the distance counter up or down in 0.01-nmi increments. The direction, up or down, depends on whether the aircraft is flying to or from the station. The distance counter circuit produces a variable synchronizer signal from the 0.01-nmi data. It applies the signal to the 4K modulo counter circuit. This synchronizes the start of the 4K modulo counter with the 0.01 nmi, since the counter is clocked in 0.1-nmi increments.

The range gate generator circuit produces a range gate 96 that is 0.8 nmi wide during each interrogation period in the search mode. The range gate is produced when the distance counter output is being counted down from 3987 to 3979 during each interrogation period. Therefore, the range gate is always produced at the same time in the distance count. The down-distance clock is inhibited during the search submode for the period from the trailing edge of the previous range gate to the next composite video pulse. The counter does not countdown a full cycle. This causes the range gate to be produced outbound of the previous range gate.

The range gate is supplied to the error sensor circuit and the update circuit on distance servo A9. Range gate 96/25 is supplied to the error sensor, distance control, and the 4K modulo circuits. Range gate 96/25 is 96 Hz in the search submode and 25 Hz in the subtrack mode. The range gate generator circuit produces a range gate center pulse that is supplied to the error sensor circuit. The pulse is also supplied to the pulse pair generator circuit.

The range gate center pulse leading edge occurs in the center of the range gate. The trailing edge occurs at the same time as the trailing edge of the range gate. The range gate generator circuit senses when the distance circuits reach their maximum range of 389.9 nmi. A range limit reset signal is then produced. This signal is supplied to the distance control circuit. This resets the distance circuits to the search submode.

The track signal indicates when the distance circuits are in the track submode. The signal is applied from the distance control circuit to the pulse pair generator circuit. When the distance circuits are in the track submode, the pulse pair generator circuit uses the T1 pulse from the 4K modulo counter circuit. The generator uses the pulse as the first pulse of the pulse pair produced by the circuit.

The pulse pair generator circuit produces a pulse pair. The time between pulses is equal to 50 sec + [12.359 sec (radar mile) x distance (nmi)]. The T1 pulse occurs at 50 sec before zero nmi, and is the first pulse of the pulse pair. The range gate center (actual distance to the station) is used to produce the second pulse of the pulse pair. The pulse pair is supplied to the output control circuit.

The offset correction and track signals from the distance control circuit are supplied to the error sensor circuit. The offset correction signal inhibits the circuit until the track signal is supplied to the sensor circuit. The track signal enables the sensor circuit to check the position of the range gate hit in the range gate. The sensor circuit produces a no-error velocity counter clock signal. The signal is supplied to the velocity counter circuit when the hit is centered in the range gate.

When the hit is not centered in the range gate, the sensor determines if the error is small or large. If the error is large, the error sensor supplies a high-frequency clock signal to the velocity counter. If the error is small, the error sensor supplies a low-frequency clock signal to the velocity counter.

The velocity counter circuit produces a velocity count from 1 to 8,163 knots. The distance circuits can only track a range reply target at a maximum of 4,800 to 5,000 knots. The counter is clocked either up or down to the correct velocity. It remains set at that velocity until an error is sensed. The circuit produces a velocity times 10 signal. The signal is supplied to the update distance circuit. A velocity count signal is supplied to the velocity data circuit.

The update circuit converts the velocity times 10 signal to 0.01-nmi distance update data that is supplied to the distance counter circuit. The 0.01-nmi update data causes the distance counter to increase or decrease proportionately with the aircraft velocity. Therefore, during the search and pretrack submodes, the distance counter is clocked by the 4K modulo counter. After the distance circuits switch to the track submode, the distance counter is updated in 0.01-nmi increments by the distance update signal.

The 8640-Hz ship clock signal from the bearing circuits is supplied to the distance data, velocity data, and output circuits. The distance count from the distance counter circuit is supplied to the distance data circuit on distance generator A8. The distance ship gate and distance status bits are 31 and 32. The distance data circuit converts the distance count and status bits to a 24-bit serial word. This word is shipped to the output circuit during the distance ship gate.

The 24-bit word is shipped to the output circuit at the 8640-Hz ship clock rate. The output ships the 32-bit distance word out of the RT. The 32-bit serial word is also supplied to the distance data circuit as serial data when this happens. The distance data circuit does not use the first 8 bits of the word but stores the last 24 bits.

The last 24 bits of the serial word is the same data that was shipped out as the 24-bit distance serial word. The same data must be stored in the distance data circuit. The distance count signal does not update the distance data as often as the 24-bit serial word is shipped to the output circuit.

The velocity data circuit operates in the same way as the distance data circuit. The difference is that the velocity data is shipped to the output circuit. Also, the distance status bits are used to establish the status of the velocity data.

DATA OUTPUT CIRCUIT

The data output circuit converts the data from the bearing, distance, and input control circuits to serial data. The information is shipped out of the RT to

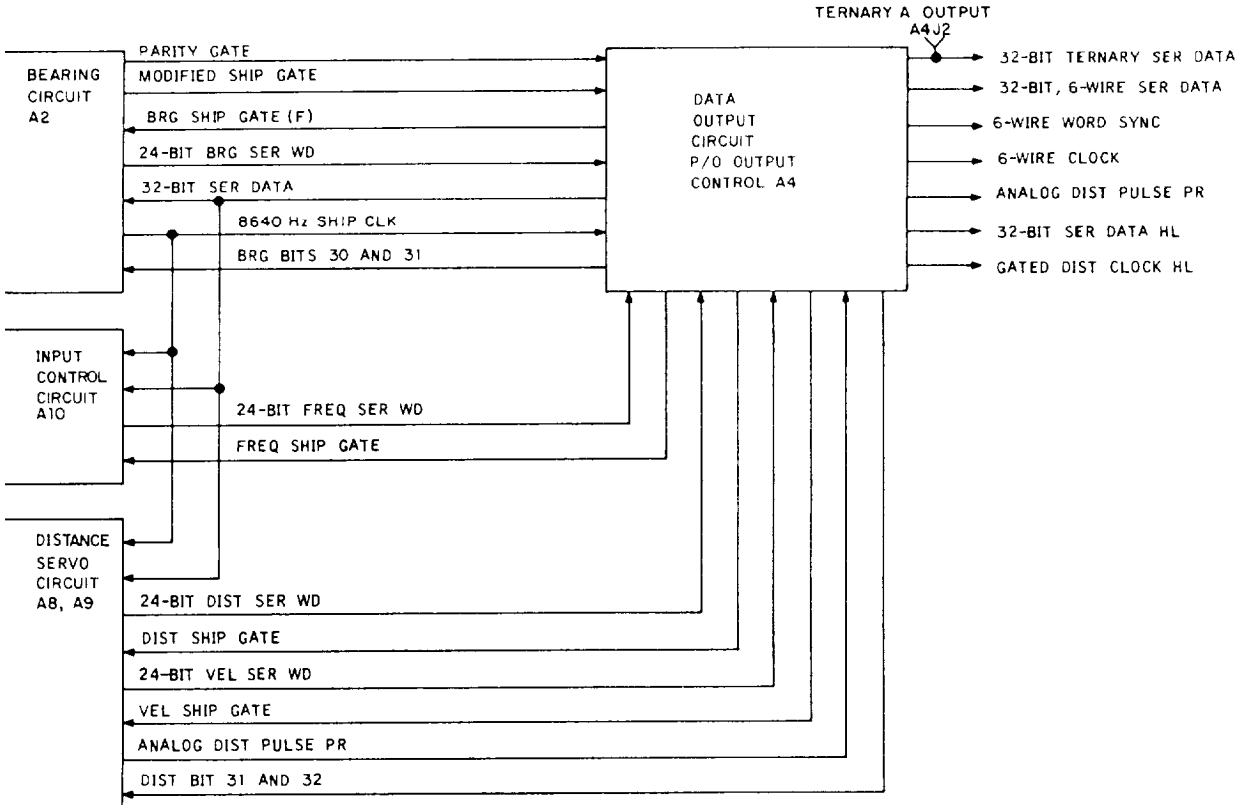


Figure 4-49.-Data output circuit operation.

indicators and/or other systems requiring the data. See figure 4-49.

The 24-bit bearing, frequency, distance, and velocity serial words are supplied to the data output on output control A4. The parity gate and modified ship gate pulses from the bearing circuits are supplied to the data output circuit. The output circuit uses the parity gate to produce odd 1's parity. This is used for the bearing and velocity serial words shipped out of the circuit. The output circuit produces the required label for each serial word and converts the serial word to ternary serial words. These words are supplied to indicators and/or systems requiring the data.

The ternary data is supplied to a ternary bus that carries the four different ternary words. The data is shipped out of the output circuit in 32-bit words separated by 32-bit spaces. The following information sequence follows: distance serial word, space, frequency serial word, space, bearing serial word, space, velocity serial word, and space. The cycle is then repeated.

The data is also shipped out of the RT as 6-wire serial data. The 6-wire data consists of 32-bit, 6-wire serial data; a 6-wire word synchronizer pulse; and 8640-Hz, 6-wire clock. The synchronizer pulse occurs as the eighth pulse of the 32-bit serial word. The 6-wire distance serial data is produced in the same sequence as the ternary data.

The data output circuit amplifies the analog pulse pair from the distance circuit. The output also supplies the data to any external system requiring the data.

The circuit produces serial data that is identical to the 6-wire serial word data. This data is supplied to the RT adapter as a 32-bit, serial-data, high-level signal. The distance data contained on the serial bus is the only data from the bus that is used by the adapter. The adapter also receives a gated distance clock high-level signal from the output circuit. The gated distance clock is an 8640-Hz, 32-bit clock that occurs only when the distance data is shipped from the RT.

When the serial data is shipped from the RT, the 32-bit serial data is shipped back to the bearing input control and to distance circuits for storing the data in the circuits. It is stored until the circuits update the data.

MONITOR AND FLAG CIRCUIT

The monitor and flag circuits check the validity of the data calculated in the bearing and distance circuits.

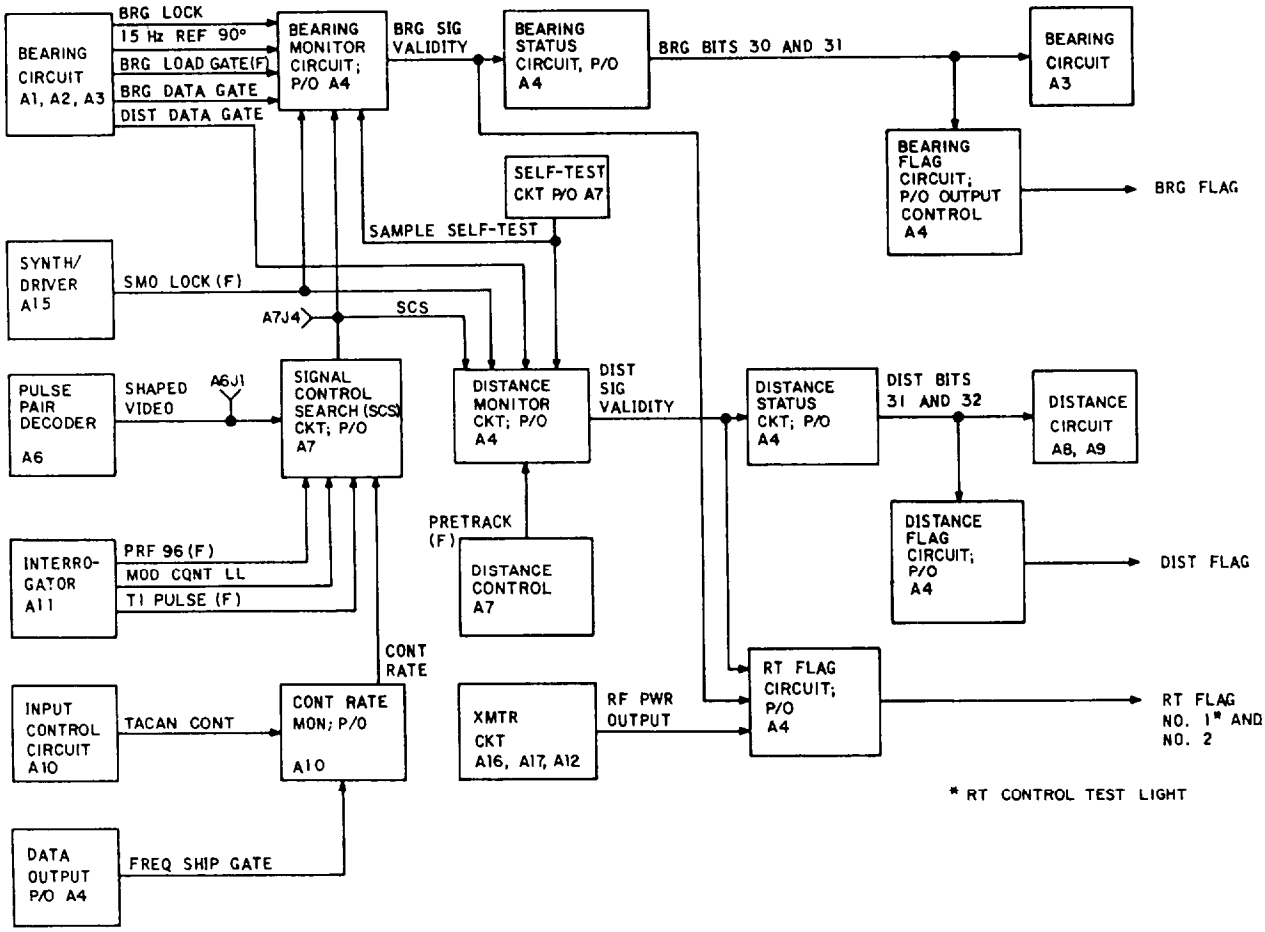


Figure 4-50.-Monitor and flag circuit operation.

They also monitor receiver-transmitter operation. See figure 4-50.

The TACAN control signal from the input control circuit is supplied to the control rate monitor circuit. It is supplied with the frequency ship gate pulse from the data output circuit. The control rate monitor is enabled by the TACAN control signal and operates only when the RT is controlled by an external system such as the radio-navigation (RNAV) system. The monitor checks that the control data is updated at a sufficient rate to ensure correct operation of the RT. The control rate signal is supplied to the signal control search (SCS) circuit.

The shaped video pulses from the decoder and the PRF, modulation control, low-level signal, and T_1 pulse from A11 are supplied to the SCS circuit. The circuit checks for at least six shaped video pulses each interrogation period, or between T_1 pulses. At least six pulses must be received each interrogation period to ensure accurate bearing and distance data. When at least six pulses are received, the circuit supplies a control rate (SCS) signal to the bearing monitor and distance monitor circuits.

If control data from the RNAV system is not being updated at a sufficient rate, the control rate signal that is applied to the monitor circuits is removed.

The bearing lock, 15-Hz reference 90-degree, bearing load gate, and bearing data gate signals are supplied to the bearing monitor. The SMO lock signal, the SCS signal, and the sample self-test signal are also supplied to the bearing monitor circuit.

When the bearing circuits are locked, the 15-Hz reference signal is present, the SMO is locked, and when the SCS signal is present, the bearing monitor circuit supplies a valid bearing signal to the bearing status and the RT flag circuits. The signals are supplied to the circuits when the sample self-test signal from the self-test circuit occurs.

The sample self-test signal occurs when the distance circuits switch from pretrack to track submodes. If any one of the input signals indicates a malfunction, the monitor circuit supplies an invalid signal to the status

and RT flag circuits. The bearing status circuit uses the bearing data from the monitor circuit to produce status bits 30 and31 of the 32-bit bearing serial word. When both bits 30 and 31 are logic 0, the bearing data is valid. Also, the bearing indication is in the 0- to 180-degree range.

When bit 30 is a logic 1 and bit 31 is a logic 0, the bearing data is invalid due to a failure. When bit 30 is a logic 0 and bit 31 is a logic 1, the status indicates that no computed data is available. When both bits 30 and 31 are logic 1, the bearing data is valid. The bearing indication is then in the 180- to 360-degree range.

Bearing status bits 30 and 31 are supplied to the bearing circuit and to the bearing flag circuit. The status bits become part of the 24-bit serial data. When the status bits indicate a failure or that no computed data is available, the bearing flag circuit produces a bearing flag signal. The signal is applied to the system indicator and causes the indicator flag to come in view.

The distance circuits operate in a manner similar to the bearing monitor and flag circuits, except the distance data gate and the pretrack signal are supplied to the monitor circuit. The pretrack signal is from the distance control circuit. When the distance circuits switch to track and all monitor inputs are normal, the monitor supplies a distance valid signal to the distance status and flag circuit. The status circuit produces distance bits 31 and 32 of the 32-bit distance serial word.

Bits 30 and 31 indicate the status of the distance data contained in the word. When both bits 31 and 32 are a logic 0, the distance data is valid. When bit 31 is a logic 1 and bit 32 is a logic 0, there is a failure in the distance circuits. When bit 31 is a logic 0 and bit 32 is a logic 1, no computed distance data is available. The status with bits 31 and 32 at a logic 1 level is not used Distance bits 31 and 32 are supplied to the distance circuits for insertion into the 24-bit distance serial data word

Distance bits 31 and 32 are supplied to the distance flag circuit. hey cause the flag circuit to produce a distance flag signal when a distance failure occurs or no computed data is available. The flag signal is supplied to a distance indicator and causes the indicator to pull the distance flag in view.

When a bearing or distance failure occurs or the transmitter power is below a predetermined minimum, the flag circuit produces indicator RT flag signals. The indicators warn the flight crew that the RT is malfunctioning.

RECEIVER-TRANSMITTER ADAPTER

Receiver-transmitter adapter MX-9577/A converts the digital outputs from the RT-1159/A to analog signals. They are converted for use by horizontal situation indicator (HSI), radio magnetic indicator (RMI), course deviation indicator (CDI), and bearing distance heading indicator (BDHI). However, during discussion of the adapter theory of operation, the HSI is used as the indicator receiving the outputs. See figure 4-51.

The distance data from the RT is applied to distance No. 1 card Al. The distance data, in serial binary-coded decimal (BCD) format, is converted to parallel format in the shift register. This is done with a serial-to-parallel bit converter. The 100's distance data is converted to analog synchro type voltages and applied to driver card A7. Card A7 is a power amplifier and the 100's distance data, in synchro format, is supplied to the HSI.

The 0.1's distance data is converted from BCD to sine/cosine format and applied to distance No. 2 card, A2. On A2, the 0. 1's and 1's distance data is combined. The sine/cosine 1's distance data output is applied to driver card A5. Card A5 converts the sine/cosine 1's data to analog synchro format that is supplied to the HSI.

The 10's and 1's distance data is processed similarly, and is converted and amplified by driver card A6. The Geneva switch causes the 1's and 10's indicators on the HSI to move at the same time when the 1's indicator is between 9 and 10.

The RMI bearing data and the HSI to-from and lateral deviation signals are derived from the 15-Hz inputs from the RT. Bearing No. 1 card, A3, converts the 15-Hz inputs to dc sine/cosine data that is applied to bearing No. 2 card, A4. The X-Y magnetic compass input is converted to sine/cosine data on A4, where it is combined with the TACAN bearing data. The 4-quadrant multipliers and summing network produce analog synchro X´- and Y´-outputs that are applied to card A8. Synchro driver power supply card A8 provides the RMI X and Y bearing outputs to the HSI.

The OBS signal generator (fig. 4-51) on card A3 produces the D-, E-, F-, and G-outputs for the course resolver in the HSI. The HSI NAV flag is controlled by the RT NAV flag input.

The OBS B input from the HSI is phase detected on card A4 to produce the HSI lateral deviation signal. The OBS A input from the HSI is also phase detected on card A4 to produce the to-from output to the HSI.

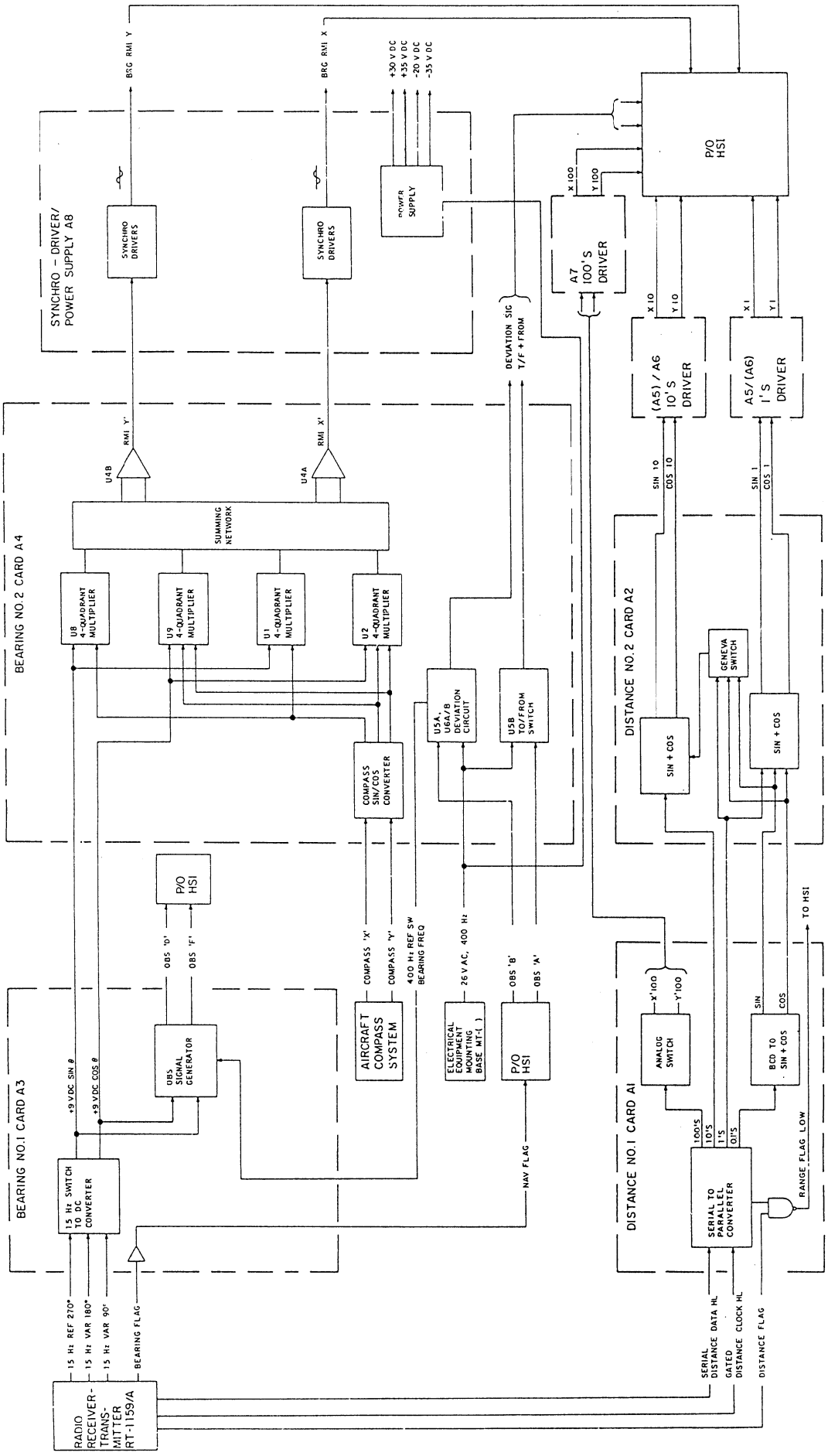
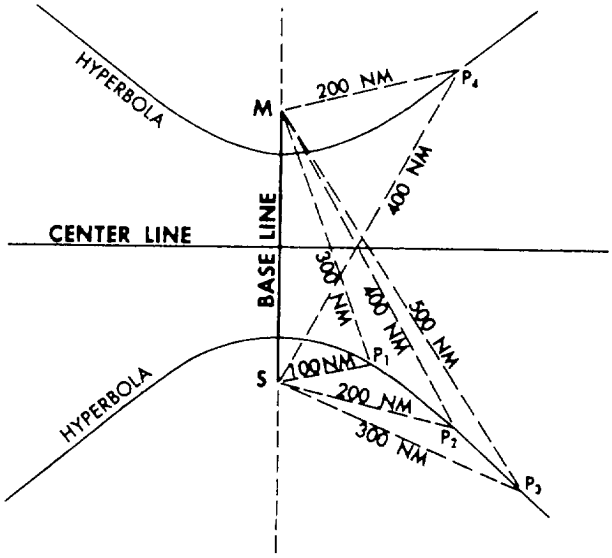


Figure 4-51.–Receiver-transmitter adapter, block diagram.



AT P_1, 300 NM − 100 NM = 200 NM
P_2, 400 NM − 200 NM = 200 NM
P_3, 500 NM − 300 NM = 200 NM
P_4, 400 NM − 200 NM = 200 NM

Figure 4-52.-LORAN hyperbolas.

LONG RANGE NAVIGATION (LORAN) SYSTEM

Learning Objective: *Identify LORAN system to include principle of operation and LORAN C and LORAN D.*

The name LORAN is derived from the words LOng RAnge Navigation, which is an appropriate description of the hyperbolic system of electronic navigation. It provides lines of position over the surface of the earth. Over water, usable LORAN signals can be received at ranges up to 2,800 miles.

The relatively long range of LORAN is made possible by employing low-frequency radio waves. At these frequencies, radio waves are capable of following the curvature of the earth.

LORAN lines of position can be crossed with each other, or with lines of position determined by any other means, to provide fixes. LORAN lines are stationary with respect to the earth's surface. Their determination is not dependent upon compass or chronometer, and it is not necessary to break radio silence to obtain them. It is possible to receive LORAN signals in all weather, except during very severe electrical disturbances.

PRINCIPLES OF OPERATION

Since the speed of radio waves is virtually constant and quite accurately known, the time needed for a signal to travel a given distance can be determined with considerable accuracy. Conversely, the measurement of the time needed for a radio signal to travel between two points provides a measurement of distance between them. All points having the same difference in distance from two stationary points, called foci, lie along an open curve called a hyperbola. Actually, there are two curves or parts to each hyperbola, as shown in figure 4-52, each representing the same time difference, but with the distance interchanged. Thus, the difference in the distances (200 nmi) from the two stations is the same at points P415, P425, P435, and P445.

The LORAN system consists of a series of synchronized chains (set) of radio transmitting stations, which broadcast pulse signals similar to those used in radar, with a constant time interval between them. These transmitting stations are the foci. The aircraft has a combination radio-receiver and time-difference measuring device. The measurements made by this equipment are used for entering tables or charts to identify the hyperbola on which the receiver is located. (Since the earth is not a perfect sphere, the hyperbolas are slightly irregular. This fact is considered when the hyperbolas are plotted on the charts.) Some receivers go one step further and provide this information to a navigation computer for instantaneous latitude and longitude readouts.

LORAN determines the difference in distance by measuring the time interval in microseconds between the arrival of the first signal and the arrival of the second signal from a pair of synchronized transmitters. One of the two transmitters constituting a pair is designated the master (M), the other the slave or secondary (S). The direct line joining these two is called the baseline. The continuations of this line beyond the transmitters are called the baseline extensions. The perpendicular bisector of the baseline is called the center line.

In some areas, a secondary transmitter maybe used as a secondary for more than one master transmitter, It is then known as a double secondary transmitter. Some master transmitters also may be used as secondary for other master transmitters. These are known as master secondary transmitters.

LORAN-C uses an arrangement with one master and two to four secondary transmitters per station. Figure 4-53 shows transmitter configurations of LORAN-C stations in the Pacific area.

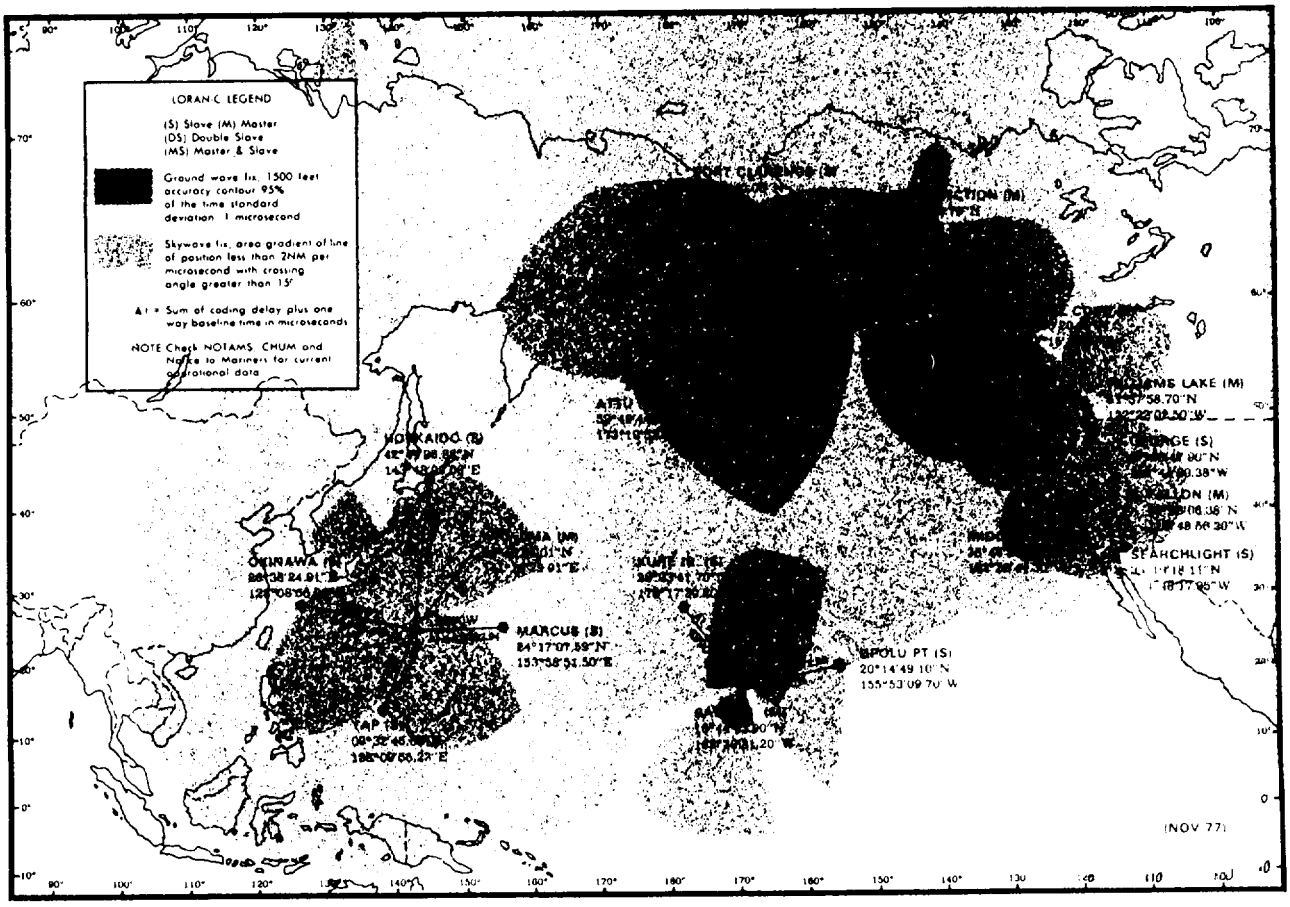


Figure 4-53.-Pacific area LORAN-C coverage.

Baseline Delay

If both signals were transmitted at the same instant, they would arrive together at any point along the center line. At any point nearer the master station, the master signal would arrive first, and at any point nearer the secondary station, the secondary signal would arrive first. Since both signals are alike, this arrangement would be unsatisfactory, as it would include an ambiguity that could be resolved only by knowing the approximate position of the receiver. Near the center line, reasonable doubt might exist as to which line to use. This ambiguity is eliminated by delaying transmission of the secondary signal until the master signal arrives at the secondary station. Radio waves travel at about 299,708 kilometers per second. Since this is equal to 161,829 nautical miles per second, the distance traveled in 1 μsec is 0.162 nautical miles, or 6.18 μsec are needed for a pulse to travel 1 nautical mile. Hence, the length of this delay is the time needed for the signal to travel the length of the baseline or, in microseconds, 6.18 times the length of the baseline in nautical miles. The length of the baseline, and therefore the length of the delay, varies from one station pair to another. This delay is called the baseline delay.

Coded Delay

With the baseline delay in use, the master station transmits a signal first. This signal travels outward in all directions. When this expanding wave front arrives at the secondary station, the secondary signal is transmitted. If no other delays are introduced, the signals travel together along the secondary baseline extension, and the time difference is zero. By the time the secondary signal arrives at the master station, the master signal is a distance away equal to twice the duration of the baseline delay. With this arrangement, however, the time difference readings are so small in some portions of the pattern that identity of each signal is not apparent until the measurement is completed, or nearly so. To avoid this, a second delay is introduced. This is called the coded (or coding) delay. The effect of this delay is to increase all time difference readings by the amount of the coded delay, thus assuring a positive value throughout the pattern. The coded delay can also be used

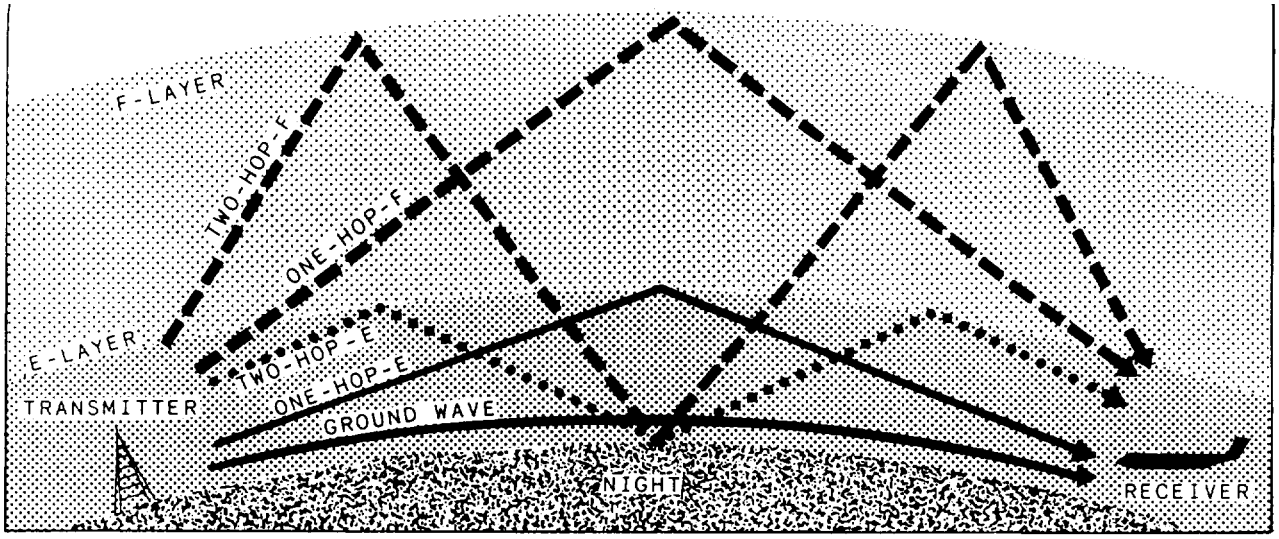


Figure 4-54.-Both ground and sky waves maybe received.

as a security measure against compromise of the system during time of war.

LORAN Reception

The LORAN receiver is similar to an ordinary radio receiver except that it has no speaker. The output of the receiver is fed to a LORAN base indicator, which is an electronic device capable of measuring with high precision the time difference between the receptions of the master and secondary signals. This indicator will measure the time difference by one of the following methods: The first method involves the use of a cathode-ray tube to provide a visual display of the incoming signals. By visually aligning these signals, a reading of the time-difference measurement can be obtained. The second method is done automatically by the LORAN set, and it provides readings of the time difference. The third type of system goes one step further and integrates with a computer to display latitude and longitude.

The readings obtained can be plotted on a LORAN plotting chart, or, in the case of direct latitude/longitude readouts, it can be plotted on any chart.

Factors Affecting LORAN Signals

LORAN signals are affected by such factors as range, the time of day at each transmitter, intervening land masses, and ground waves and sky waves.

DISTANCE FROM EACH TRANSMITTER.– The distance of the master and secondary transmitters from the aircraft is one factor that affects LORAN signals. It is possible to receive a ground wave from one transmitter and a sky wave from the other. For example, when the ground wave from the secondary transmitter is beyond the range of the aircraft receiver, the first pulse in the secondary pulse train will be a sky wave, not a ground wave. A pulse train is the order in which the pulses appear on the trace.

TIME OF DAY AT EACH TRANSMITTER.– Sky waves are normally received at night, but they are also received occasionally during daylight hours. It is not unusual for the first reflection of sky waves to occur in the late afternoon before sunset, and to continue into morning daylight for 3 or 4 hours. This is especially true when the transmitter is in an area that is still dark.

INTERVENING LAND MASSES.– When a ground wave passes over land, its range is significantly reduced because of the attenuation properties of land, As little as 30 miles of land between the transmitter and the receiver can decrease ground wave range by as much as 150 miles. Ground waves that are normally received may not appear because of intervening land.

These factors–range, time of day, and intervening land–should be considered when the pulses on a LORAN indicator are interpreted.

GROUND WAVES AND SKY WAVES.– The path over which LORAN signals travel affects their range, their characteristics, and the reliability of their time

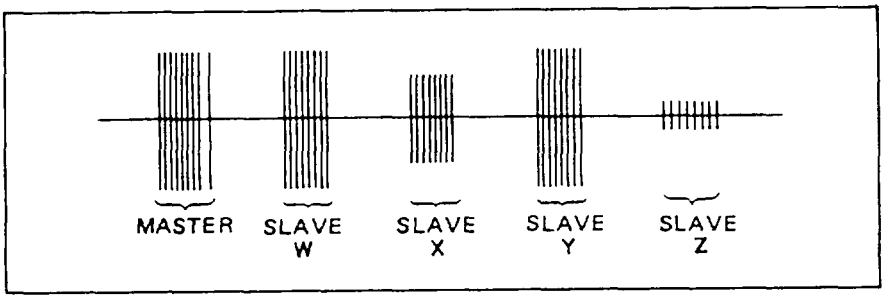


Figure 4-55.-Typical indicator display–time base 1.

difference readings. Radio energy that travels along the surface of the earth is called the ground wave, and that which is reflected from the ionosphere is called the sky wave. The sky wave is named after the atmospheric layer that reflects it and the number of hops (bounces) it takes. See figure 4-54 for examples of sky waves and ground waves.

A LORAN pulse that travels to the ionosphere and back travels a greater distance than one that follows the surface of the earth. The additional distance it travels depends on the height of the reflecting layer, the number of hops it takes, and the distance of the receiver from the transmitter. Because of these variables, sky wave time difference readings are not as accurate as ground wave readings. For this reason, ground waves should be used when available even though they may be considerably weaker than sky waves. When a LORAN receiver is within 250 miles of a LORAN transmitter, sky waves produce an unacceptable error in time.

When ground waves are not available, one-hop-E sky waves can be used to obtain lines of position (LOP) with reasonable accuracy. But, when two-hop-E sky waves are used, the error is multiplied to the point where the time difference readings produce unusable LOPs. The F-layer of the ionosphere is too unstable to provide reliable time difference readings, and is not applicable to LORAN-C.

LORAN-C

Because LORAN-C operates in the low-frequency band between 90 and 110 kHz, it is less subject to attenuation, giving it greater range. This allows the baselines between transmitters to be longer, thereby reducing the number of stations required to provide complete coverage.

LORAN-C uses pulse groups instead of a single pulse for measurement of time differences. The use of pulse groups not only increases average transmitting power, but permits the measurement of time differences with accuracies not attainable before. Phased coding of the multipulsed groups permits station identification and discrimination between ground waves and sky waves.

The master transmitter of a particular LORAN-C network transmits nine pulses in its group; the secondaries transmit eight pulses to a group. The additional pulse in the master group provides visual identification of the station. See figure 4-55.

Transmission Irregularities

The accuracy of LORAN-C transmissions depends upon the correct tuning or synchronizing of the signals. LORAN-C transmitting stations use a “blink” code as a warning of transmission irregularities. Some irregularities are as follows:

- Station not transmitting
- Incorrect phase coding
- Incorrect number of pulses
- Incorrect pulse spacing
- Incorrect pulse shape
- Observed time difference outside specified limits at monitor station

Both master and secondary stations of a pair blink if either station is operating incorrectly; readings obtained from that pair must be treated with caution until both stations have stopped blinking. When a secondary station blinks, the first two of the eight pulses are transmitted for only one-fourth second in every 4 seconds. The master station blinks the ninth pulse in a

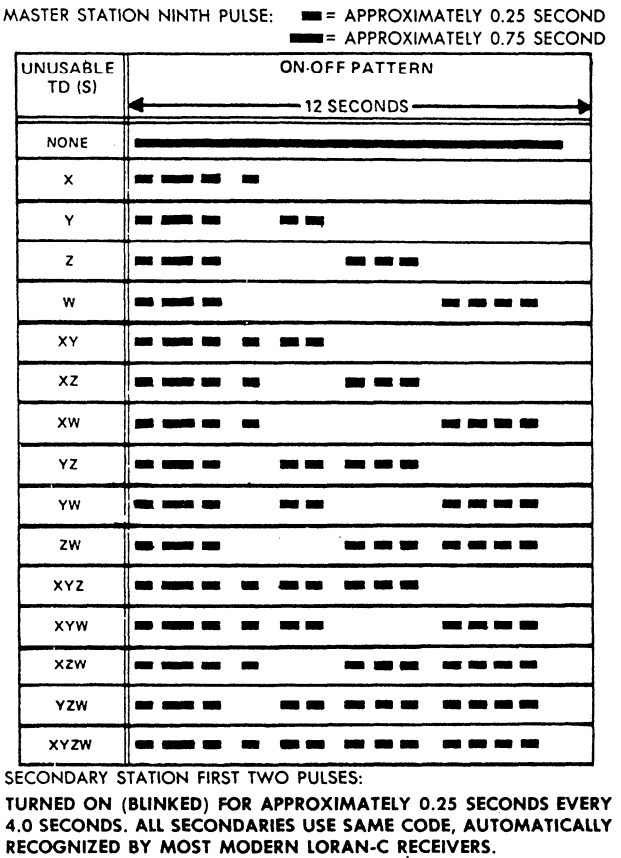


Figure 4-56.—LORAN-C blink code.

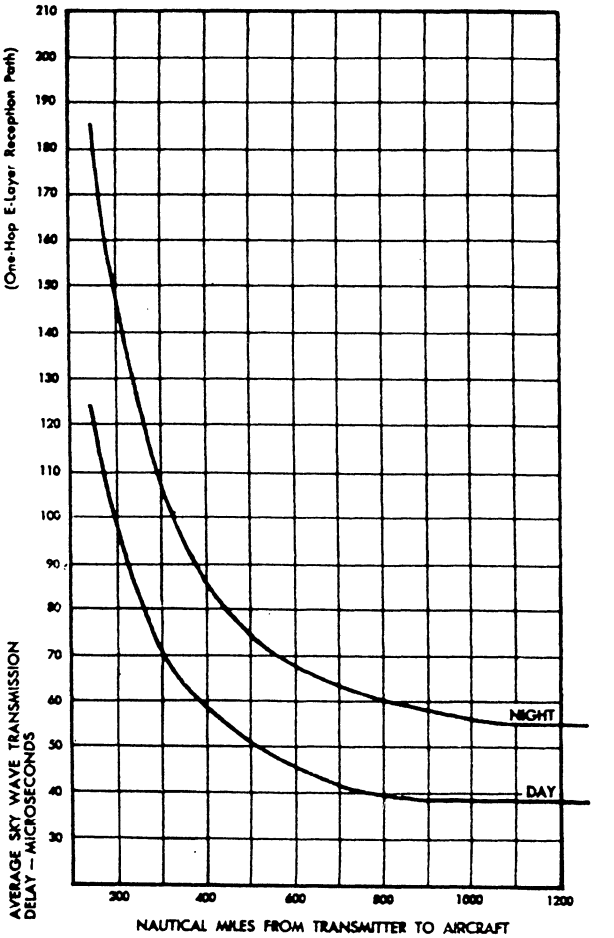


Figure 4-57.—LORAN-C ground wave-sky wave correction graph.

code, which is repeated in a 12-second cycle. See figure 4-56.

The LORAN-C receiver performs four basic functions. It measures and tracks the carrier phase difference of two station pairs. It measures and tracks the envelope time difference of the same two station pairs. It continuously adjusts the amplitudes of the three pulse groups in order to present signals of constant amplitude to the error detectors. It continuously monitors **any** combination of the three pulse groups for evidence of sky wave tracking.

Identifying Sky Waves

Distance from the station and your position are the major determinants pertaining to whether you are receiving ground or sky waves. Sky waves are present both day and night. As a general rule, both are present within 1,000 nmi, with the ground wave strong enough to use for LORAN-C. The area beyond 1,400 nmi will usually be only sky waves. It is the area from 1,000 nmi to 1,400 nmi that poses the largest identification problem. In this area, the ground wave mayor may not be available, which is why identification of the type of wave is so important.

At this point, it is important to look up the possible combinations of ground and sky waves and determine the expected values for each combination. This will aid in determining which combination is present.

LORAN microsecond time delays printed on the charts are for ground waves only. Therefore, corrections for sky wave and ground wave reception must be applied, when applicable, to obtained readings to plot on the chart. This correction is necessary because a sky wave takes longer to get from transmitter to receiver because it must travel a longer distance. The possible combinations are GS (ground from the master, sky from secondary), SG (sky from master, ground from secondary), and SS (sky from both master and secondary). See figure 4-54.

To understand the GS, SG, and SS correction, let's consider a case with GS. The ground wave travels straight from the master to the receiver. The sky wave

from the secondary trounces off the ionosphere and then to the receiver, causing a longer time difference between the reception of the master and the secondary. Therefore, you must subtract a correction for this situation from the obtained readings. This reading can be obtained by using the graph shown in figure 4-57. In the case of SG, the same thing occurs, except that the master signal is delayed. The reading obtained is less than what it should be. We add the correction value obtained from the chart shown in figure 4-57. SS corrections are much smaller than GS and SG for apparent reasons. SS corrections can be found on individual LORAN charts for each chain depicted. A good rule is G master S secondary subtract, SG add, and SS see chart. The only other consideration for the graph in figure 4-57 is whether to use the day or night correction figure. To determine this, time at both the transmitter and the receiver should be considered.

Static

Static caused by a buildup of static electricity due to motion through a moist atmosphere can greatly diminish the maximum ranges of signals. This static usually causes interference with the received signal, and when strong enough, it can completely mask the LORAN signal. This masking looks very similar to "grass" around the signal and can effectively hide the LORAN signal visually and electronically. It is commonly known as "precipitation static noise."

LORAN-D

LORAN-D is very similar in characteristics to its predecessor, LORAN-C. The LORAN-D system has a relatively short range capability and is designed for tactical uses such as close air support and interdiction, reconnaissance, air drop, and rescue. LORAN-D transmitters may be transported to forward operating locations, and it can be operational in short periods of time.

LORAN-D operates in the 90-110 kHz band and has transmission characteristics very similar to LORAN-C. The major difference between the two systems is that LORAN-D transmits 16 pulses per group as opposed to 8 pulses per group transmitted by LORAN-C. LORAN-D is designed to provide precise navigation fixes (average predictable error of 600 feet) out to 250 nautical miles from the master station and usable fixes out to 500 nautical miles from the master station.

Combining LORAN-C/D makes possible more precise navigation and position fixing for a variety of missions. LORAN-C/D receivers provide continuous time difference measurements, which, through the use of modern airborne computers, can provide readouts of aircraft position in latitude and longitude. See figure 4-58.

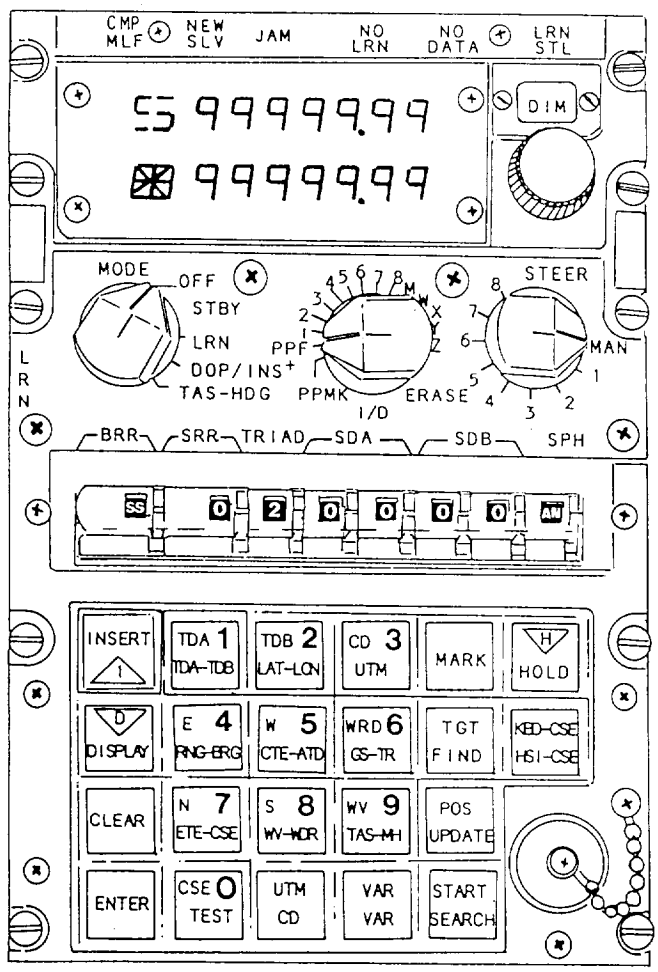


Figure 4-58.-LORAN-C/D receiver.

INERTIAL NAVIGATION SYSTEM (INS)

Learning Objective: *Identify INS principles of operation to include accelerometers, integrators, stable platforms, and computers.*

Inertial navigation is now accepted as one of the best navigation systems for two reasons:

1. An inertial system neither transmits nor receives any signal, so it is unaffected by enemy countermeasures.

2. Theoretically, there is no accuracy limitation in an inertial system. Technology and manufacturing precision can be considered as the factors affecting accuracy.

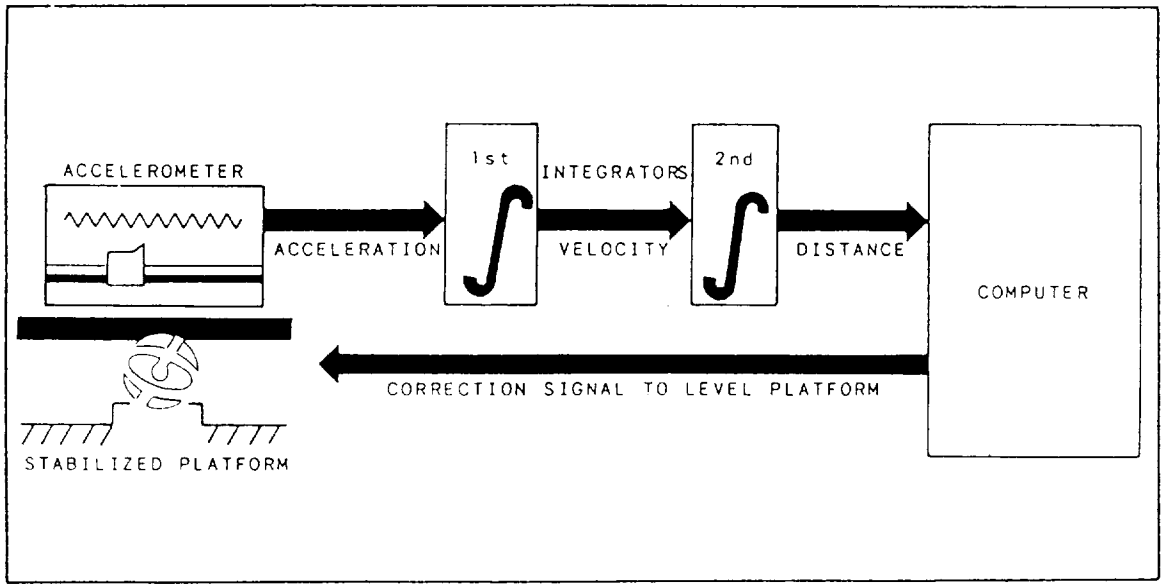


Figure 4-59.-A basic inertial system.

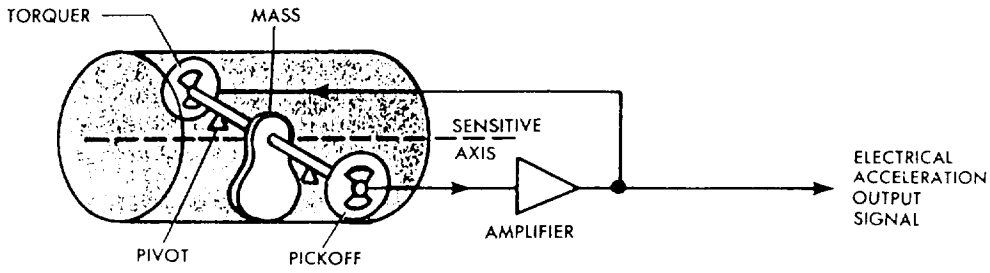


Figure 4-60.-Accelerometer.

An inertial navigator can measure ground speed in the presence of wind and is completely independent of operating environments. The need for a system with these properties has spurred development to the point where the inertial navigator is as good as, or better than, other automatic navigation systems. The inertial navigator provides accurate velocity information instantaneously for all maneuvers, as well as accurate attitude and heading reference.

Inertial navigation system technology has advanced very rapidly within the past few years. Inertia is rapidly becoming the basic element around which advanced navigation systems are designed. Inertial navigation systems with excellent reliability and present position errors of less than 3 nmi per hour are currently employed in a number of operational aircraft, and accuracies of 1 nmi per hour and less are within the state of the art.

PRINCIPLES OF OPERATION

The basic principle of inertial navigation is the measurement of acceleration or displacement rather than the measurement of airspeed and wind velocity, as is necessary in the use of dead reckoning. This measuring of displacement is done with accelerometers. The four basic components in any inertial navigation system are as follows:

1. The accelerometers arranged on the platform to supply specific components of acceleration

2. The integrators to receive the output from the accelerometers and to furnish velocity and distance

3. A stable platform oriented to maintain the accelerometers horizontal to the earth and to provide azimuth orientation

4. A computer to receive the signals from the integrators and to change the distance traveled into position in the selected coordinates

Figure 4-59 shows that the accelerometers are maintained horizontal to the earth by means of a gyrostabilized platform. A signal is transmitted from the accelerometer to the integrators, which perform a

double integration. Distance is fed into the computer, where two operations are performed; first, a position is determined in relation to the reference system used, and second, a signal is sent back to the platform to reposition the accelerometer.

ACCELEROMETER

Acceleration-measuring devices are the heart of all inertial systems. It is most important that all possible sources of error be eliminated and that the accelerometers have a wide range of measurements. Very slight accelerations or even decelerating quantities need to be recorded. Changes in temperature and pressure must not affect the output of acceleration. An accelerometer consists of a pendulous mass, which is free to rotate about a pivot axis in the instrument. There is an electrical pickoff that converts the rotation of the pendulous mass about its pivot axis into an output signal. This output signal is used to torque the pendulum to hold it into position, and, since the signal is proportional to the measured acceleration, it is sent to the navigation computer as an acceleration output signal. See figure 4-60. However, the accelerometers cannot distinguish between actual acceleration and the force of gravity. Acceleration, to be meaningful, must be computed relative to the earth. This means that the accelerometers must be kept level in relation to the earth's surface (perpendicular to the local vertical) if acceleration in the horizontal plane is to be measured. The gyroscopes keep the accelerometers level and oriented in a north-south and east-west direction.

In an aircraft, acceleration must be measured in all directions. To do this, three accelerometers are mounted mutually perpendicular (orthogonal) in a fixed orientation. To convert acceleration into useful information, the acceleration signals must be processed to produce velocity, and then the velocity information must be processed to get the distance traveled. It is true that if acceleration is integrated with respect to time, velocity results. It is also true that if velocity is integrated with time, the result is distance. Any inertial system is based on the integration of acceleration to obtain velocity and distance. Acceleration is a vector quantity, and has not only magnitude but also direction.

INTEGRATOR

The integration of both acceleration and velocity is very critical, and the highest accuracy is essential. There are two types of integrators: the analog and the digital. One of the most used analog integrators is the RC

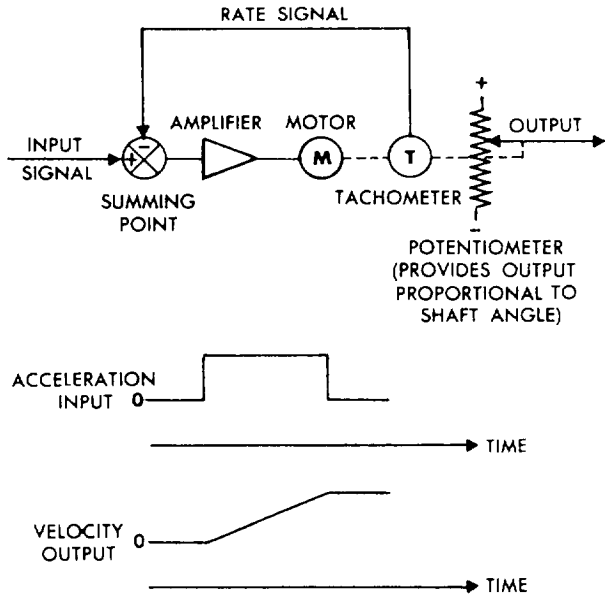


Figure 4-61.-Integrator.

amplifier, which uses a charging current stabilized to a specific value proportional to an input voltage. Another analog integrator is the ac tachometer-generator, which uses an input to turn a motor; the motor physically turns the tachometer-generator, producing an output voltage. The rotation of the motor is proportional to an integral of acceleration. Simply stated, the processing of acceleration is done with an integrator. All an integrator does is produce an output, which is the mathematical integral of the input, or, in other words, the input signal multiplied by the time it was present. See figure 4-61.

STABLE PLATFORM

Gyros are mounted on a platform with the accelerometers and control the orientation of the platform. All inertial systems use a gyrostablized platform to maintain accelerometer orientation. Each platform must contain a minimum of two gyros. If rate gyros are used, three gyros are needed. Each gyro must have its own independent operating loop. The effectiveness of the platform is determined by all parts of the platform, not just the gyros, to include torque motors, servomotors, pickoffs, amplifiers, and wiring. The gyro presents a major problem, particularly concerning precession. Many later developments have appeared, including the air-bearing gyro, which has only 1/10,000,000 the friction of a standard gyro. Other types of gyros have real precession rates of less than 360 degrees in 40 years. The air-bearing gyro has little or no precession. Platforms have been used for years in bombing and fire control systems; autopilots use a basic

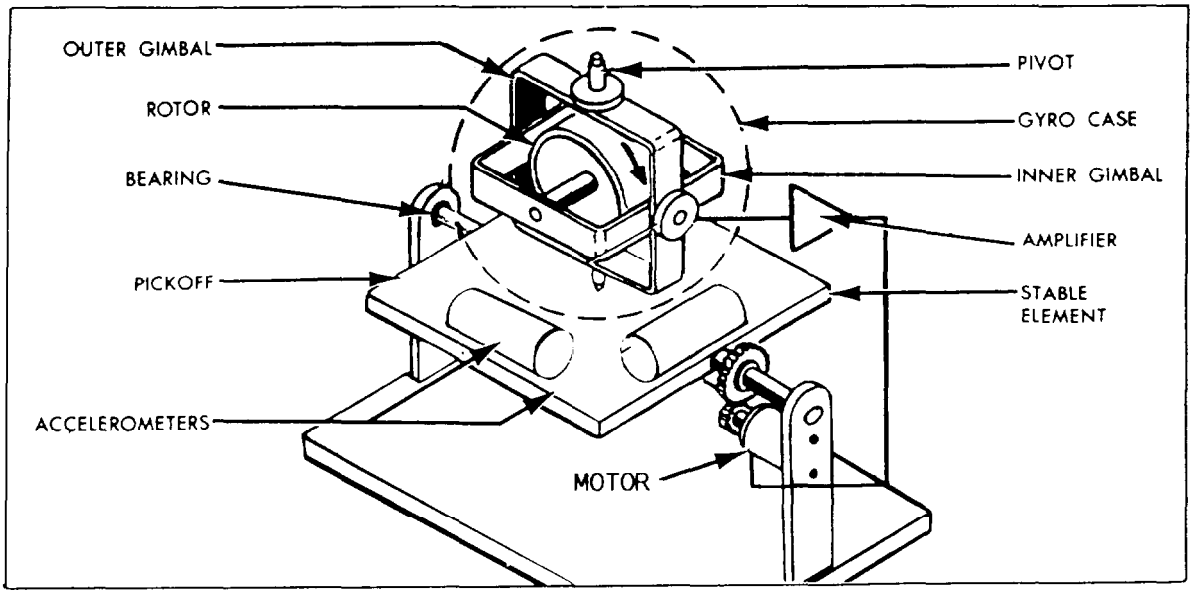


Figure 4-62.-Stable platform

platform. Inertial navigation simply requires a stable platform with higher specifications of accuracy.

A gyrostabilized platform on which accelerometers are mounted is called a stable element. It is isolated from the aircraft's angular motions by gimbals. A simple diagram of a two-degree-freedom gyro mounted on a single-axis platform is shown in figure 4-62.

A gyro tends to remain in its original position when it is up to speed. Any displacement of the stable element from its frame of reference is sensed by the electrical pickoffs in the gyroscopes. These electrical signals are amplified and used to drive the platform gimbals to realign the stable element.

More advanced inertial navigation systems have a four-gimbal platform in a three-axis configuration. The order of gimbal axis is as follows, starting with the innermost axis: azimuth, inner roll, pitch, and outer roll. See figure 4-63.

The four-gimbal mounting provides a full 360-degree freedom of rotation about the stable element, thus allowing it to remain level with respect to local gravity and orientated to true north. This is north as established by the gyros and accelerometers, regardless of the in-flight attitude of the aircraft. The azimuth, pitch, and outer roll gimbals have 360-degree freedom of rotation about their own individual axis. The fourth gimbal, or inner roll gimbal, has stops limiting its rotation about its axis. This gimbal is provided to prevent gimbal lock, which is a condition that causes the stable element to tumble. Gimbal lock can occur during flight maneuvers, such as a loop, when two of the gimbal axes become aligned parallel to each other, causing the stable element to lose one of its degrees of freedom.

Measuring Horizontal Acceleration

The key to a successful inertial system is absolute accuracy in measuring horizontal accelerations. A slight tilt will introduce a component of earth's gravity and incorrect acceleration will be measured. See figure 4-64.

Keeping the accelerometers level is the job of the feedback circuit. The computer calculates distance traveled and, via the feedback link, moves the accelerometer through an equivalent arc. The problem of aligning the accelerometer using this method is complicated by the following factors:

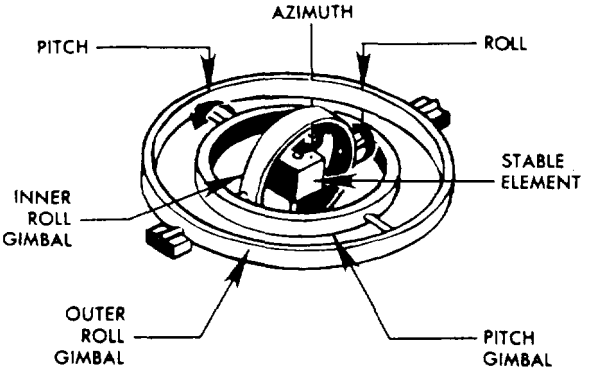


Figure 4-63.-Gimbal platform.

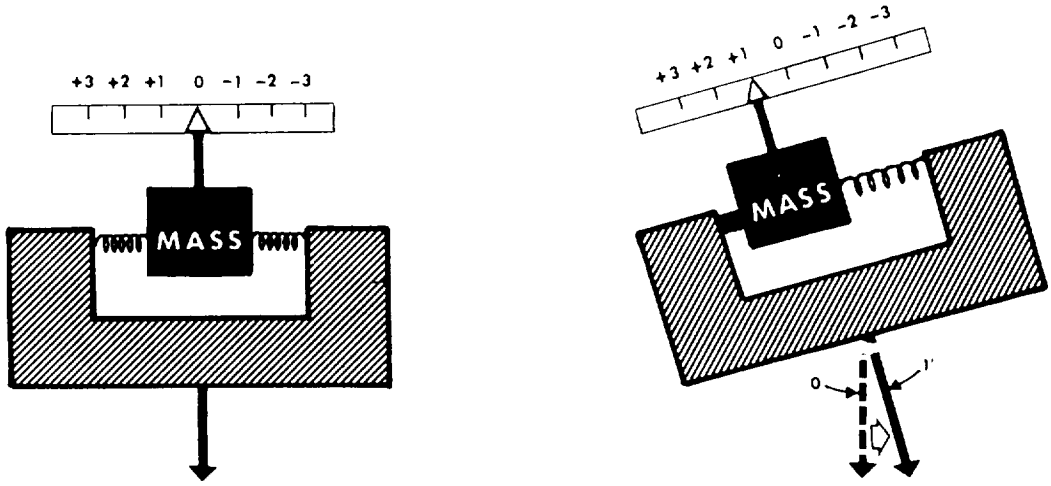


Figure 4-64.-Effect of accelerometer tilt.

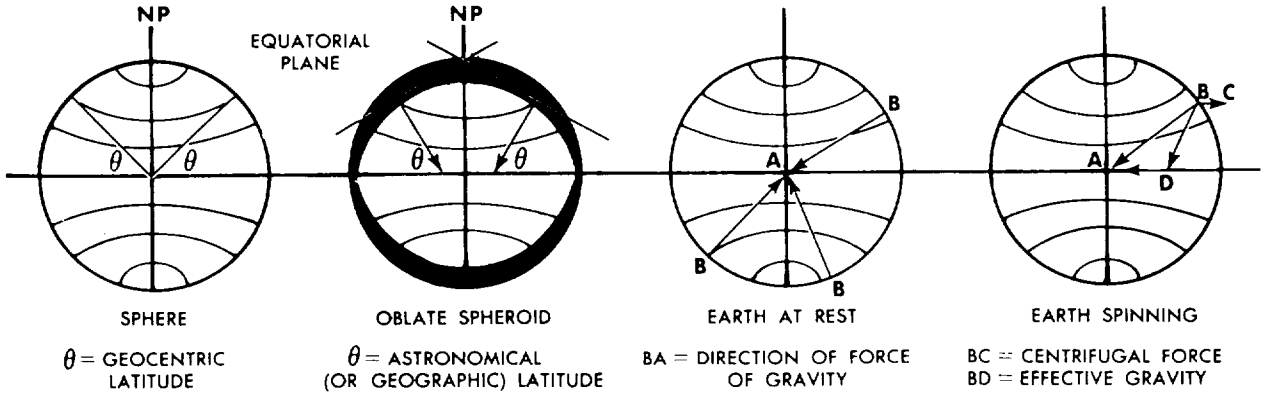


Figure 4-65.-Effect of earth rotation on gravity field.

- The earth is not a sphere, but an oblate spheroid or geoid.
- The rotation of the earth produces a centrifugal force, which deflects the specific force of gravity.
- Because the earth is not a smooth surface there are local deviations in the direction of gravity.

The feedback circuit operates on the premise that the arc transverse is proportional to distance traveled. Actually, the arc varies considerably because of the earth's shape; the variation is greatest at the poles. The computer must solve for this irregularity in converting distance to arc.

The accelerometers are kept level relative to astronomical rather than geocentric latitude. The accelerometers are kept aligned with the local horizon and also with the earth's gravitational field by using the astronomical latitude. The earth's rotation produces a centrifugal deflection that causes gravity to be perpendicular to astronomical latitude. See figure 4-65.

Local abnormalities in the earth's gravitational field are of minor concern. They are compensated for only in vehicles with short inertial guidance terms, such as ballistic missiles.

Accelerometers are kept level by feedback from the computer. Feedback is needed because of two effects, both called "apparent precession." If the inertial unit were stationary at one point on the earth, it would be necessary to rotate the accelerometers to maintain them level, because of the earth's angular rotation of 15 degrees per hour. Also, movement of the stabilized platform would require corrections to keep the accelerometers level. When using a local horizontal system in which the accelerometers are maintained directly on the gyro platform, the gyro platform must be processed by a signal from the computer to keep the

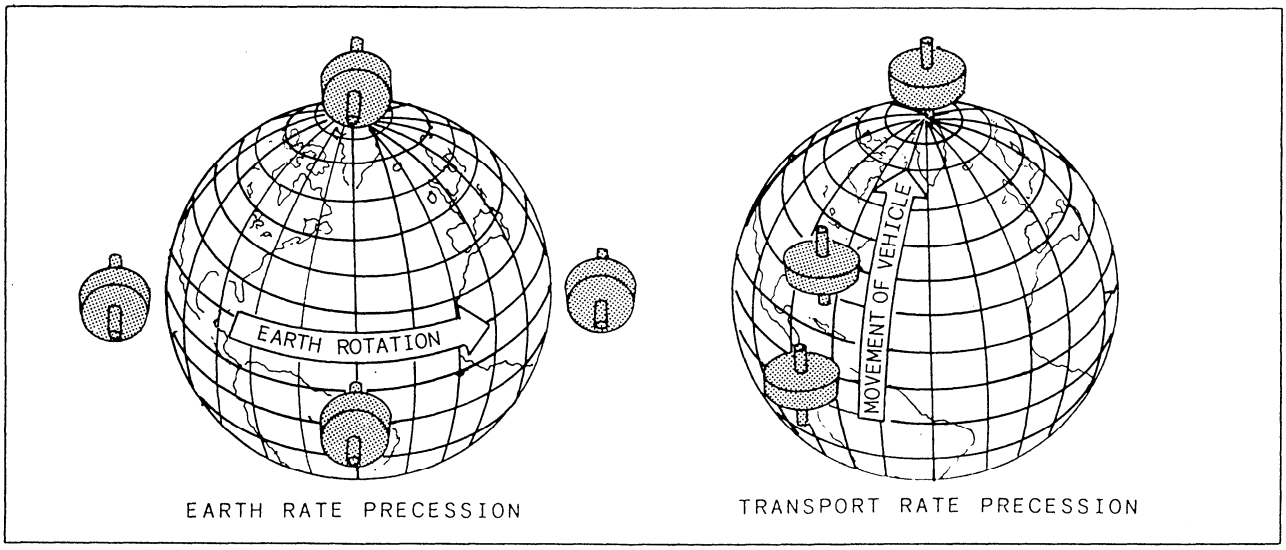


Figure 4-66.—Apparent precession.

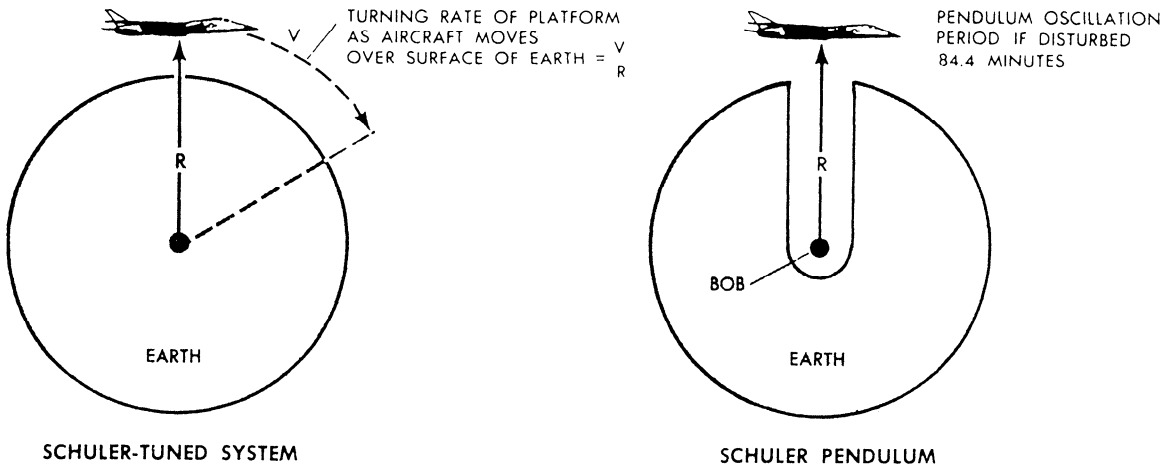


Figure 4-67.—Schuler pendulum phenomennon.

platform horizontal. Apparent precession is illustrated in figure 4-66.

A slight error in maintaining the horizontal will induce a major error in distance computation. If an accelerometer picked up an error signal of 1/100 of the g-force, the error on a 1-hour flight would be 208,000 feet. Dr. Maxmillian Schuler, in 1923, showed that a pendulum with a period of approximately 84 minutes could solve the problem of eliminating inadvertent acceleration errors. The fundamental principle of the 84-minute theorem is that if a pendulum had a radius equal to that of the earth, gravity would have no effect on the bob because the center of the bob would be at the center of gravity of the earth. If a pendulum has a period of 84 minutes, it will indicate the vertical regardless of acceleration of the vehicle. The Schuler pendulum phenomenon prevents the accumulation of errors caused by the measurement of gravity, although it will not compensate for errors in azimuth resulting from the precession of the steering gyro. The amplitude of the Schuler cycle depends upon the overall accuracy of the system. Figure 4-67 shows the Schuler-tuned system.

A gyro that is up to speed and is unslaved or not torqued is space-oriented and will appear to move with respect to the surface of the earth. This is undesirable for aircraft inertial navigators, because the accelerometers will not be kept perpendicular to the local vertical. To earth-orient a gyro, the control of apparent precession is used. If a force is applied to the axis of a spinning gyro wheel that is the to move in a gimbaling structure, the wheel will move in a direction at right angles to the applied force. This is called “torquing” a gyro, and can be considered as induced precession. A continuous torque applied to the appropriate axis by electromagnetic elements called “torquers” reorients the gyro wheel to maintain the

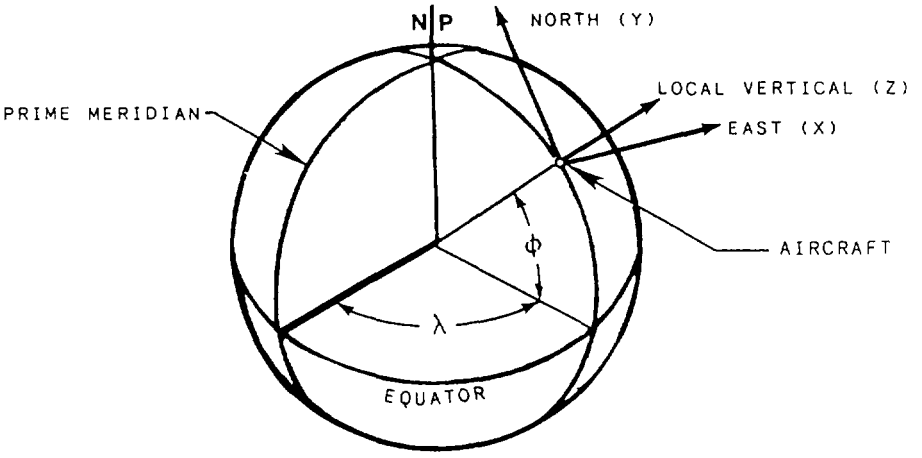


Figure 4-68.-Geographic references.

stable element level, with respect to the earth, and keeps it pointed north. An analog or digital computer determines the torque to be applied to the gyros through a loop that is tuned by using the Schuler pendulum principle. The necessary correction for earth rate depends on the position of the aircraft; the correction to be applied about the vertical axis depends on the velocity of the aircraft.

It is important that the stable element be accurately leveled with respect to the local vertical and aligned in azimuth with respect to true north. Precise leveling of the stable element is accomplished prior to flight by the accelerometers that measure acceleration in the horizontal plane. The stable element is moved until the output of the accelerometers is zero, indicating that they are not measuring any component of gravity and that the platform is level.

Azimuth alignment to true north is accomplished before flight by starting with the magnetic compass output and applying variation to roughly come up with true north reference. From this point, gyrocompassing is performed. This process makes use of the ability of the gyros to sense the rotation of the earth. If the stable element is misaligned in azimuth, the east gyro will see the wrong earth rate and will cause a precession about the east axis. This precession will cause the north accelerometer to tilt. The output of this accelerometer is then used to torque the azimuth and east gyro to ensure a true north alignment and a level condition.

Solving Navigational Problems

The frame of reference of an inertial system will govern, to some degree, the uses of the system. The geographical coordinate system with north reference is the most common, but not the only system used. A north-oriented system requires that one accelerometer be mounted aligned to north and another mounted 90 degrees to the first to sense east-west accelerations. This arrangement allows for any movement to indicate distance traveled east-west and north-south. Distance north-south is converted to coordinates by dividing miles traveled by 60 to obtain degrees; east-west travel requires that distance be multiplied by the secant of latitude and divided by 60 to obtain degrees. This is due to the convergence of meridians and is performed by computers.

Although convenient, latitude and longitude reference has the distinct disadvantage of not being adaptable to use in the polar regions because of convergence of Longitudes. It is possible to offset the pole to a point on the equator. This offset would result in the polar areas being covered by a square grid. There is no specific reason to use a north-oriented system, for no external reference such as magnetic north is used in the inertial system. As a matter of fact, some inertial systems use a principle known as “wander angle,” which does not require the gyros to be oriented to true north. A wander angle inertial system has the advantage of being able to operate in polar regions.

The earth is not a perfect sphere but an ellipsoid, the equator diameter being 27 miles longer than the polar diameter. The inertial navigation system (INS) maintains a continuous local vertical reference and measures distance traveled over a reference spheroid, which is perpendicular to the local vertical. This reference spheroid is mechanized by the INS computer. On this spheroid, the latitude and longitude of the present position are continuously measured by the integration of velocity. In figure 4-68, phi (ϕ) represents latitude (N-S) and lambda (λ) represents longitude (E-W).

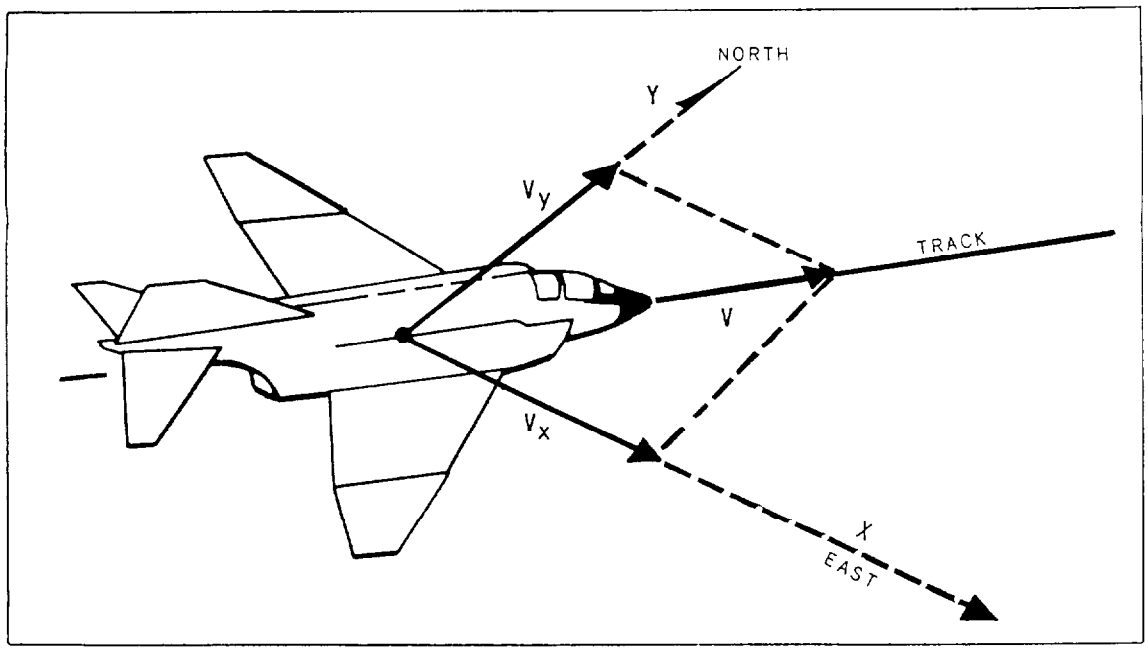


Figure 4-69.-Measurement of aircraft ground speed.

The axes are arbitrarily designated X, Y, and Z–which correspond to east, north, and local vertical, respectively. This defines their positive directions. From now on, reference to velocities, attitude angles, and rotation rates will be about the X, Y, and Z axes. The local vertical (Z) is established by platform leveling. This is the most fundamental reference direction. To complete platform alignment, north (Y) must be known; this is accurately established by gyrocompassing. However, prior to gyrocompassing, the platform is course aligned; which is rotating the platform about the vertical (Z) axis through an angle equal to magnetic heading, plus local variation, to an accuracy of 0.5 degree or less. It should be pointed out here that gyrocompassing established platform alignment to the earth's axis of revolution, or North Pole. The INS is capable of doing this to an accuracy of 10 minutes of arc or less. After the platform is aligned, it remembers its alignment and always stays pointing to true north and the local vertical regardless of the maneuvers of the aircraft.

Ground speed components of velocity in track (V) are measured by the system along the X and Y axes, as shown in figure 4-69. These components, V_X and V_Y, include all effects on the aircraft such as wind, thermals, engine accelerations, and speed brake decelerations. The ground speed (V) is usually displayed by some form of digital readout.

The angles between the aircraft attitude and the platform reference attitude are continuously measured by synchros. The aircraft yaws, rolls, and pitches about the platform in a set of gimbals, with each gimbal being rotated through some component of attitude. True heading is measured as the horizontal angle between the aircraft's longitudinal axis and platform north. This is shown in figure 4-70. Roll and pitch angles are measured by synchro transmitters on the platform roll and pitch gimbals.

COMPUTER

Three of the basic components in any inertial navigation system–accelerometers, integrators, and the stable element with its gyros–have been discussed. The fourth component is the computer.

The principle of inertial navigation does not include fixing en route; thus, there is a need for much greater accuracy in the computers used with inertial navigation than in those used with other systems. The computer function is less complex than that of basic GPI (ground position indicator) units. Since the input from the integrators is already defined as distance, the operation requires only the solution of present position. The second function of the computer is to send a positioning signal to the stabilized platform. Additional operations may be performed by computers (in selected units) such as solution and display of true headings, ground track ground speed, wind direction and velocity. These additional operations are not required of inertial systems computers.

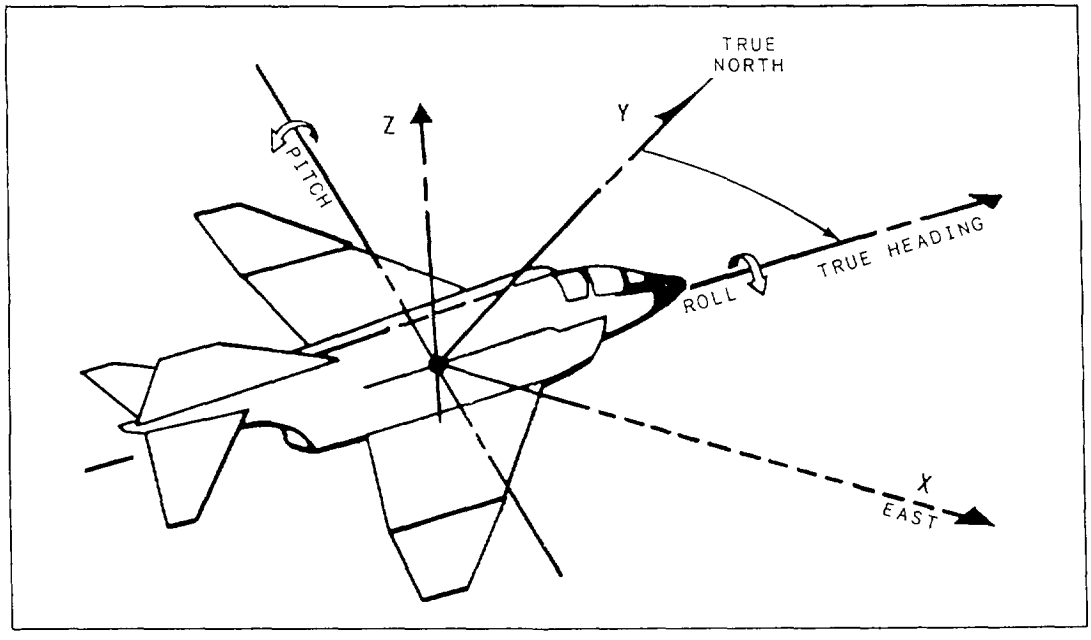


Figure 4-70.-Measurement of aircraft attitude.

DOPPLER NAVIGATION SYSTEM

Learning Objective: *Identify Doppler operating principles and a functional theory of a Doppler radar navigation set.*

In this section, you will learn about a representative radar navigation set. However, before introducing this set, the Doppler principle, as it applies to Doppler radar navigation, is presented.

DOPPLER RADAR PRINCIPLES

Doppler molar uses CW RF transmission along with the Doppler effect. Remember, pulse-type radar determines the distance to the target by measuring the period between transmission of a pulse and receipt of the reflected pulse. The CW Doppler radar, however, senses velocity by measuring a proportional shift in frequency of the reflected signal. This frequency shift is the Doppler effect.

In operation, the airborne CW Doppler radar transmits fixed-frequency RF signals as two or more narrow beams. The beams arc transmitted earthward and displaced laterally and longitudinally at fixed and equal angles. The same airborne set receives a portion of the earth-reflected CW signals, each of which has undergone a Doppler frequency shift caused by the Doppler effect. Velocity proportional difference frequency signals are extracted by continuously mixing the received energy with samples of the transmitted energy, followed by electronic detection. These signals are Doppler signals, and they contain composite velocity information. Since they are of audio frequency, they may be amplified, electromechanically or electronically tracked, and compared with pitch, roll, and altitude rate information to derive the individual components of the aircraft's velocity relative to the axes or relative to the earth's surface.

DOPPLER EFFECT

Electronically, the Doppler effect is the apparent increase or decrease in frequency in a received signal that results from the following conditions:

- The movement of either the transmitter or receiver, or both relative to each other
- The simultaneous movement of a combined transmitter-receiver relative to each other, or relative to a fixed reflecting surface

Disregarding angular motion for the present, the magnitude of the frequency shift is directly proportional to the closing or receding velocity along a straight line distance between the transmitter and receiver, or between the combined transmitter-receiver and fixed reflecting surface.

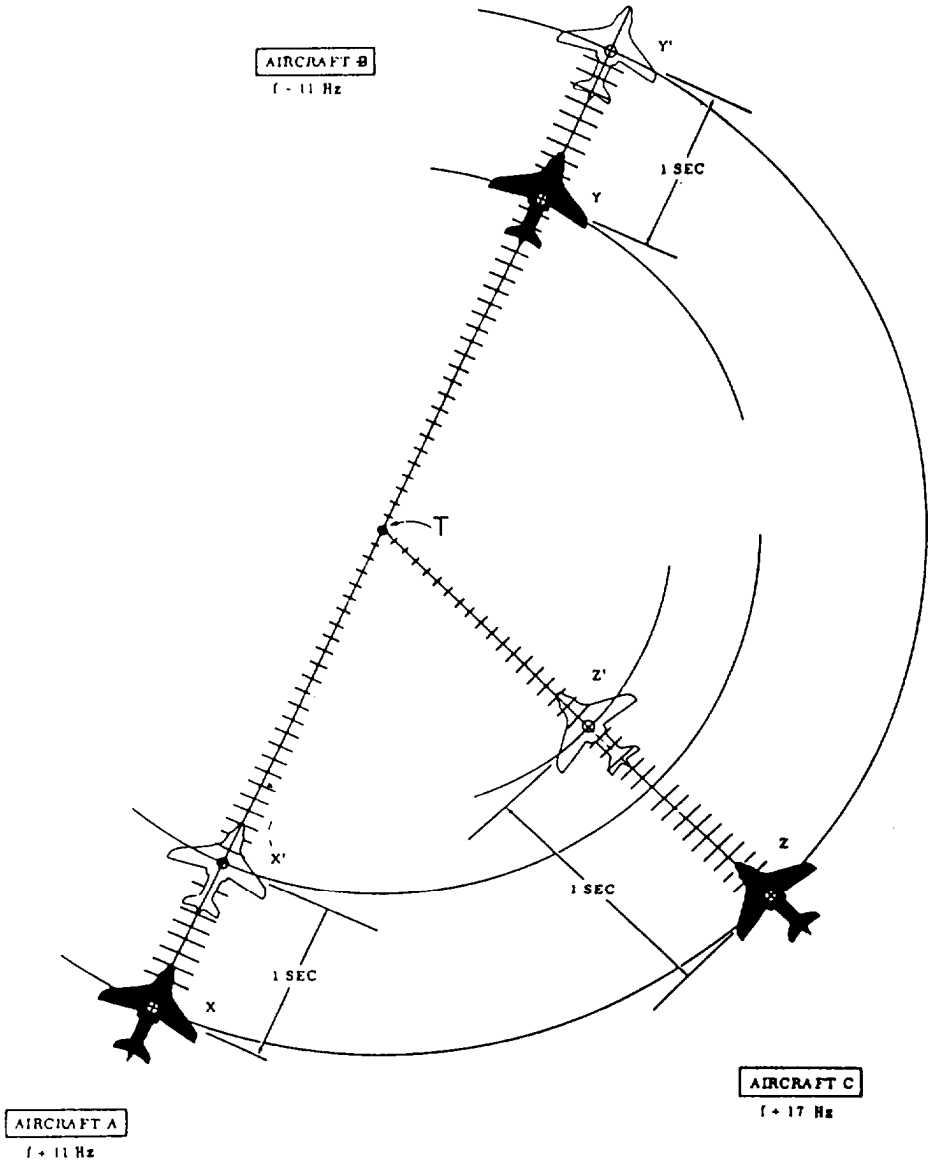


Figure 4-71.-Doppler frequency shift, stationary transmitter, airborne receiver.

Figure 4-71 will help you understand the Doppler effect. Assume three conditions of flight relative to transmitter T, represented by aircraft A, B, and C.

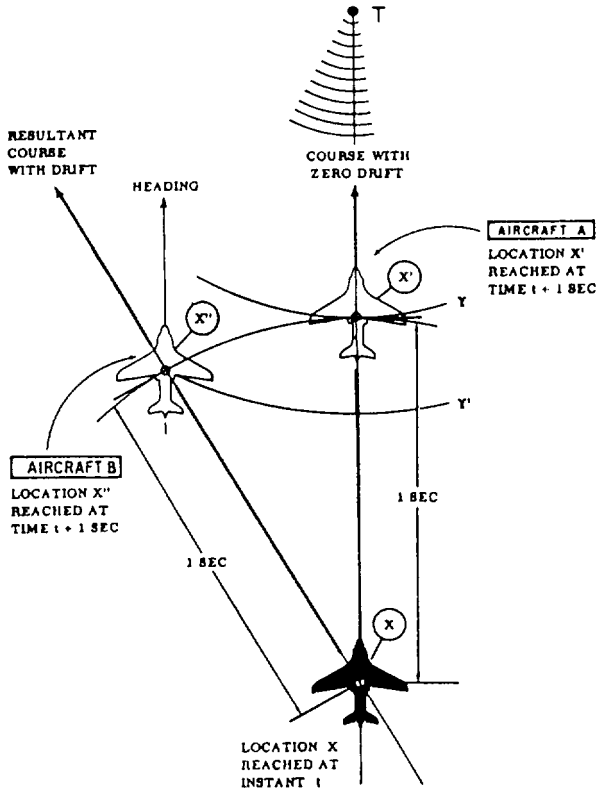
Aircraft A. Aircraft A is heading on a straight line course toward T. It is flying at a constant speed of approximately 11 wavelengths per second and covers the distance from point X to X´ in 1 second.

Aircraft B. Aircraft B is heading on a straight line course away from T. It is flying at the same constant speed as A, and covers the distance from Y to Y´ in 1 second.

Aircraft C. Aircraft C is heading on a straight line course toward T. It is flying at a constant speed approximately 1 1/2 times the speed of A. Aircraft C covers the distance from Z to Z´ in 1 second.

During all three preceding conditions, transmitter T is transmitting a VHF, CW signal of f cycles per second. Therefore, each distance (from X to X´, from Y to Y´, and from Z to Z´) is equal to a specific number of wavelengths at frequency.

Each aircraft (A, B, and C) is equipped with a relatively broadband receiver capable of accepting signals at frequencies higher or lower than transmitted



15Figure 4-72.-Effect of drift on Doppler frequency shift.

frequency f. During the 1-second interval in which A is flying from point X to X′, f number of cycles of the signal from T reach point X. Since A is advancing in a straight line toward the signal source, it receives the following signals:

- The f number of cycles it would receive if stationary for 1 second over point X.
- It simultaneously receives additional wavelengths of the signal en route between point X′ and X.

Therefore, aircraft A receives a signal that is of apparent higher frequency than that emitted by the transmitter.

The magnitude of the preceding frequency shift is proportional to the speed of the aircraft toward T, and inversely proportional to the frequency of the transmitted signal. The direction of frequency shift depends on the direction of flight, relative to the signal source. The effect of flight direction and speed on the direction and magnitude of frequency shift is shown by aircraft A, B, and C. See figure 4-71.

Since aircraft B is flying away from the signal source at a speed equal to the speed of A, the apparent frequency is lower than the transmitted signal. However, it is equal in magnitude to the apparent frequency received by A.

Since aircraft C is flying in the same direction, but at approximately 1 1/2 times the speed of A, the frequency received by C is also apparently higher than the transmitted signal, but greater than that received by aircraft A. The apparent frequency shift is the Doppler effect.

See figure 4-72. An aircraft sets its heading toward a fixed ground transmitter. However, it experiences drift, which if not corrected, results in the position shown by aircraft B. The Doppler frequency shift is proportional to the amount of drift.

Aircraft A. Aircraft A is flying with zero drift and headed straight toward transmitter T.

Aircraft B. Aircraft B is flying at the same speed as Abut drifting past the transmitter.

In the interval of 1 second, aircraft A has approached closer toward the transmitter than aircraft B. You can see this by comparing the arcs scribed at Y and Y′. Because of the faster rate of approach to the transmitter, aircraft A receives more cycles of the transmitted frequency per unit of time than aircraft B. This results in an apparent increase in frequency received by aircraft A over that received by aircraft B. The difference in the Doppler effect on received frequency of zero drift and drift conditions suggests a means of determining drift angle.

TWO-BEAM SYSTEM-AN/APN-122(V)

The Doppler frequency shift previously described also occurs when the transmitter and receiver are in motion together; for example, when they are carried in an aircraft, and received signals are signals reflected from a fixed point on the earth's surface. The receiver senses a Doppler frequency shift that is directly proportional to the speed of the aircraft. The ratio of frequency shift to ground speed is twice that shown in figures 4-71 and 4-72, since a proportional shift occurs in the signal on its way to the earth's surface and occurs again to an equal degree, during straight line flight, in the signal returning to the receiver. To obtain the largest possible frequency shift, you must focus the transmitted beam toward the earth at the angle that yields the largest ratio of change in distance between the transmitter-receiver and the earth's surface. The desired rate of change is obtained when the transmitted beam is

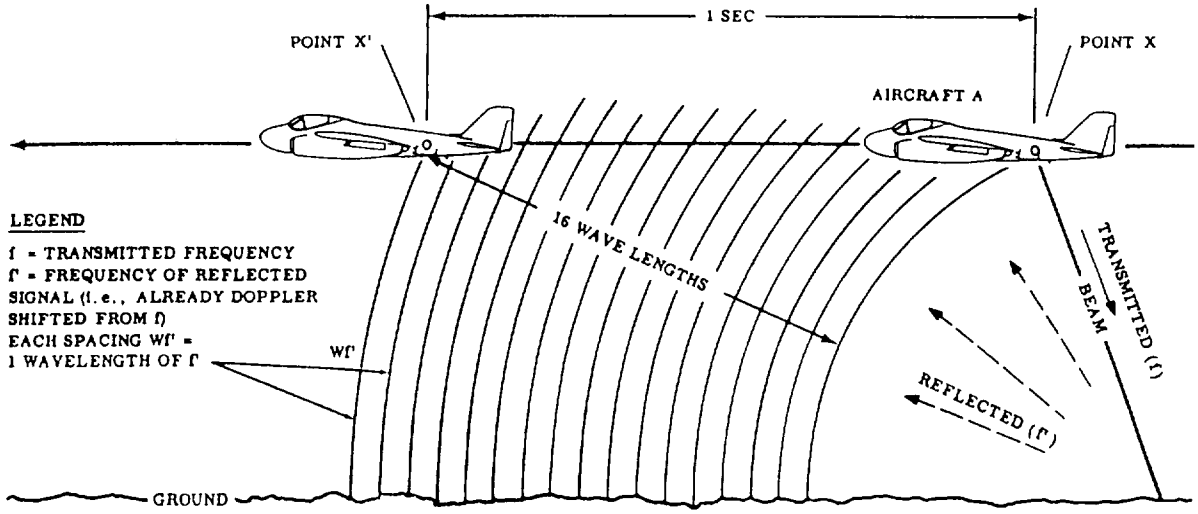


Figure 4-73.-Frequency shift: airborne transmitter-receiver with oblique beam.

inclined toward or away from the direction of the aircraft's heading. Figure 4-73 shows the Doppler shift sensed by an airborne transmitter-receiver in which the beam is focused obliquely downward and aft of the aircraft's longitudinal axis.

In figure 4-73, aircraft A flies from point X to X´ in 1 second. The signal f' Hz is reflected from the ground toward the receiver. Each second the receiver recedes from the reflected signal, a distance equal to 16 wavelengths of frequency f' is traveled. Because each wave must catch up with the receiver, the time between each of the reflected waves is increased, and therefore delayed in arrival. In effect, the reflected signal frequency f' decreases to f' - Hz.

To sense drift and heading velocities simultaneously, two transmitter beams are used. They are directed laterally relative to aircraft heading as well as earthward and aft. In Radar Navigation Set AN/APN-122(V), the two beams are directed symmetrically-one to port and one to starboard of the aircraft's longitudinal axis. The orientation of the transmitter beams relative to the horizontal and vertical planes is shown in figure 4-74.

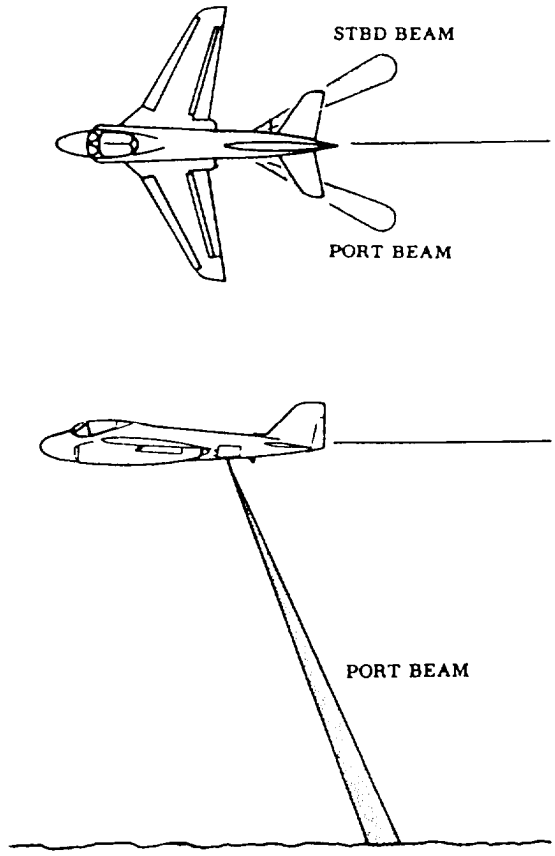


Figure 4-74.-Beam orientation of Radar Navigation Set, AN/APN-122(V).

Effect of Drift On Frequency

Figure 4-75 shows the Doppler effect in an aft-oriented, dual beam system. Assuming the aircraft is not drifting, simultaneous radiations of each beam are reflected back to the aircraft from separate points on the earth's surface. The time in transit of the microwave energy in each beam is a direct function of the aircraft velocity. If they could be seen, the continuous waves of transmitted energy would form dual traces across the earth's surface; these traces comprise successive instantaneous points of reflection.

In figure 4-75, the port transmitted beam P has a rate of travel equal to its corresponding starboard transmitted

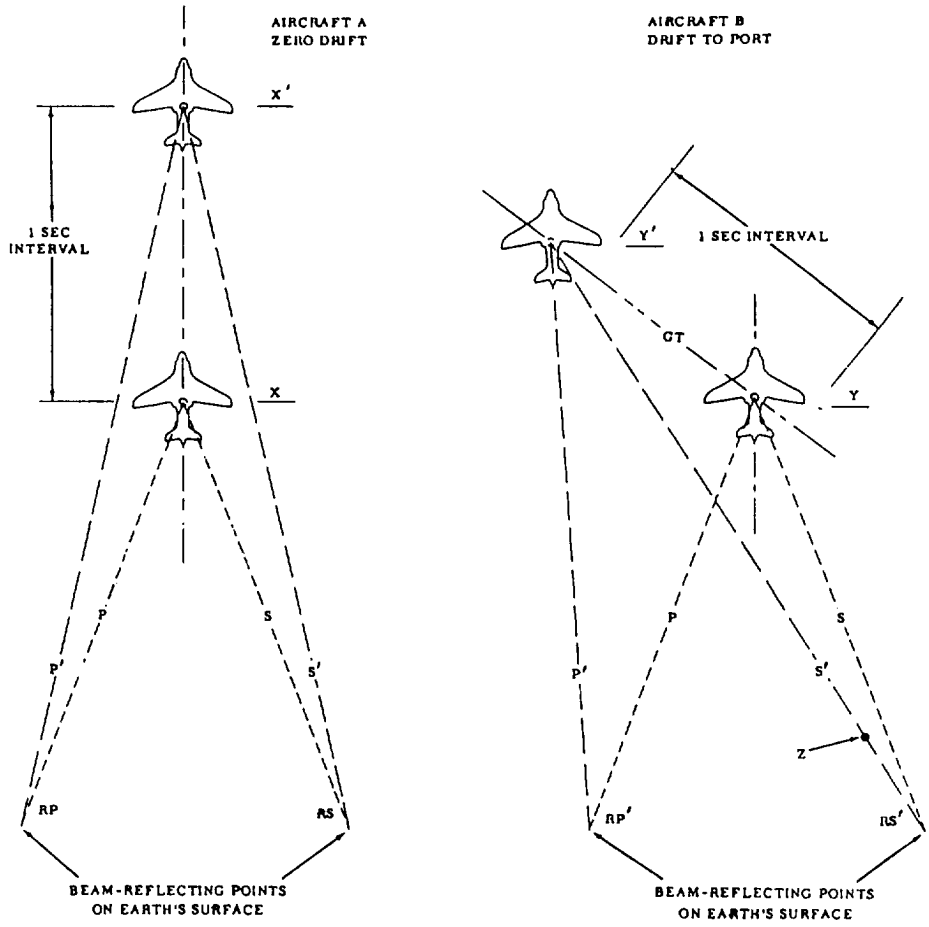


Figure 4-75.-Effect of drift on frequency shift in dual beams.

beam S. Therefore, points RP and RS and RP′ and RS′ are the instantaneous points of reference for determining the rate of travel by the received beams of aircraft A and B, respectively. In the zero drift condition, aircraft A maintains heading H from X to X′ for 1 second. The rate at which A recedes from reflecting points RP and RS is equal; therefore, the effective wavelength of the RF energy in beams P′ and S′ is equal. Since Doppler frequency shift is proportional to the relative velocity between the reflector (earth) and the receiver, the Doppler frequency shift in beams P′ and S′ of A is equal.

In the drift to port condition, aircraft B encounters drift forces and covers ground track GT in 1 second. The rate at which B recedes from reflecting point RS′ is greater than the rate at which it recedes from point RP′. Therefore, the distance that beam S′ of aircraft B travels in 1 second is greater than its associated beam P′ in equal time by the distance from point Z to RS′. Since the rate of travel of effective wavelength for beam S′ is greater than beam P′ in a port drift condition, the Doppler frequency shift is greater in S′ than in P′. If the drift were to starboard instead of port, the largest Doppler frequency shift would occur in the port beam.

You can see that by using dual oblique beams, the direction of drift can be derived from the received beam that registers the least frequency shift. The velocity of drift can be computed from the difference between the magnitude of frequency shift in each beam, and ground speed can be computed from the sum of the magnitudes of frequency shift in each beam. The method used for deriving direction and velocity of drift and ground speed in radar navigation sets is similar to this.

Functional Theory of the Radar Navigation Set

Two narrow beams of continuous microwave energy are transmitted by the transmitter-receiver and focused obliquely toward the earth. The beams are displaced laterally, one to port and one to starboard, and aft (fig. 4-74). A portion of the transmitted energy from each beam is reflected from the earth's surface at an angle sufficient to be intercepted by the receiving portion of the transmitter-receiver antenna. Each

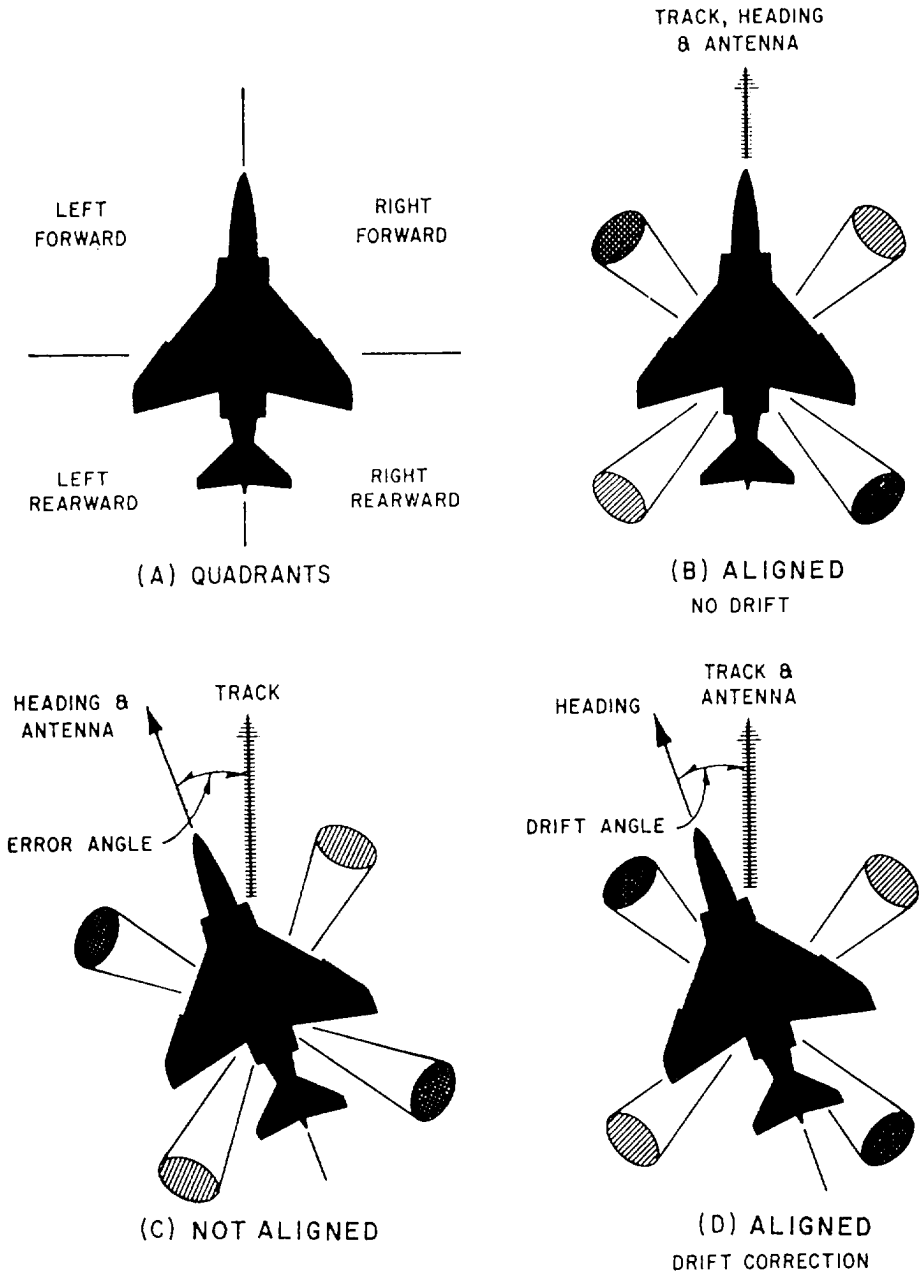


Figure 4-76.-Four-beam pattern, using two-beam transmission.

received CW signal is mixed with continuous samplings of the transmitted signal within the waveguide assembly of the antenna, then coupled to their respective port or starboard crystal detectors. The resultant difference frequency output signal (f_{dp} and f_{ds}) of each detector is proportional to the aircraft's velocity along the port and starboard beam axes, respectively. These Doppler signals are of equal frequency as long as the aircraft maintains a straight line heading with no pitch or roll. If the aircraft drifts to port or starboard, the following occurs:

- Its drift is sensed as a decrease in Doppler frequency derived from the beam that is focused in the direction of drift.
- An increase in Doppler frequency is derived from the beam focused opposite to the direction of drift.

Because of the lobe pattern of the transmitted beams, the reflected energy of each beam comprises a narrow band of frequencies rather than a single frequently. The result is narrow-band Doppler difference signals (port Doppler frequency and starboard Doppler

frequency), which are fed from the port and starboard crystal detectors, respectively, to the dual channel amplifier assembly. At this point, each signal is amplified under automatic gain control and then fed to the signal data converter.

Using what are initially only noise inputs, the signal data converter initiates a search function to acquire well-defined Doppler signals in both port and starboard channels. Simultaneously, this converter supplies gain-increasing AGC feedback voltages to the amplifier assembly and a memory signal used in the ballistics computer. This memory signal also causes an indicator lamp to come on, indicating that the signal data converter is operating in the search mode. When a satisfactory Doppler signal is acquired, the signal data converter locks on and tracks the center of each Doppler energy spectrum, and converts these signal frequencies to rate-proportional pulse outputs. The memory signal is cut off, and the indicator or lamp goes out. These rate-proportional pulse outputs, which are representative of aircraft velocity data, are then routed to the ballistics computer.

FOUR BEAM SYSTEM-AN/APN-153(V)

The AN/APN-153(V) is a lightweight, miniaturized ground speed and drift angle measuring system. It is designed to satisfy the navigational requirements of modem military aircraft. It uses Doppler pulsed radar techniques to measure ground speed and drift angle directly, continuously, and accurately. It is easily installed in most fixed-wing aircraft because of its small size and lightweight.

The AN/APN-153(V) emits short pulses of microwave energy, so the transmitter is not operative while an echo is being received. Therefore, in a single-beam radiation pattern, there is no second frequency with which to compare the echo. However, two beams–one projected dead ahead and the other directly behind-would experience equal but opposite Doppler shifts. In this case, one-half the difference between the two echo frequencies is the Doppler shift of either beam, and this shift is proportional to ground speed. This condition assumes no drift. If drift is present, the two-beam system described above does not detect it when used with pulsed beams. Antennas radiating to the left and right detect sideways motion, but not fore and aft motion.

To make simultaneous measurement of both ground speed and drift angle, a modification of the two-beam pattern is used in the AN/APN-153(V). The vicinity of the aircraft is divided into four quadrants-left forward, left rearward, right rearward, and right forward (fig. 4-76, view A). Simultaneous beams are radiated into diagonal quadrants with switching of the quadrants at regular intervals.

When there is no drift, the ground track and aircraft heading coincide. In this condition (fig. 4-76, view B), the radiation pattern is symmetrical about the axes of the aircraft, and the left and right beams are at the same angle from ground track. The Doppler shift is equal in the beams.

When the antenna is not aligned with ground track (fig. 4-76, view C), an angle exists between the antenna axis and ground track. The beams experience unequal shifts. The difference frequency resulting from one pair of beams is higher than that from the other. As the switching action occurs, this difference is processed as an error signal, which, through servo action, drives the antenna into alignment with the ground track.

When the antenna is aligned with ground track, the angle between the antenna axis and the aircraft heading represents the drift angle. Through a synchro system, this information is supplied to the computer along with the ground speed data.

EIGHT-BEAM SYSTEM-AN/APN-190(V)

The Doppler Radar Navigation Set AN/APN-190(V) is an airborne ground speed and drift measuring system. It uses the Doppler principle to extract data from reflected signals. A lightweight system, operating on a frequency of 13.325 (0.05) GHz, precisely measures, processes, and indicates ground speed and drift angle information aboard aircraft. The AN/APN-190(V) is essentially a solid-state system of low operational reliability that requires maximum maintenance. It is used with a navigation/weapon delivery computer, inertial measurement system, and a heads-up display set.

In measuring the ground speed and drift angle of an aircraft, the system transmits narrow beams of electromagnetic energy downward from an antenna stabilized in pitch, roll, and azimuth to illuminate small areas on the earth's surface. A portion of the transmitted energy is reflected from the surface, received, and processed by the system to determine the Doppler frequency shift.

The AN/APN-190(V) radiates a four-beam pattern that consists of two-beam pairs alternately switched from left to right of the aircraft's fore-aft axis.

Frequency shifts occurring in each beam are proportional to the components of the aircraft's forward and lateral velocity along the respective antenna beam. The frequency shifts are detected by mixing a portion of the transmitted signal with the received signal and detecting the audio beat frequency.

After detection, the Doppler signals are fed to ground speed and azimuth tracking circuits. These circuits convert the signals into voltages proportional to ground speed and azimuth. Outputs proportional to these voltages are provided in proper electrical form to drive the appropriate auxiliary equipment.

The magnetron in the radar transmitter generates 200-volt pulses of RF energy at a PRF of 80 to 120 kHz with an average power of 10 watts. The magnetron output is applied to a crystal switch that passes the magnetron pulses to a duplexer, which routes the microwave energy to a waveguide assembly. During the transmit cycle, the receiver crystal switch is closed, minimizing coupling of transmitted energy into the receiver. The duplexer permits the use of the same antenna for both transmitting and receiving, and it isolates the receiver from the transmitter. The duplexer also couples the transmitted signal to the waveguide assembly, and, during the receive cycle, couples the return signal from the waveguide to the receiver switch. The waveguide couples transmitter RF energy to the antenna. Four radiating elements composed of a J-band waveguide, four reflectors, and a waveguide flange complete the antenna array.

When the set is transmitting, the generated RF energy is applied through the crystal switch to the duplexer and waveguide assembly, coupling the energy to the antenna so that the switching and array combination generates eight beams of energy. See figure 4-77. The beams are radiated two at a time in the following order:

1. Front right (A) and aft left outer (A´)
2. Front right (B) and aft left inner (B´)
3. Front left (C) and aft left outer (C´)
4. Front left (D) and aft right inner (D´)

Left-right switching and fore-aft beam lobing are caused by coded signals supplied by the receiver-transmitter to the switching modules in the antenna assembly.

The antenna is stabilized in pitch and roll to compensate for changes in aircraft attitude. Pitch and roll stabilization is achieved by means of servo loops

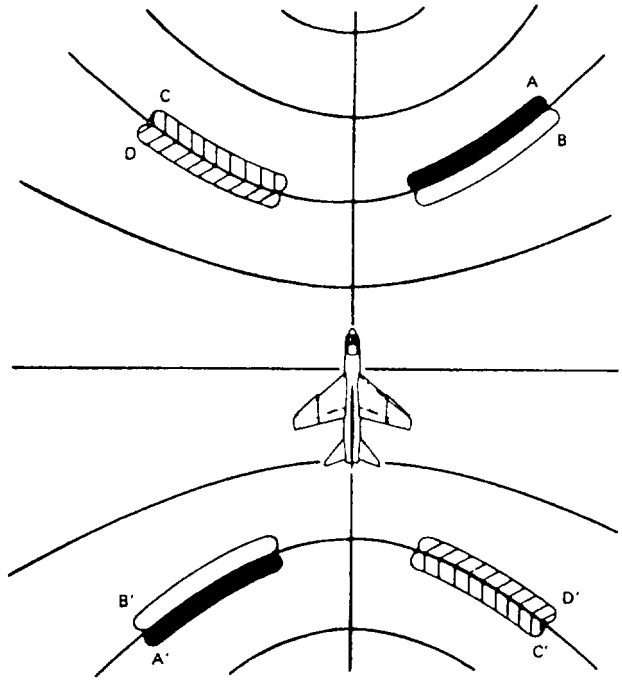


Figure 4-77.-Antenna radiation pattern.

that receive pitch and roll attitude signals from a pitch and roll reference input. Servomotors drive the antenna array pitch and roll gimbals through their respective gear trains. The array is also driven in azimuth by a servo loop, aligning it with the aircraft ground track to compensate for aircraft drift conditions up to 30 degrees. During receive, the return signal is coupled with the antenna through switching modules, through the waveguide assembly to the duplexer, where it is then routed to a receiver switch.

The receiver accepts the Doppler return and processes it to provide output frequencies falling within a 1-to 36-kHz band, which contains both ground speed and azimuth drift information. These output frequencies are applied to frequency tracking circuits, which consist of ground-speed tracking and azimuth tracking loops. The ground tracking circuits measure Doppler frequency and provide an analog ground-speed output. The azimuth tracking loops compare Doppler shift of two transversely switched beam-pair returns and drive the antenna in azimuth until it is aligned with the ground track. At this time, the Doppler shifts from both beams are equal. A potentiometer positioned by an antenna azimuth drive shaft provides an electrical analog output of drift angle to analog-to-frequency conversion circuits. Frequency-to-digital conversion circuits use input frequencies to represent ground speed and drift angle to provide a 60-bit word. This word represents ground speed, drift angle, and system status for use by

digital navigation computers and in inertial system correction.

NAVIGATIONAL COMPUTER SYSTEM

Learning Objective: *Identify the navigational computer system to include the sensors, computer and navigation panels.*

A navigation computer system relieves the navigator of many manual operations required to direct the aircraft in flight. When automatic sensing devices are tied into a navigation computer system, the navigator is automatically provided current readings of present latitude and longitude, ground speed, and heading. The navigation computer system eases the navigator's workload and frees him or her to make the decisions that are beyond the capability of computers.

Modern aircraft are capable of speeds and ranges that require the navigator to perform extensive calculations rapidly and accurately. For example, on a flight from the United States to a foreign country, the route could pass over land, water, and ice caps. The navigator must contend with overcast or undercast conditions, day and night flight, altitude changes, turning points, and mandatory ETA requirements. To handle all these conditions at the speed of sound or faster, the navigator uses automatic navigation computers.

The navigational computer system consists of the following components:

The data-gathering units (sensors) such as radar, Doppler, INS, LORAN, and TACAN

Computer units where the computations and comparisons are made

Navigation panels containing the dials and controls that give the navigator a system monitoring and control capability

SENSORS

Sensors are data-gathering units such as conventional radar, Doppler radar, inertial navigation system (INS), long range navigation (LORAN), and tactical air navigation (TACAN) systems.

Conventional Radar

When a radar set is incorporated into the computer system, movable electronic cross hairs are displayed on the radarscope so that range and direction of radar returns are measured and inserted into the computer. See figure 4-78. The cross hairs consist of a variable range mark and a variable azimuth mark. They are maneuvered with a cross hair control handle. On the radarscope, they resemble a single fixed-range mark and a heading mark. By moving the cross hair control handle, the navigator simultaneously changes the position of the cross hairs and the corresponding coordinate measurements (east-west and north-south) being fed to the navigation computers. The function is completed almost instantaneously.

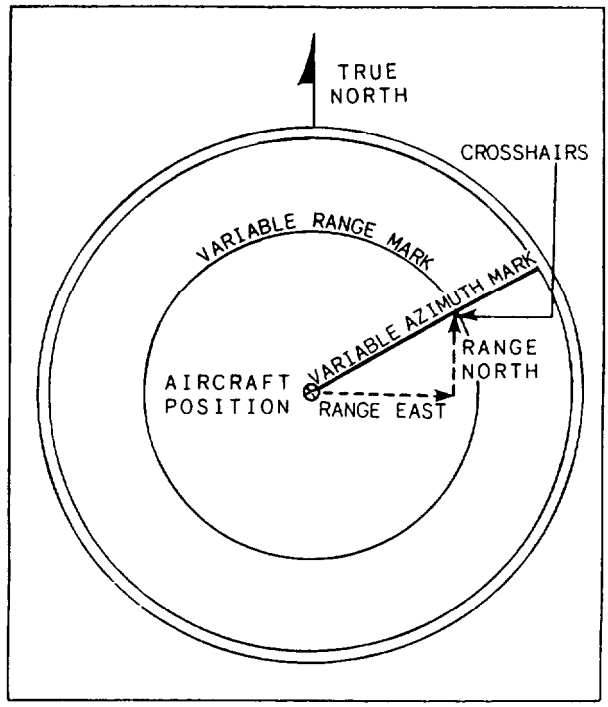


Figure 4-78.-Radar cross hairs.

When the navigator positions the cross hairs on a given return, the computers determine the distance between the aircraft and the return. If the coordinates of the return have been set in the computer, the computer can maintain a running account of the aircraft latitude and longitude.

Doppler Radar

Doppler radar's contribution to the computer system is ground speed and drift angle. These two outputs are put to several uses in the computer system. Doppler ground speeds is used to drive the present position latitude and longitude counters. Doppler outputs are used in platform leveling and in checking inertial ground speed in an inertial system. Doppler

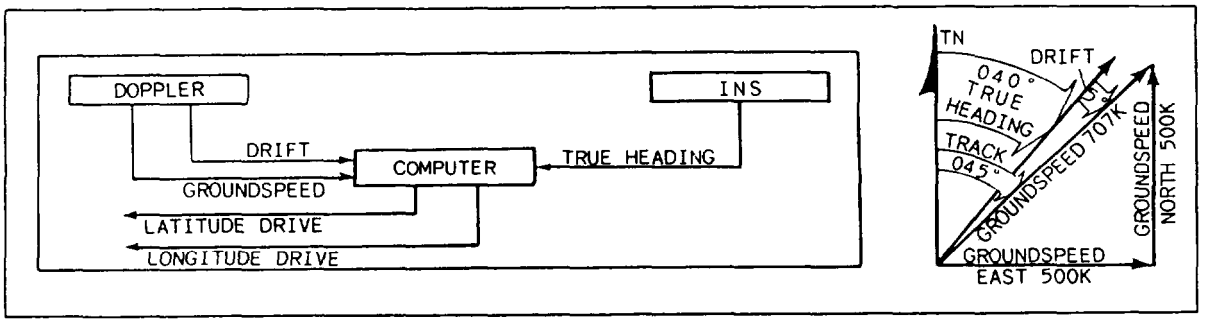


Figure 4-79.-Present position counter drive.

radar is an essential part of many navigation computer systems.

Inertial Navigation System (INS)

The INS is used to feed velocity information into the computers. Once the inertial sensor is leveled and in operation, it is used to continually update the present position counters.

Long Range Navigation (LORAN)

LORAN fits in well with an automatic computer system. Some computer systems have the coordinates of LORAN stations stored in them. During flight the navigator selects the stations he or she wants to use and the computer does the rest. Fixing is automatic and occurs in the same way that the navigator takes a celestial fix. An assumed position is determined by the computers; then, the LORAN position is applied to this assumed position. A series of credibility checks and approximations are applied automatically to the computer. The result is an accurate LORAN fix. When the computer functions in the LORAN mode, continuous present position and ground speed information is still available.

Tactical Air Navigation (TACAN)

TACAN can easily be added to a computer system. Since the TACAN output is given in the form of a range and bearing, the computers only need the coordinates of the TACAN station being used. This data is set into the computer before the mission begins. Some confections must be applied to TACAN outputs to increase accuracy. The bearings received from TACAN are magnetic; therefore, the computer must have an accurate magnetic variation value at all times. This is usually built into the computer. TACAN range output is expressed in slant range. The computer applies absolute altitude above the station to the slant range to produce exact ground range.

COMPUTER UNIT

The two basic types of navigation computers are the analog and the digital computer.

Analog

An analog computer is comparable to the navigator's handheld computer because a graphic replica of the problem to be solved is constructed to find the answer. The analog computer is generally larger than the digital computer because many components must be added to solve a wide variety of problems. Some of these computers weigh as much as 2,000 pounds. The analog computer has one main advantage-it is not as sensitive to temperature and pressure changes as the digital system.

Digital

The digital computer is generally lighter and more compact than the analog system. In some cases, the digital computer weighs less than 100 pounds. It computes navigation problems in the same way as the analog computer. It is unnecessary to design a digital computer expressly for the navigation problems it is to solve. Properly programmed, the same computer could be used in fields other than navigation. This is possible because the digital computer deals strictly with numbers. This requires that all inputs be changed to a numerical value before they are sent to the computer. Likewise, all outputs must be converted back to terms that are meaningful to the navigator.

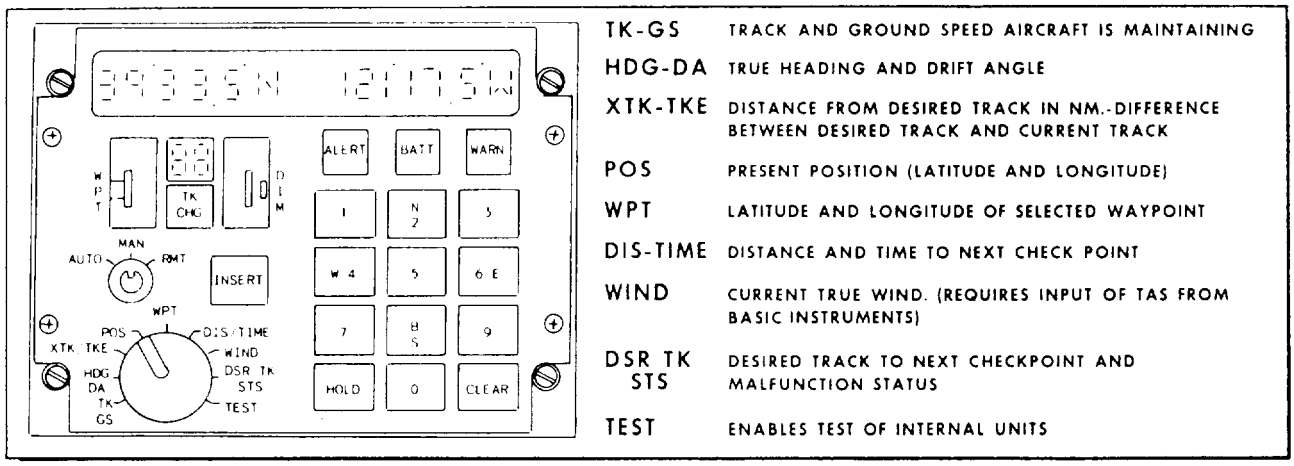


Figure 4-80.-Typical control display unit.

Determining Position

Regardless of computer type, the problems to be solved by a navigation computer remain the same. The ever-present problem facing the navigator is determination of aircraft position. With a computer system, it is not necessary to estimate a position based on a track and ground speed derived from the last known position. The computer always displays the current position for convenient reading.

Figure 4-79 illustrates an example that depicts determination of present position using astrotracker and Doppler information.

The INS sends a true heading of 040 degrees to the computer, and the Doppler registers ground speed of 707 knots and adrift of 5 degrees right. The true heading and drift are combined in the computer to produce a value of track; in this case, 045 degrees.

The ground speed can be resolved around the direction of track to produce values of ground speed to drive the latitude and longitude counters. In this case, the ground speed north and east is 500 knots. Though this process seems basically simple, a few corrections must be applied to the ground speed components before they are sufficiently accurate for present position drive. These corrections, done within the computer, include such things as compensation for convergence of meridians and for the imperfect shape of the earth.

Determining Heading to Destination

Another question the navigator often faces is, "What is the heading to destination?" This question is also answered by many navigation computer systems. The computer first computes the required track, either rhumb line or great circle. To do this, it computes the direction and distance from the present latitude and longitude to the destination latitude and longitude. The present track taken from the inertial system in this instance, is then compared to the required course to destination; the difference is a heading correction. Ground speed may be applied to the distance to destination, and a time-to-go may be computed to provide a continuous ETA. Figure 4-79 illustrates some of the computer outputs associated with a typical INS. It is interesting to note that many of the new computers are equipped with an automatic troubleshooting system. In the event of malfunction, the system may automatically shut down or downgrade its mode of operation and display the location of the actual malfunction and some possible causes.

NAVIGATION PANELS

The navigation panels makeup the greatest part of the computer system visible to the navigator. Panel appearance and operation vary with each computer system. The multitude of counters, dials, switches, buttons, control knobs, and selectors give the navigator maximum use and control of the system. Selectors that determine which sensors will be used and which readouts will be given permit the navigator switch from one mode of operation to another, as shown in figure 4-80.

The computer system aids the navigator in other ways. Most modern computers have limits built into them so they will not accept unreasonable information. For instance, if the coordinates of a fix point are set 1 degree of latitude in error, the computer rejects the fix

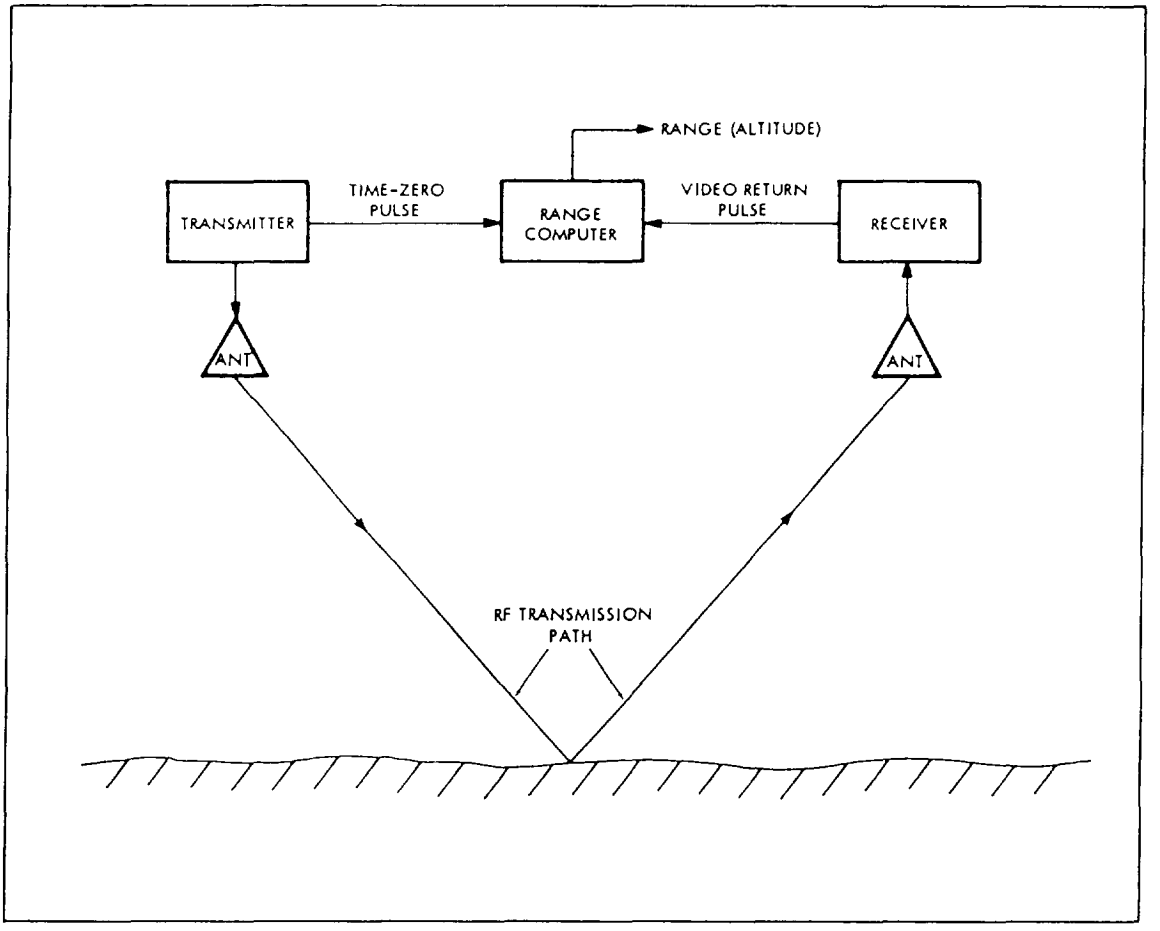


Figure 4-81.-Basic elements of radar altimeter system.

because the information is totally incompatible with information already in the computer. A rapid change in ground speed from a sensor might be rejected, and that sensor output no longer used because it would be considered unreliable.

So far in this discussion, only basic navigation has been considered. A sophisticated computer system can solve ballistic problems and automatically release bombs and missiles. If the system is installed on a transport-type aircraft, cargo drops and notification of bailout time to paratroops can be controlled by the navigation computer.

All computer systems do not contain all the sensors or have all the capabilities described above. The mission requirements of the aircraft dictate what the computer system should include. With advancements in science and engineering, automatic computer systems have increased capabilities and uses. The systems will become lighter and more compact, thereby increasing the practicability of installing them on more aircraft.

ELECTRONIC ALTIMETER SYSTEM

Learning Objective: *Recognize the basic operating principles for the electronics altimeter to include transmitters and receivers.*

Airborne electronic altimeters (commonly called radar altimeters) are absolute altimeters because they measure and indicate the height (altitude) of the aircraft above the terrain, rather than with respect to sea-level, as do barometric altimeters. They may operate on frequency-modulation principles (FM), pulse modulation principles, or a combination of both. Most modern radar altimeters use pulse-modulation principles.

Because they operate by transmitting and receiving radio-frequency (RF) energy, they are accurate over all types of terrain and under all types of weather conditions.

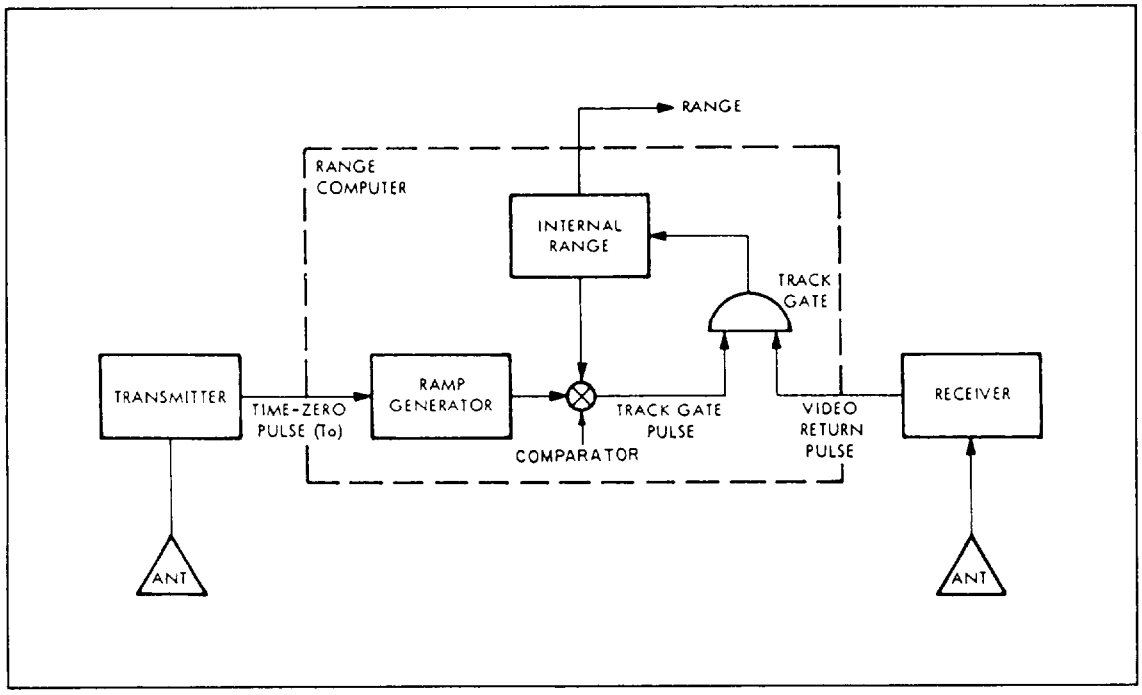


Figure 4-82.-Basic closed loop of range computer.

BASIC SYSTEM PRINCIPLES

A typical radar altimeter system operates on the accurate timing of the interval required for an RF pulse to travel from the aircraft's transmitting antenna to the terrain below, and return to the aircraft's receiving antenna. The system converts the time interval to a range signal that is used as the input to various readout circuits. Basic functions of the radar altimeter systems are shown in figure 4-81. When the transmitter fires, it transmits a pulse of RF energy out the transmitting antenna, and at the same time, generates a time-zero (T-zero) pulse that is used to start the measurement of the transmission-reception time interval. The time measuring circuit is commonly called the range computer or tracker. When the RF return (echo) pulse arrives at the receiver, a video return pulse is generated. A track gate (which is generated in the range computer by the T-zero pulse) is positioned on the leading edge of the video return pulse, which, in turn, causes the output of the range computer to be an analog signal that represents range or altitude. The output of the range computer may also be in the form of a bit digital word, which is used by some system readout indicators.

Closed-Loop Tracking

Typically, most radar altimeter receiver-transmitter units use a closed-loop tracking system contained in the unit's range computer. See figure 4-82. For explanation purposes, we'll analyze the closed-loop tracking system used by the APN-194 range computer. The APN-194 maximum altitude readout is 5,000 feet.

Each time the transmitter fires, a time zero pulse is generated. This pulse triggers the ramp generator, which generates a ramp voltage that varies linearly from 0 to 28 volts in 10.17 μsec. (Time required for RF energy to travel 10,000 feet; that is, to echo from a 5,000-foot altitude.)

At the same time, an internal range voltage is being swept from 0.5 to 32 volts at a relatively slow rate (2 to 3 times a second). When these two voltages coincide at the comparator, a track gate pulse is generated and appears at one input to the track gate. The level that the ramp voltages reach at the time the track gate pulse is generated is proportional to altitude or range. When a track gate pulse and a video return pulse occur simultaneously (in time) at the track gate inputs, a track gate output pulse is generated, which controls the value of the internal range voltage. The internal range voltage at the instant the track gate output pulse is generated represents the radar range (altitude). This output is applied to an indicator readout unit servo system, which drives the indicator needle.

When the track gate pulse is applied coincidentally with the video return pulse, the track gate output consists

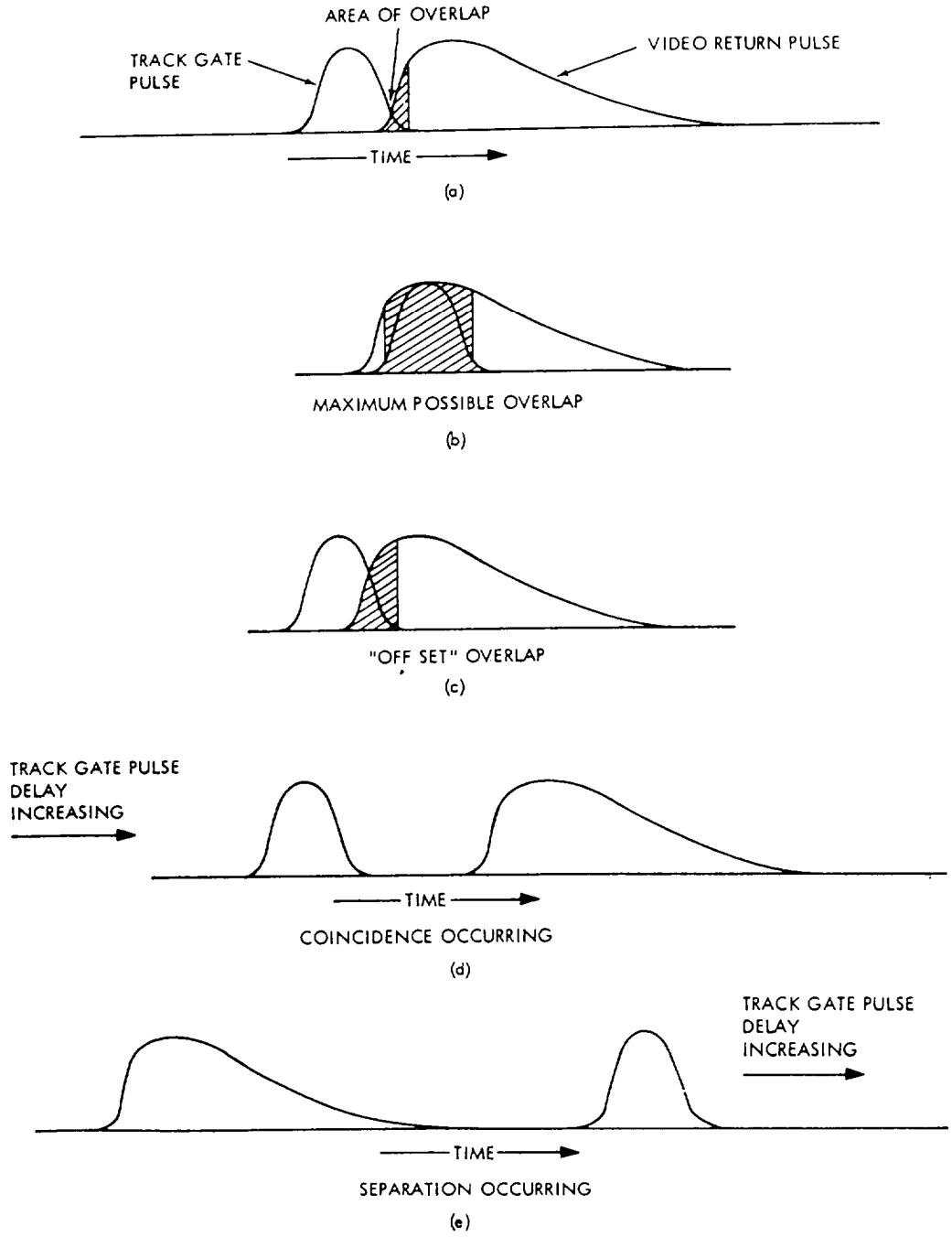


Figure 4-83.-Various pulse relationships.

of the overlapping portion of the track gate pulse and the video return pulse. Figure 4-83 shows various possible time relationships of the two pulses. When the altimeter system is properly tracking the RF echo, the track gate output (shaded area) will be as illustrated in figure 4-83, view (c). The track gate output will consist of pulses, at a rate of 8,500 pulses per second (transmitter firing rate), with individual values (amplitude) equal to the overlapping (shaded) portion of the two pulses. The track gate output pulses are then processed to maintain a constant overlap of the track gate and video pulses. If the RF transmission path distance should change (height above terrain), the track gate output pulse values would also change. This change in the track gate output pulse will change the internal range voltage, which, in turn, will change the position (time) of the track gate. (The track gate pulse would occur at a different time because the ramp generator would have a different time to reach the new voltage value of the internal range voltage.)

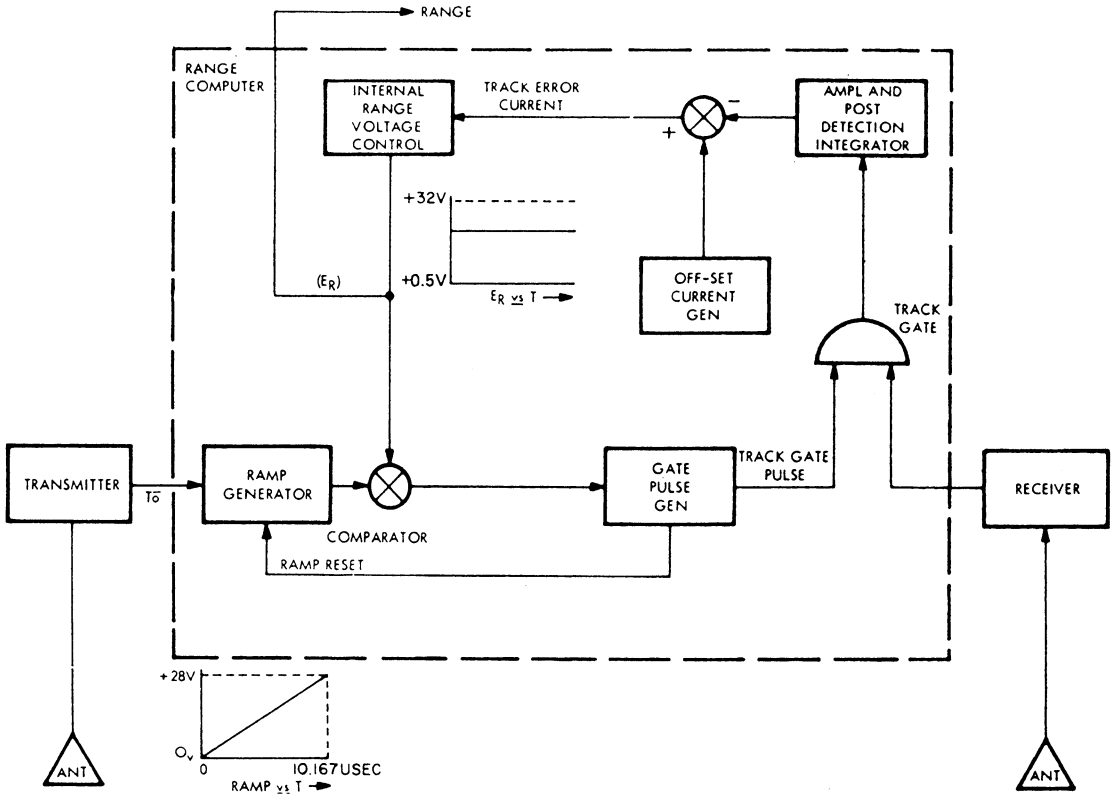


Figure 4-84.—Radar altimeter system-tracking elements.

Leading-Edge Tracking

You will notice that the trailing edge of the track gate pulse intercepts the leading edge of the video return at the approximate midpoint. This characteristic is called "leading-edge tracking." By controlling the amplitude of the video return pulse so that it has a constant amplitude under all conditions (altitude, attitude, terrain, etc.) and by controlling the position of the track gate with respect to the leading edge of the return video, the radar altimeter system is made essentially independent of all characteristics except the clearance of the aircraft over the terrain.

Figure 4-84 shows the basic functions of the leading-edge, closed-loop portion of the range computer. As previously mentioned, a track gate pulse is generated whenever the ramp generator voltage and the internal range voltage are equal at the comparator. The internal range voltage can be changed from 0.5 volt to 32 volts only relatively slowly by track error current inputs, and will hold its value whenever the track error current becomes zero. The ramp generator voltage also changes from zero volts to 28 volts. However, this voltage changes linearly in 10.17 microseconds, as previously explained. This ramp is generated each time the altimeter's transmitter fires (8500 PPS). Figure 4-85

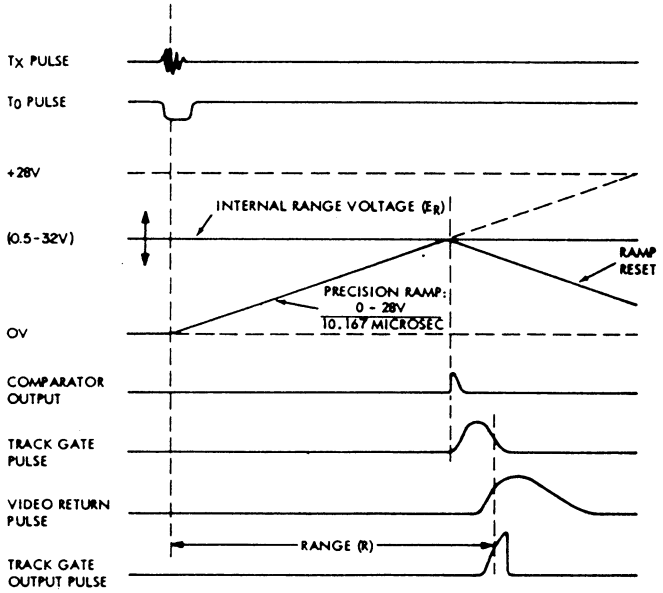


Figure 4-85.—Developing track gate output.

shows the sequence of events and the voltage relationships at the comparator during one transmission cycle. A comparator output pulse is generated each time the ramp generator and internal range voltages are equal. When the comparator output pulse occurs, a track gate

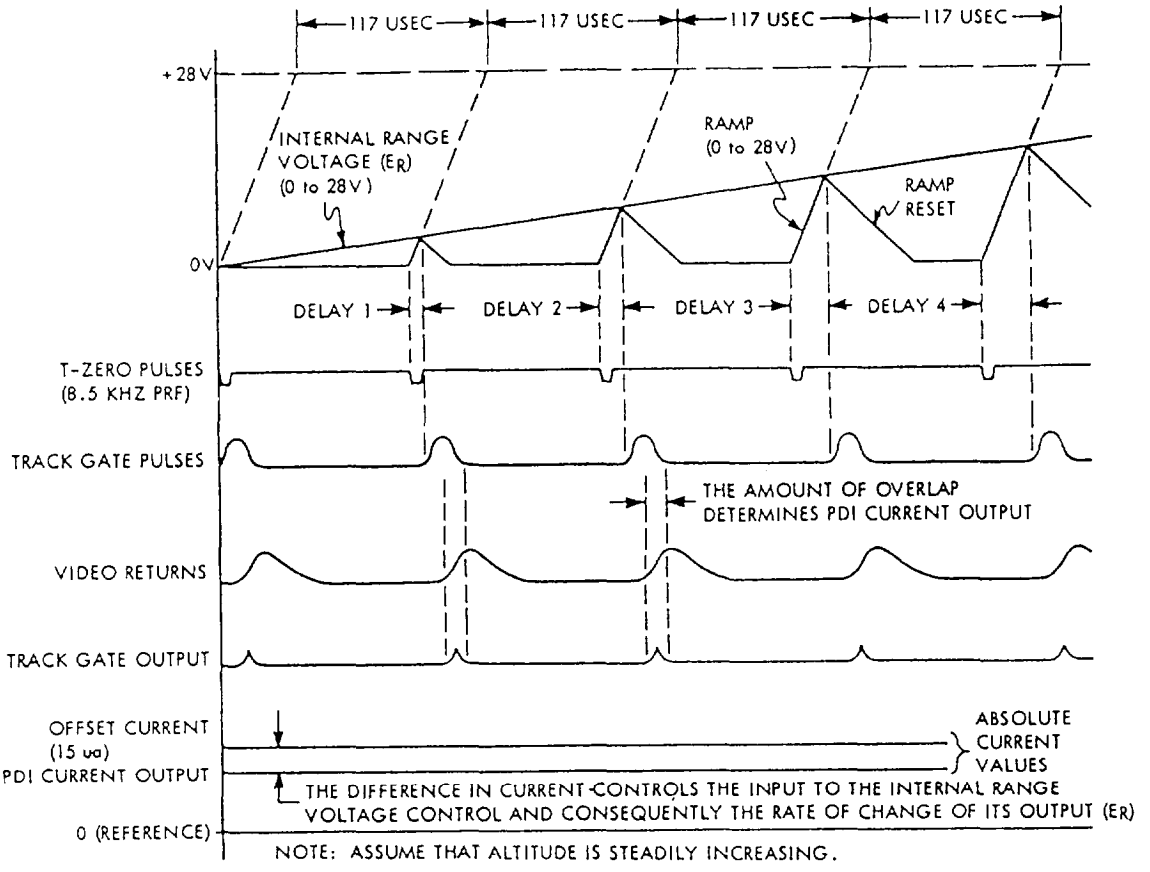


Figure 4-86.-Track loop relationships.

pulse is generated. Each time a track gate pulse is generated, a ramp reset signal is also fed back to the ramp generator commanding the generator to return to zero.

To keep the trailing edge of the track gate pulse intercepting the midpoint of the video return pulse, the video leading-edge output pulses from the track gate are amplified, filtered, and summed with an off-set current to produce an error current to control internal range voltage. See figure 4-85. When the track gate pulse overlaps the leading edge of the video return signal by the amount desired, the output of the amplifier and post detection integrator (PDI) will be 15 microampere of current. The off-set current is a negative 15 microampere of current. Therefore, under ideal conditions, the track error current to the internal range voltage control will be zero. Consequently, the time delay used in generating the track gate value of the internal range voltage will not change.

When the altitude of the aircraft increases gradually, the amount of overlap between the track gate pulse and video return pulse decreases (video arrives later) with each succeeding return video pulse applied to the track gate. See figure 4-86. Consequently, the output of the amplifier and PDI function decreases from 15 microampere, and the track error current out of the summation network becomes negative (some negative portion of the negative off-set current). This negative input to the internal range voltage control is phased to cause the internal range voltage to increase, and more time will be required for the ramp generator voltage to equal it. The result is that the track gate pulses follow the leading edge of the received signals at the track gate.

When the altitude of the aircraft decreases, the overlap of the track gate pulse with the video return pulse increases. The signal arrives sooner, so more of the video is in the gate. The output of the amplifier and PDI function then becomes greater than the 15 microampere. The track error current becomes positive out of the summation network causing the internal range voltage to decrease. The ramp generator voltage will reach this new value sooner. Therefore, the track gate pulse will be generated sooner and, with decreasing altitude, the echo signal will still be tracked.

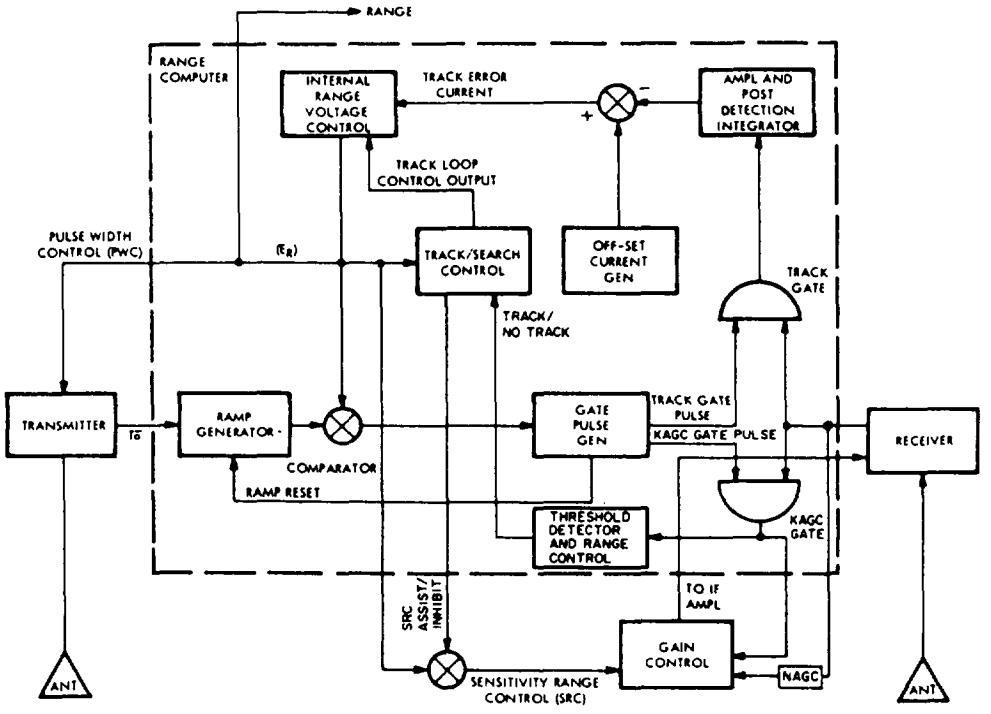


Figure 4-87.-Gain and track search mode control.

Gain Control

The accuracy of the radar altimeter altitude measurement is dependent not only on the slope and linearity of the ramp generator voltage, but also on the consistency of the rise time of the video return pulse. The purpose of an altimeter's gain control system is to keep the rise time of the video return pulse the same under all received signal strength conditions. Figure 4-87 shows the relationship of the gain control function and the tracking/search mode control function. There are two system gain control functions: signal strength controlled, and range controlled. The signal controlled gain consists of noise automatic gain control (NAGC) and keyed automatic gain control (KAGC). Range controlled gain consists of sensitivity range control (SRC) and sensitivity range control assist.

The receiver's IF gain control system is phased so that the IF gain control system is the internal range voltage that is used to drive the sensitivity range control (SRC). Its gain schedule is a nonlinear function of altitude, decreasing the receiver's sensitivity only at very low altitudes to prevent false tracking of the transmitter antenna leakage pulse (stray radiation). There are two gain controlling sources from the range computer: NAGC and KAGC. The NAGC is a wide band control used to sense all noise and video signals from the receiver's IF amplifier. It will limit the IF gain to maintain a usable signal-to-noise ratio. The KAGC circuit is a fully keyed (gated) circuit. That is, it allows only those IF signals that occur during the KAGC pulse to control IF amplifier gain. The KAGC pulse is generated simultaneously with the track gate pulse. However, the KAGC pulse is wider than the track gate pulse; therefore, it overlaps much more of the video return pulse than does the track gate pulse. It must include the video pulse peak. The function of the KAGC circuit is to control the receiver's IF gain so that the peak of the video pulse remains constant under all conditions of signal strength.

The sensitivity range control (SRC) is assisted during the search mode of operation. Because the IF gain control circuit has a time constant of approximately 50 milliseconds, it cannot react fast enough to reduce the IF gain during the rapid 20-millisecond retrace cycle of the internal range generator. If not compensated, the increased sensitivity would allow the system to begin tracking antenna stray radiation signals because the IF gain would be relatively high as the track/search control caused the internal range voltage to begin its outbound sweep. Therefore, during the retrace, the SRC (provided by an output from the track/search control) forces the IF gain control to reduce the IF gain; consequently, receiver sensitivity before the retrace ends and the outbound sweep begins.

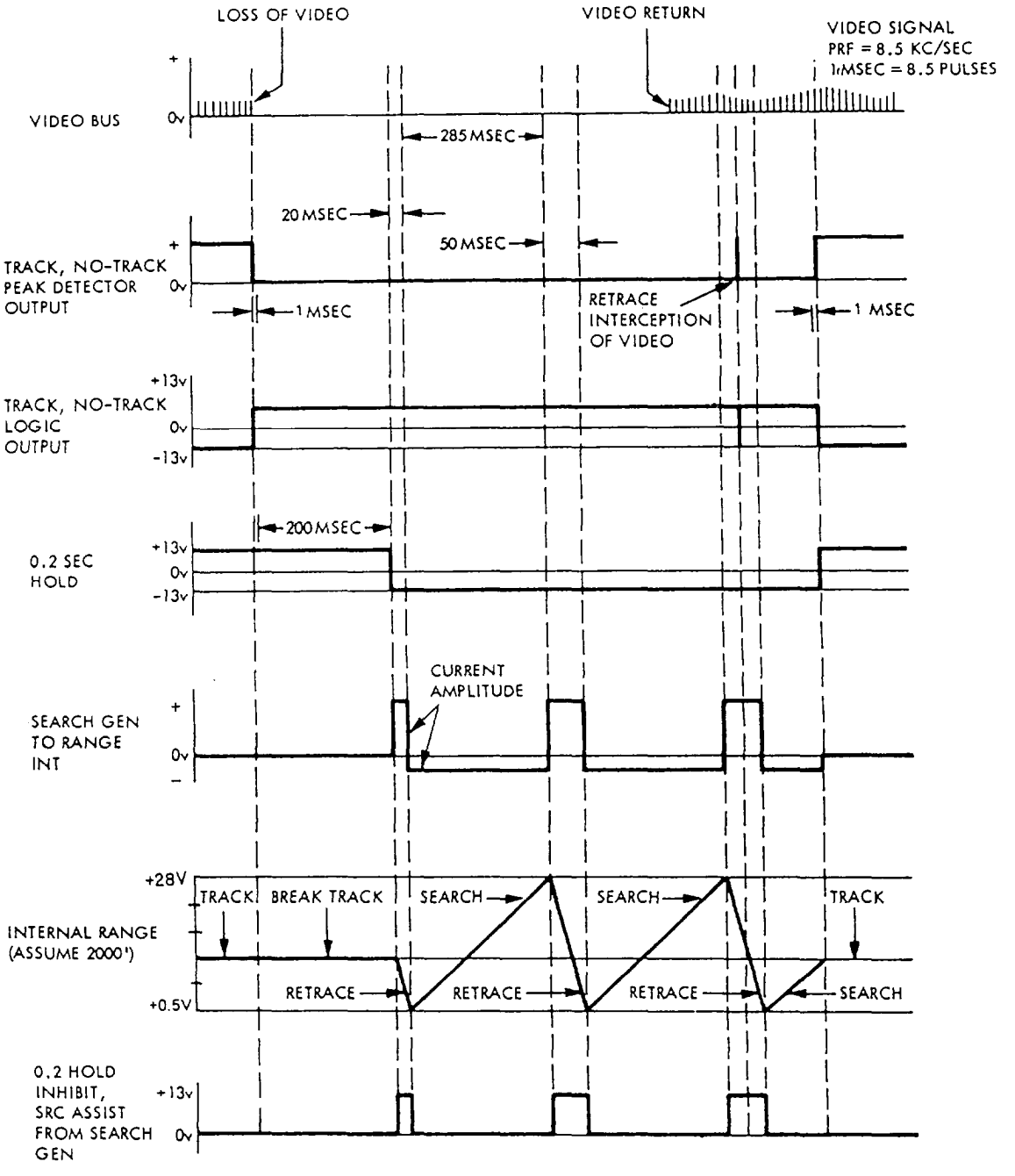


Figure 4-88.-Search and acquisition waveforms.

Mode Control

The previous discussion has assumed that the system was tracking an echo pulse. In actual operation, the system is either in an automatic mode or the self-test mode. The automatic mode contains three submodes: track, search, and memory, during which the system searches for an echo pulse or tracks that pulse. The self-test mode is manually commanded and fully exercises the automatic mode, causing the system to acquire and track a synthetic target and display a 100-foot range. The automatic mode control system receives its input from the KAGC circuit and return pulse. If 8 to 10 consecutive video return pulses have sufficient amplitude, the mode control will not modify the tracking mode if the system is tracking, and will switch to track if it is searching. The mode control first switches to the tracking operation when 8 to 10 consecutive synchronous video pulses, of 6 volts amplitude or greater, are applied to the KAGC gate. The

time constant of the track/no-track detector of the track/search control circuit is selected so that it will require 8 to 10 pulses before its output will exceed the tracking threshold. This prevents the system from being switched into the track mode by random pulses from other systems operating in the area

The mode control has a 0.2-second hold condition to provide for momentary drops in signal strength. If the video return pulse does not return to sufficient strength within the 0.2-second interval, the mode control enables a memory mode of operation. The memory mode lasts for approximately 1 second, during which time the internal range circuits search for a new target, and the last valid external range output is maintained If a new target is detected during the memory mode, the system reverts to the normal track mode. If a target is not detected, the system switches to the search mode. During the search mode, the track/search control overrides and controls the track loop. The track loop current is removed from the internal range control circuits, and the track/search control forces the internal range voltage through its limits of 0.5 to 32 volts at about three times a second (the search out rate is about 15,000 to 20,000 feet per second). This causes the KAGC and track gate pulses to run through their full range of delays, seeking to find the video return pulse.

During the search retrace portion of the cycle, the track/search control is inhibited from switching to the track mode of operation. This assures that the system will not acquire a secondary target (one having greater range than the primary target) during the retrace cycle. Figure 4-88 shows the search and acquisition waveform relationships. When video return pulses of adequate amplitude return, the KAGC gate pulses will overlap them. After successfully sampling several video pulses, the command to the track/search control causes it to remove all its overriding controls, enabling the normal track mode of operation.

ALTIMETER TRANSMITTER

The altimeter transmitter module contains the circuitry necessary to generate and control the RF energy pulse, which is needed to detect that pulse and supply a corresponding time-zero pulse to the range computer. Basically, the circuits are those of a typical radar transmitter.

ALTIMETER RECEIVER

The altimeter receiver module contains the circuits necessary to produce a usable video signal for the range computer. It is typical of most superheterodyne receivers.

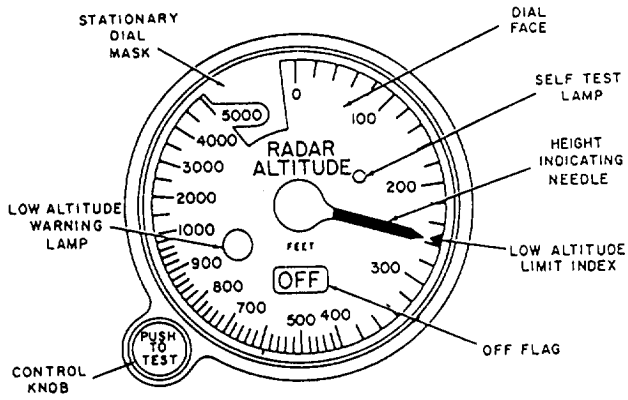


Figure 4-89.-Front view of typical radar altimeter height indicator.

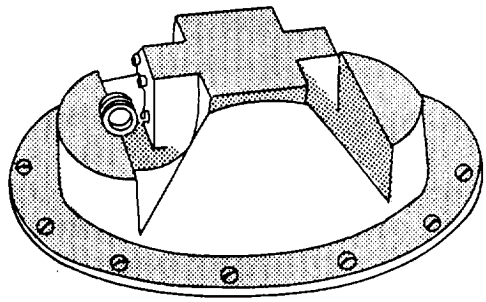


Figure 4-90.-Typical radar altimeter flush-mounted antenna.

ALTIMETER HEIGHT INDICATOR

The height indicator is both the control box and height readout for the radar altimeter system. Figure 4-89 shows the front view of a typical altimeter height indicator. It is a null-balancing servo device, which operates from the radar range voltage (analog signal) from the R-T range computer, and a reference voltage also supplied by the R-T unit's power supply. The indicator also contains a low-altitude warning circuit, which is energized whenever the height indicating needle reads lower than the altitude selected on the low-altitude limit index. The warning circuit will illuminate the low-altitude warning light. In some indicators, a 400-Hz tone to the headset is also supplied in the event of low altitude.

TRANSMITTING AND RECEIVING ANTENNAS

The transmitting and receiving antennas are identical. They are flush-mounted, dielectric-loaded, flared-horn type antennas. See figure 4-90, Operation is

in the 4250- to 4350-MHz range (typical frequency range of radar altimeters).

Interference Blanker

An interference blanker is required by a radar altimeter system only if the receiving and transmitting antenna isolation is less than 85 dB. It consists of RF attenuators, RF isolators, and a pulse-shaping network. Its sole purpose is to attenuate the RF signal, resulting from direct antenna leakage, to a level that is below the tracking capability of the altimeter system. This prevents possible zero-altitude indication (zero-lock) above the altimeter's maximum altitude (5,000 feet for the APN-194).

System Operation

Refer to figure 4-89 during the following discussion: The control knob is the system's sole operating control. In the fully counterclockwise position (indented), the power is "off." Turning the control clockwise past the indent turns the system "on." After a short warm-up period, the OFF flag will disappear, indicating the system is on. If you further adjust the control knob, it will adjust the low-altitude index needle. The altimeter should lock on at 0-6 feet of altitude, depending on the particular aircraft.

The "self-test" feature is initiated by pushing in on the control knob. Most self-test circuits will cause the altimeter to read 100 feet. (This could vary.) With the indicator needle reading the self-test altitude, adjust the low-altitude index to some altitude reading below it. The low-altitude warning light should light, and a low-altitude warning signal (if supplied) should be heard in the headset.

Once the aircraft is airborne, the operator need only to turn the system "on" and adjust the low-altitude index pointer to the desired low limit. The system will then automatically read out the aircraft's height above the terrain, and will give a low warning indication any time the aircraft drops below the preset low-altitude index needle.

RADAR ALTIMETER WARNING SET

The Radar Altimeter Warning Set (RAWS) AN/APQ-107 is used in conjunction with altimeters such as Radar Altimeter AN/APN-117, Radar Altimeter AN/APN-22, or the Electronic Altimeter AN/APN-141 previously discussed. It provides aural and visual warning signals to the pilot and copilot when any of three unsafe flight conditions exist:

1. Aircraft flies below preselected altitudes.
2. Input power to the radar altimeter fails.
3. Radar altimeter warning circuit indicates unreliability.

Functional Characteristics

The RAWS receives a signal that is a function of altitude from a radar altimeter. When the aircraft descends to a preselected altitude (designated "high-altitude index"), the incoming altimeter signal is interpreted by the warning set. The RAWS produces a signal that causes warning lamps to flash in the cockpit, and a concurrent 1000-HZ interrupted tone signal is heard in the pilot's and copilot's headsets. Both the aural and visual signals are repeated twice a second for a duration of 3 seconds. No further warning indications are supplied until the aircraft descends to a preset "low-altitude index." When the low-altitude index is reached, a warning is indicated by the warning lamps and an interrupted 1000-Hz tone sounds in the headsets again. These warning signals continue as long as the aircraft remains below the low-altitude index setting. When the aircraft ascends through the high-altitude index setting, the warning signals are not presented. It is important to note that the high and low index settings and the type of aircraft for which the RAWS has been precalibrated are indicated on the front of the set.

In addition to the altitude index warnings, the power failure (115 V, 400 Hz ac, or 28 V dc) to the radar altimeter indicator causes the warning set to produce aural and visual warning indications at a two-times-per-second rate. Another hazardous flight condition that will cause the set to produce warning signals is when the radar altimeter becomes unreliable. This condition could be caused by either the radar receivers' signal being too weak to provide reliable altitude information, or if the altimeter itself malfunctions. These warning signals, however, may be altered by a flight or altitude inhibit condition.

Warning Inhibit Conditions

A warning inhibit (overriding) condition is a flight or altitude situation that applies a bias potential to a diode gate, altering one or more signals or indications transmitted by the warning set.

The following conditions inhibit warning signals as indicated:

1. When all wheels are down and locked, the low-altitude index audible signal is inhibited.

2. When the wheels are down and locked and the altimeter is unreliable, both the audible and visual signals are inhibited.

3. Aircraft operation at a barometric altitude over 700 feet (referenced to takeoff altitude), together with an unreliable altimeter, inhibits the high-altitude index, low-altitude index, and unreliability warning indications.

4. When automatic stabilization equipment (ASE) is engaged the 3-second high-altitude index warning is inhibited. This condition has no effect on the low-altitude index warning or altimeter limit indicator warning.

NOTE: In aircraft where it is desirable to have the pilot and/or copilot adjust the critical altitude setting with the radar altimeter limit indicators, the low-altitude index is provided by the limit indicator setting. The low-altitude index warning signal is produced when the aircraft is flying below the limit set by the pilot's and/or copilot's limit indicator.

5. A weight-on-wheels condition inhibits the aural portion of the power failure warning. All other inputs to the RAWS are overridden by a power failure warning input.

6. With proper external connections, the wheels down and lock condition will inhibit the high-altitude index warning indications.

CHAPTER 5

ANTISUBMARINE WARFARE (ASW)

Chapter Objective: *Upon completion of this chapter, you should be able to identify acoustic and nonacoustic ASW systems and describe their associated maintenance procedures and equipment characteristics.*

As an Aviation Electronics Technician (I), you need a vast amount of electronics knowledge to troubleshoot, repair, and maintain equipment. The in-depth information on circuitry in the previous chapters and within the Navy Electricity and Electronics Training Series (NEETS) modules reduces this chapter's emphasis to specific circuitry and electronics peculiar to ASW systems. It is beyond the scope of this manual to train you on the complete repair of any system. The C schools and appropriate maintenance instruction manuals (MIMs) teach you to effectively troubleshoot and repair. This chapter will familiarize you with the characteristics, uses, and peculiarities of selected pieces of equipment. Basic electronics, troubleshooting, and circuit analysis are covered in the NEETS modules.

The ASW community includes acoustic and nonacoustic systems in support of the ASW mission. Several aircraft platforms supporting the ASW mission include both rotary-wing and fixed-wing aircraft. Some of the major platforms are the P-3, S-3, SH-2, SH-3, and SH-60 communities. We will use some of these aircraft systems and components as representative ASW acoustic and nonacoustic systems.

Each type of ASW aircraft is equipped with an acoustic ASW system suite. The acoustic system suite consists of passive and active types of sonobuoys deployed from the aircraft and components installed in the aircraft. Sonobuoys are used to detect sound(s) from a submerged submarine and to transmit the sound(s), via VHF radio signals, to an aircraft. The acoustic system components in the aircraft receive and process the sound-modulated VHF signals for use by ASW operators. These operators, a tactical coordinator (TACCO) and/or a sensor operator (SENSO), are considered part of the acoustic ASW system. In this chapter, we will discuss ASW acoustic platforms, their components, and the functions of each. We give special emphasis to the acoustic data processor (ADP) unit, which is the heart of the acoustic ASW platform.

ASW SYSTEMS

Learning Objective: *Identify acoustic and nonacoustic ASW systems to include platforms and general maintenance concepts.*

The ASW mission includes long-range search, detection, localization, and classification of submarines. This mission involves the use of most of the avionics equipment aboard ASW aircraft. Some of the systems supporting the ASW mission include radar, navigation, communication, and detection equipment. This chapter will concentrate on the acoustic and nonacoustic systems that have a primary ASW role.

ACOUSTIC

A typical ASW acoustic system is composed of components that detect, process, and display sonobuoy detected data for analysis. Figure 5-1 shows a functional diagram of the acoustic system components in an S-3 aircraft. This diagram is a good representation of a typical acoustic suite. The deployed sonobuoy's acoustic data is received by a sonobuoy receiver system (SRX) and is input

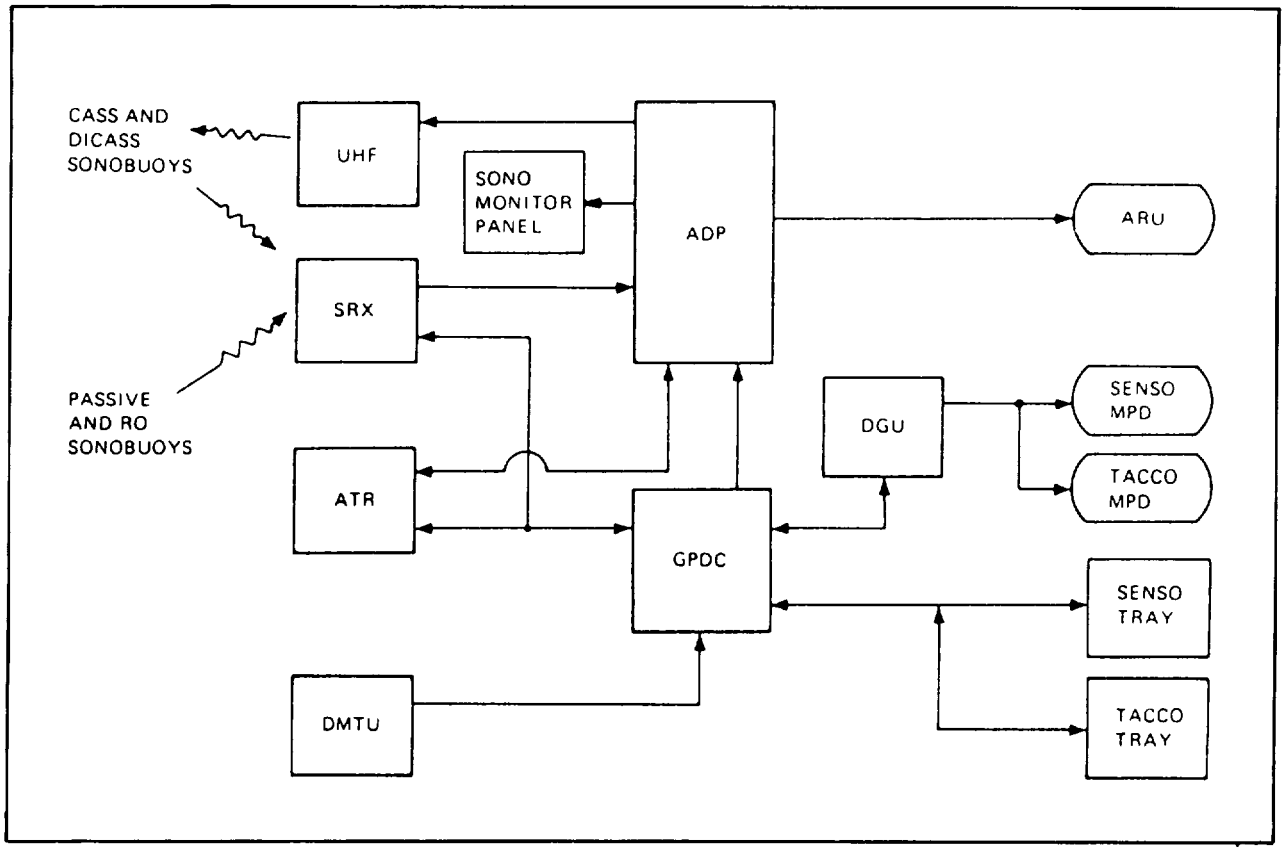


Figure 5-1.-Acoustic system components functional diagram.

into an acoustic data processor (ADP) and an analog tape recorder (ATR). The ADP processed data goes to an auxiliary readout unit (ARU) or multipurpose display (MPD) through the controls of the sensor operator (SENSO) or tactical coordinator (TACCO). Figure 5-1 is a simplified view of acoustic system components. Most platforms and acoustic suites contain similar equipment. Specific SRX such as the AN/ARR-72, AN/ARR-75, or AN/ARR-78 may vary, but they all function to receive sonobuoy acoustic data.

ASW acoustic equipment includes all the systems that relate to sound. Acoustic systems include sonobuoys and receivers, processors, recorders, and sonar equipment. Sonar is the name of the major acoustic system that includes transmitters, receivers, and signal processing. Analysis and recording of detected information is also included under acoustic equipment, and includes magnetic recorders and chart recorders. The AN/AQA-7 sonar computer recorder group, the UYS-1, and the AQS13 sonar systems are examples of ADP analyzers or signal processors. The ATR magnetic recorders include the AQH-4, AQH-7 sound recorder reproducer systems, and the ASH-27 recorder.

NONACOUSTIC

Radar, navigation, MAD, ECM, ESM, and infrared detecting sets (IRDS) are examples of nonacoustic systems. Our discussion will be limited to the MAD system. Although the nonacoustic system's purpose and use aboard other platforms may be different from the acoustic systems, the circuitry and theory are basically the same.

PLATFORMS

The major platforms for our discussion will be the P-3C, S-3B, and SH-60 aircraft. The

Antisubmarine Warfare 3 & 2, module 4, (NAVEDTRA A60-04-45-91) is an excellent reference to nonacoustic system's platform description and operation. This reference covers most of the acoustic systems.

The SH-60 acoustic suite includes the AN/ARR-75 SRX, the AN/UYS-1 ADP, a general-purpose digital computer (GPDC), and the ATR. The acoustic equipment in the P-3C includes the AN/ARR-72 SRX, the AN/AQA-7 sonobuoy recorder group ADP, the CP901 GPDC, and the AN/AQH-4 ATR. The S-3B suite includes the AN/ARR-75 or AN/ARR-78 SRX, the AN/UYS-1 ADP, the GPDC, and the AN/AQH-7 ATR.

GENERAL MAINTENANCE CONCEPTS

The maintenance concept for ASW acoustic and nonacoustic systems is the same as other avionics systems. We have organizational-, intermediate-, and depot-maintenance support. Intermediate maintenance includes automatic test equipment (ATE) and actual hot bench testing and repair. Many weapons replaceable assemblies (WRA) undergo troubleshooting at an organizational or intermediate activity, down to a shop replaceable assembly (SRA) or subshop replaceable assembly (SSRA). An AIMD usually receives the SRA and either runs ATE or bench checks the components, or, in many cases, performs both.

The technology of digital electronics has enabled the maintenance and repair of electronics to take on a computer image. Technology is increasing the speed and accuracy of repairs, and is also providing highly reliable and sophisticated equipment. Computer-age equipment lends itself to ATE testing and repair. Increasingly, sophisticated equipment and test equipment require even more experienced technicians. The Navy still has some older equipment that can only be repaired on a hot bench and many hybrid systems that use both ATE and bench testing for repair. The ATE for ASW acoustic and nonacoustic systems is basically identical to our discussion of ATE.

Any ASW acoustic system equipment repairable is routed through the local supply department and intermediate maintenance activity. The activity individual component repair list (ICRL) will indicate the repair capability and specific work center responsible for repair. All repairable will be sent to the applicable ATE work center or hot bench for test and check or indicated repair capability. The ATE work center is 650 and the major work centers for ASW are usually designated as work centers 660, 66A, and 66B.

The specific work center will troubleshoot the repairable down to another repairable SRA or SSRA, or circuit component, depending on the item's maintenance philosophy. If the philosophy calls for troubleshooting to the next lower repairable, then the faulty repairable will be removed, ordered, and routed through supply. The repairable may come back to the original work center or go to ATE for repair. In-depth AIMD procedures can be found in OPNAVINST 4790.2 (series).

BASIC ACOUSTIC ASW SYSTEM OVERVIEW

Learning Objective: *Recognize the functional relationship of acoustic ASW systems to include acoustic data processor component.*

The acoustic ASW system components and their functional relationship are shown in figure 5-1. This system is typical of the acoustic systems used in S-3A, and in other fixed-wing ASW aircraft. Notice that five data and control signal paths interface with the ADP unit, and that five paths interface with GPDC.

The sawtooth-shaped arrows represent RF signals transmitted from the UHF TX (transmitter) to the active-type sonobuoys, and RF signals transmitted from active-, passive-, and RO (range only) -type sonobuoys to the sonobuoy receiver (SRX) in the aircraft. The sound-data output of the SRX is fed to three other components: the analog tape recorder (ATR), the ADP, and the GPDC.

In the ADP (a 16-channel input, frequency-analyzing group), the analog acoustic signals detected by the SRX are converted to digital form. The ADP processes the digital signals for frequency, bearing, and range for display and for audio monitoring and recording.

ACOUSTIC DATA PROCESSOR

The ADP consists of the two half systems, as shown by the dashed line in figure 5-2. Observe that half system number 1 and half system number 2 contain almost identical components. These components are known as weapons replaceable assemblies (WRAs). The two half systems are capable of analyzing acoustic signals detected by a total of 16 sonobuoys simultaneously. Figure 5-2 shows the SRX providing eight audio signals to a signal data converter in half system 1 and eight audio signals to a signal data converter in system 2, for a total of 16 signals. The functions of the signal data converter and each of the other components are given below. Locate each component in the figure as you read about it.

Signal Data Converter

This unit accepts audio signals from the SRX. Processing the signals in the converter involves demultiplexing, heterodyning, translation, and filtering. The processed data is routed to the sonobuoy monitor panel, to the spectrum analyzer-signal generator of the signal data computer, and to the data storage magnetic drum. The signal data converter also generates built-in test equipment (BITE) signals.

Sonobuoy Monitor Panel

The sonobuoy monitor (SONO MON) panels provide the SENSO and TACCO with manual controls to detect sonobuoy target bearing and to monitor sonobuoy audio. By monitoring this panel and adjusting controls, the SENSO or TACCO is able to control the gain of the system audio outputs, the direction of steerable nulls, and the selection of sonobuoy audio.

Spectrum Analyzer-Signal Generator

This unit provides the active sonobuoy commands that are transmitted by the UHF TX

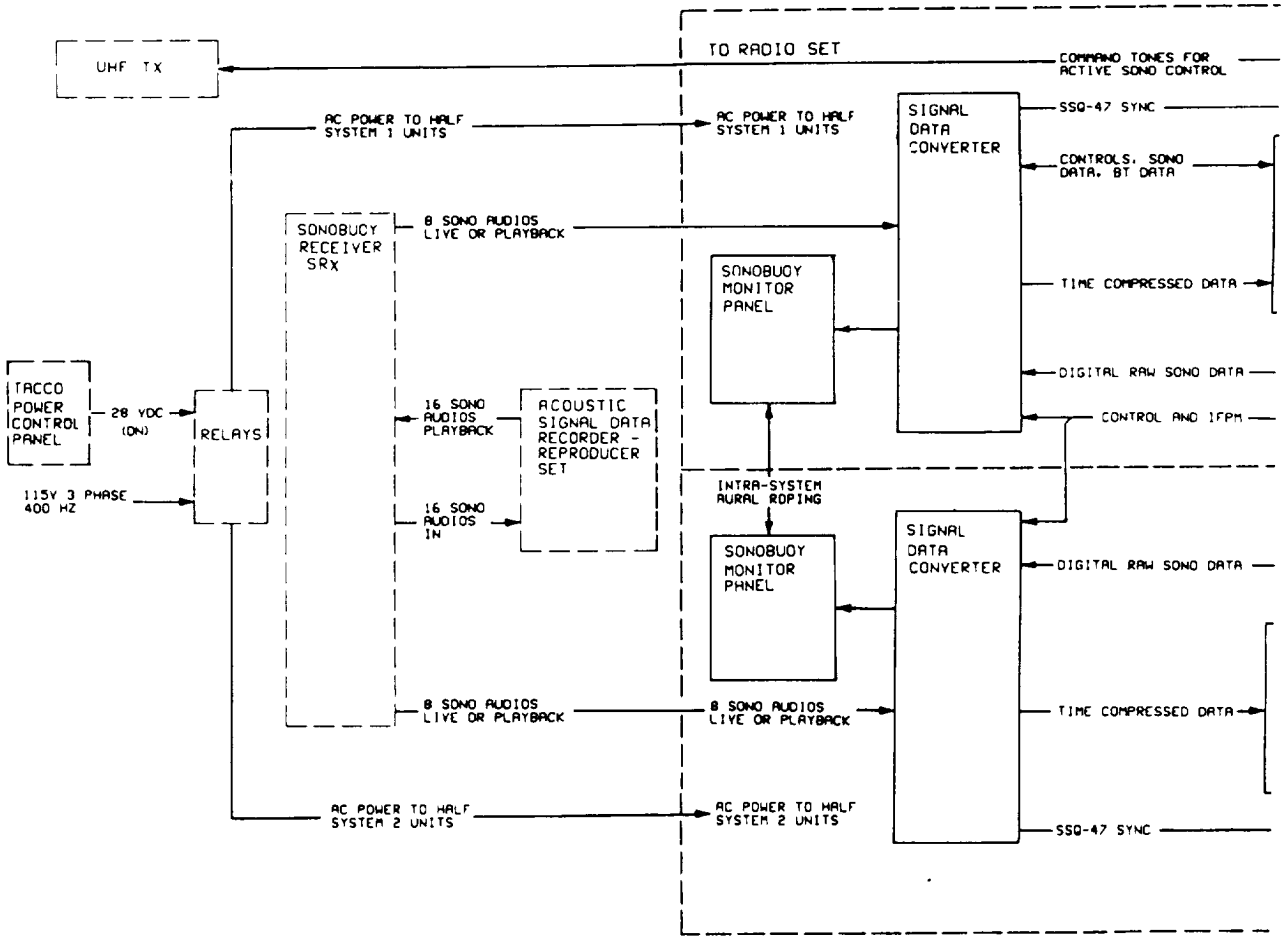


Figure 5-2.-Acoustic data processor signal flow diagram.

to displayed sonobuoys. The unit also performs narrow-band analysis, directional processing, BT processing, and frequency standardization of data received from the signal data converter unit. This unit also functions as a computer interface.

Spectrum Analyzer Converter

This unit provides interfacing with the tactical display auxiliary readout unit (ARU). The unit also performs narrow-band auxiliary readout unit analysis, directional processing, and frequency standardization of data received from the spectrum analyzer converter and the spectrum analyzer-signal generator units.

Data Storage Magnetic Drum

This unit has memory provisions for time compression, gram storage, automatic line integration (ALI) storage, computer auxiliary program storage, and ARU and multipurpose display (MPD) refresh.

Power Supply

This unit supplies electrical power for the magnetic drums and controls drum speed and sync.

Sonar Data Computer

This unit provides multipurpose display (MPD) format and interfacing, computer interfacing, drum data flow control, DIFAR bearing calculations, and ARU refreshing.

TACCO POWER CONTROL PANEL

The power switch on this panel activates the acoustic group system. When the power switch

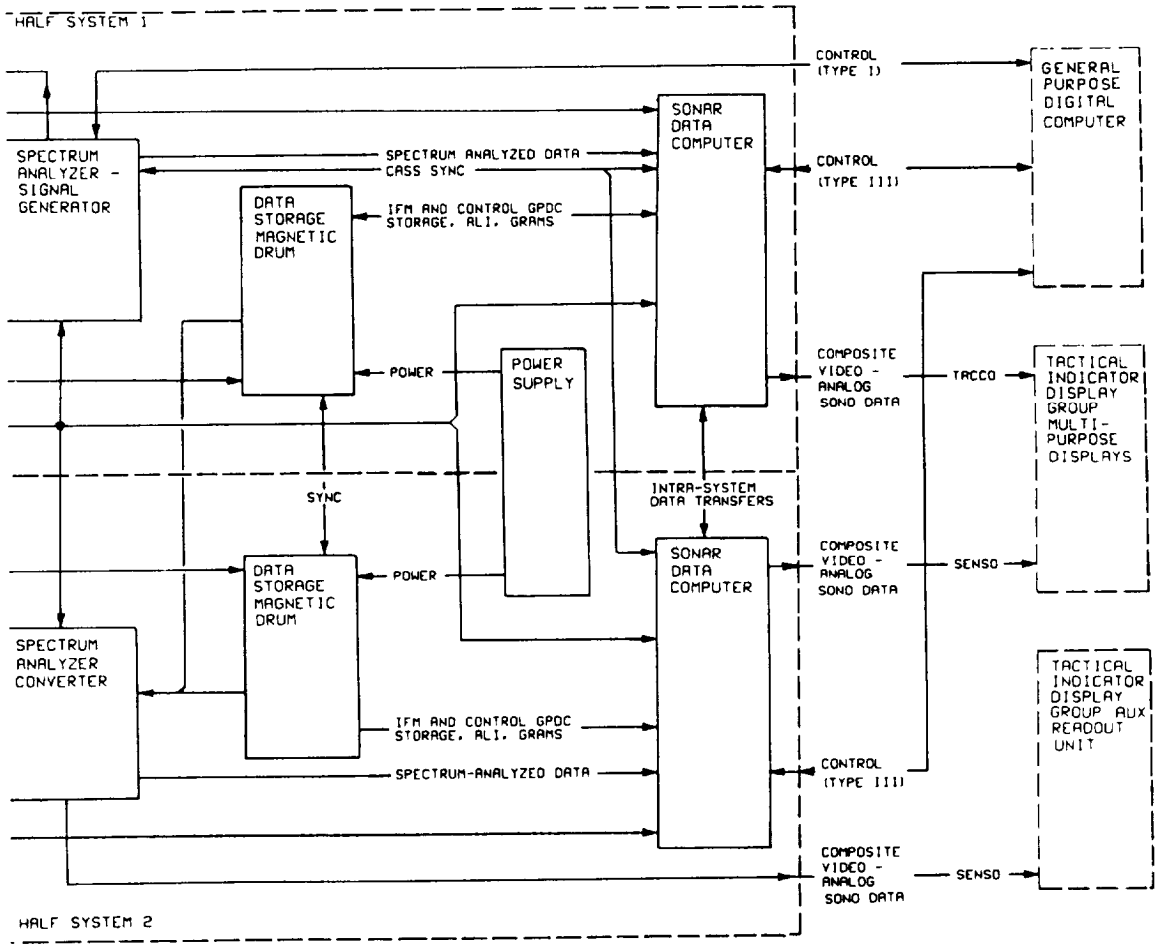


Figure 5-2.-Acoustic data processor signal flow diagram—Continued.

is in the ON position, 28-volt dc and 115-volt, three-phase, 400-hertz ac are applied through the power relays to the two half systems.

RADIO RECEIVING SET

This unit is the sonobuoy receiver (SRX). It is capable of receiving and demodulating up to 99 sonobuoy signals. Any 16 audio outputs can be selected, eight of which can be routed to the signal data converter in each half system. The 16 audio signals may be live signals or signals played back from the acoustic signal data recorder-reproducer set.

SIGNAL DATA RECORDER-REPRODUCER SET

This unit records sonobuoy audio data for later use. For example, personnel may remove the tape from an aircraft following a mission for data analysis at an antisubmarine warfare operations center (ASWOC).

GENERAL-PURPOSE DIGITAL COMPUTER

This unit provides overall control for the ADP system. It also processes information for display on the SENSO and TACCO display units.

TACTICAL INDICATOR DISPLAY GROUP MULTIPURPOSE DISPLAYS

These units display sonobuoy signal characteristics and locations. Appropriate alphanumerics, symbols, vectors, and conics are also displayed.

TACTICAL INDICATOR DISPLAY GROUP AUX READOUT UNIT

This unit is an auxiliary display unit. It provides a master display of sonobuoy acoustic characteristics.

HELICOPTER ACOUSTIC DATA SYSTEM

Learning Objective: *Recognize functions of helicopter acoustic data system components.*

The ASW helicopter acoustic data platform is similar to those used in fixed-wing aircraft. The main difference is that this sonobuoy system is designed primarily for detection and classification of submarines by providing an airborne data link between acquisition sonobuoys and a shipboard-installed telemetric data receiving set. In addition, this system is capable of processing onboard range-only (RO) sonobuoy signals to provide a permanent recording, and a means for converting scalar distances of these recordings into direct reading range indications.

Briefly, the acoustic sensor subsystem receives sonobuoy data from deployed sonobuoys. The raw acoustic data is sent to the ship, via a radio terminal set (data link), for shipboard processing, and to a spectrum analyzer (SA) for use aboard the helicopter. The subsystem then processes the sonobuoy data for display and aural monitoring aboard the helicopter. Major components of the helicopter acoustic system include a radio receiving set and a spectrum analyzer. Functions of these units are discussed in the following text.

RADIO RECEIVING SET

The SRX receives, demodulates, and amplifies sonobuoy data. The SRX consists of four VHF radio receivers. Each of the four receivers can operate on a separate channel, independent of the others. The RF signals received by the sonobuoy antennas are applied to each of the four receiver modules, where tuned filters select the signals for each module. The signals then pass through a series of amplifiers, filters, and mixers to produce the output audio signals. The output signals are applied to the spectrum analyzer (SA) and the data link. The SA produces the audio signal for the communications group to allow monitoring by the aircrew. In SHIP CONTROL, ASW mode, the sonobuoy receivers are tuned by the shipboard operator (SO) through the data link.

SPECTRUM ANALYZER

The spectrum analyzer (SA) is a high-speed signal processor designed to extract acoustic target information from both active and passive sonobuoy data. The SA determines frequency, amplitude, bearing, Doppler range, and other signal characteristics for acoustic targets.

In addition to narrow-band processing, the SA performs the following functions:

- Filtering, analog-to-digital conversion, and demultiplexing of directional sonic data

- Band separations by octaves

- Generation of low-power audio signals and of command-activated sonobuoy control signals

- Passes sonic gauge tones coincident with or with the start and stop of sonobuoy commands for down link to a ship, or when ping signals are received from range-only (RO) sonobuoys

ASW ACOUSTIC SYSTEM CHARACTERISTICS

Learning Objective: *Describe the ASW acoustic system to include sonobuoy characteristics, receiver characteristics, and general maintenance.*

The ASW acoustic system includes sonobuoy characteristics, receiver characteristics, and general maintenance.

SONOBUOY CHARACTERISTICS

Sonobuoys are aircraft-deployed, expendable sonar sets that contain a VHF radio transmitter to relay acoustic information to the deploying aircraft. Certain special-purpose sonobuoys vary from this general description and will be discussed later in this chapter.

Because the sonobuoy is expendable, it does not require periodic inspections, preventive maintenance, repair, alignment, disassembly, or testing. However, you should be familiar with sonobuoy characteristics so you can assist in repair of the associated processing and recording systems. The *AT3* TRAMAN, NAVEDTRA 12329, contains specific information and should be reviewed along with the *Sonobuoy Instructional Manual,* NAVAIR 28-SSQ-500-1.

Sonobuoys may be grouped into three categories: passive, active, and special purpose. Passive sonobuoys are used in LOFAR and DIFAR systems; active sonobuoys are used in CASS and DICASS systems. Special-purpose sonobuoys—BTS, SAR, and ATAC/DLC types—are used for missions other than ASW. Passive and active sonobuoys, systems, and acronyms, and their meanings and relationships to each other, are discussed below.

Passive Sonobuoy

The passive sonobuoy is referred to as a listen-only sonobuoy. The basic acoustic sensing system that uses the passive sonobuoy for detection and classification is known as the low-frequency analysis and recording (LOFAR) system.

LOFAR SYSTEM.— Sounds emitted by a submarine are detected by a hydrophore that has been lowered from a passive omnidirectional sonobuoy. A block diagram of a LOFAR sonobuoy is shown in figure 5-3. Data regarding the frequency and amplitude of these sounds are transmitted by the sonobuoy antenna to a receiving station. At this station, normally located on board the deployment aircraft, the sounds are analyzed, processed, displayed, and recorded. The basic LOFAR display plots the frequency of the sound waves against the intensity of their acoustic energy and the duration of the sound emission. Frequency and duration data can be displayed on a video screen and printed out. The data is also recorded on magnetic tape for storage and retrieval.

DIFAR SYSTEM.— The directional low-frequency analysis and recording system (DIFAR) is an improved passive acoustic sensing system.

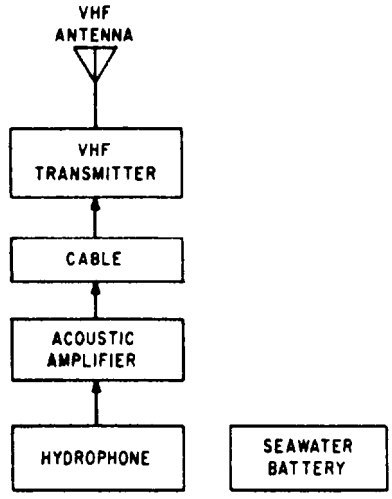


Figure 5-3.-Block diagram of the passive omnidirectional (LOFAR) sonobuoy.

Using the passive directional sonobuoy (fig. 5-4), DIFAR operates by detecting directional information and then frequency multiplexing the information to the acoustic data transmitted by the sonobuoy to the deployment aircraft. This information is processed by the aircraft's acoustic analysis equipment to compute a bearing and is then displayed. Subsequent bearing information from the sonobuoy can be used to pinpoint, by triangulation, the location of the sound or signal source.

Active Sonobuoy

The active sonobuoy is either self-timed (the sonar pulse generated by the sonobuoy at a fixed pulse length and interval) or commendable, as determined by a UHF command signal from the controlling aircraft. An active sonobuoy uses a transducer to radiate a sonar (sound) pulse that is reflected from the hull of a submarine. The time interval between the ping (sound pulse) and the echo return to the sonobuoy is measured. Taking into account the Doppler effect on the pulse frequency, this time-measurement data is used to calculate both range and speed of the submarine relative to the sonobuoy.

RO SONOBUOYS.— Self-timed active sonobuoys, known as range-only (RO) sonobuoys, are set to ping for a limited period, starting from the time they are deployed. Figure 5-5 is a block diagram of an RO sonobuoy.

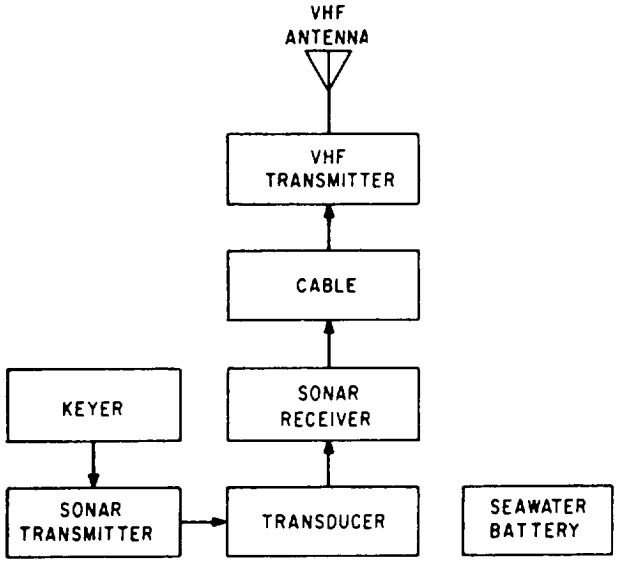


Figure 5-5.-Block diagram of the active omnidirectional self-timed (RO) sonobuoy.

CASS SONOBUOYS.— The command active sonobuoy system (CASS) allows the aircraft to deploy the sonobuoy, but the sonobuoy remains silent until it receives a command signal from the aircraft to radiate a sound pulse. This technique allows the aircraft to surprise a submarine.

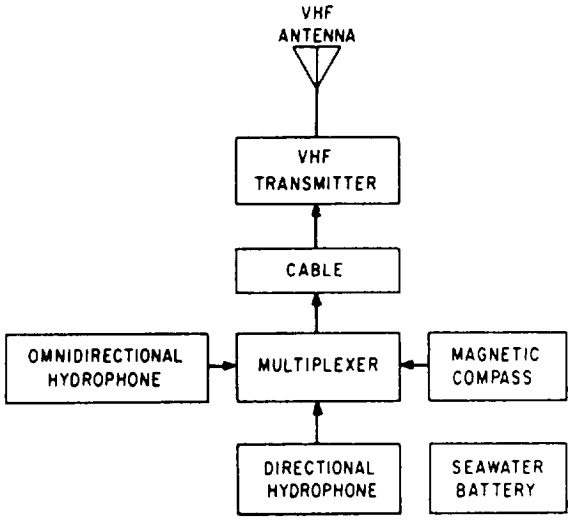


Figure 5-4.-Block diagram of the passive directional (DIFAR) sonobuoy.

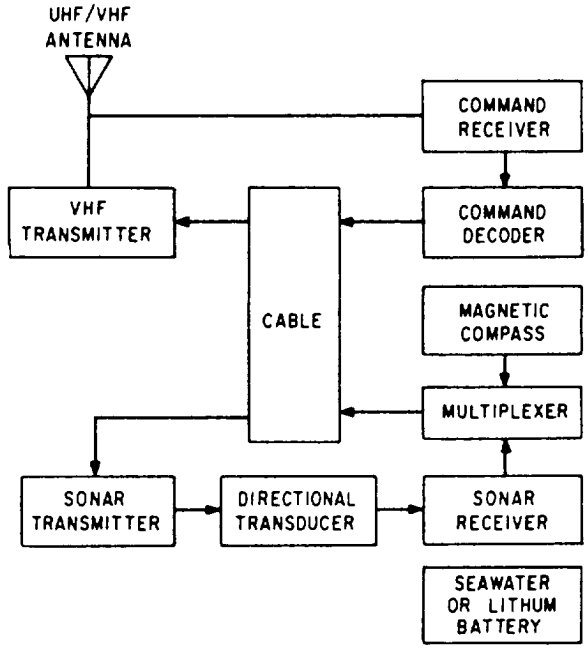


Figure 5-6.-Block diagram of the active directional commendable (DICASS) sonobuoy.

DICASS SONOBUOY.— A CASS sonobuoy, equipped with a directional hydrophore, is known as a directional commendable sonobuoy (DICASS). Figure 5-6 is a block diagram of a DICASS sonobuoy. A DICASS sonobuoy allows the aircraft acoustic analysis equipment to determine both range and bearing to a target with a single sonobuoy. DICASS sonobuoys are replacing the older RO and CASS sonobuoys.

Special-Purpose Sonobuoys

Currently there are three categories of special-purpose sonobuoys in use in the fleet: namely, the bathythermobuoy (BTS), the search and rescue (SAR), and the air transportable communication (ATAC) types. These sonobuoys are NOT designed for use in submarine detection or localization.

BATHYTHERMOBUOY.— The bathythermobuoy (BTS) (fig. 5-7) is used to measure water temperature versus depth. The water depth is determined by timing the descent of a temperature probe. Once the BTS enters the water, the probe (fig. 5-8) descends automatically at a constant 5 feet per second.

The probe uses a thermistor, a temperature-dependent electronic component, to measure the temperature. The electrical output of the probe is applied to a voltage-controlled oscillator, whose output signal frequency modulates the sonobuoy transmitter. The frequency of the transmitted signal, which is recovered at the sonobuoy receiver in the aircraft, is linearly proportional to the water temperature. The water temperature and depth are recorded on graph paper that is visible to the ASW operator.

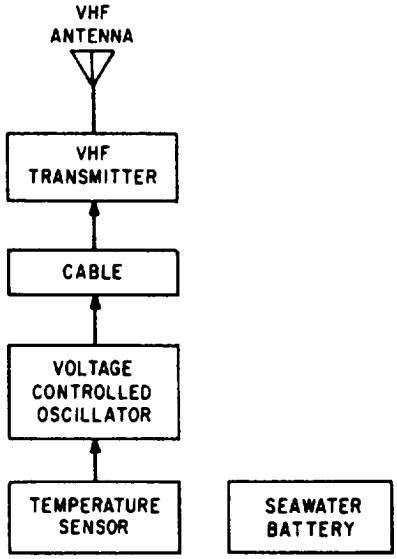


Figure 5-7.-Bathythermograph sonobuoy block diagram.

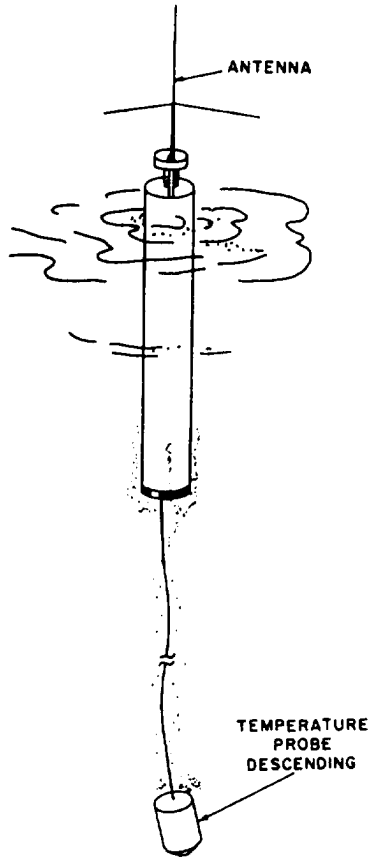


Figure 5-8.-Bathythermograph sonobuoy deployment.

SAR BUOY.— The search and rescue (SAR) sonobuoy is designed to operate as a floating RF beacon. It is used to assist in marking the location of an aircraft crash site, a sunken ship, or survivors at sea. The buoy is designed for SAR use only, and cannot be used as a sonobuoy for ASW purposes. This buoy can be launched from aircraft equipped to launch sonobuoys or deployed over the side by hand. Nominal RF output is 1 watt for 60 hours on sonobuoy channel 15 (172.75 MHz). A floating microphone is provided for one-way voice communication. The RF beacon radiates automatically and continuously, regardless of whether the microphone is used. A flashing light and dye marker are incorporated in the buoy. The SAR buoy also has an 8-foot tether

line for attaching the buoy to a life raft or a person.

Over-the-side (OTS) hand deployment of the SAR buoy is covered in detail in the *Sonobuoy Instructional Manual,* NAVAIR 28-SSQ-500-1.

ATAC/DLC.— Air transportable communication (ATAC) and down-link communication (DLC) buoys are intended for use as a means of communication between aircraft and submarines. Figures 5-9 and 5-10 are block diagrams of ATAC and DLC buoys. The ATAC buoy is commendable from the aircraft and provides up-link and down-link communications by a preselected code. The DLC buoy is not commendable and provides down-link communications only by a preselected code.

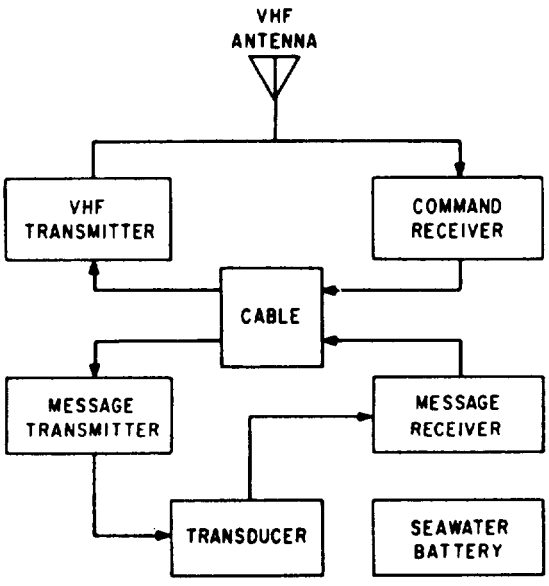


Figure 5-9.-ATAC sonobuoy block diagram.

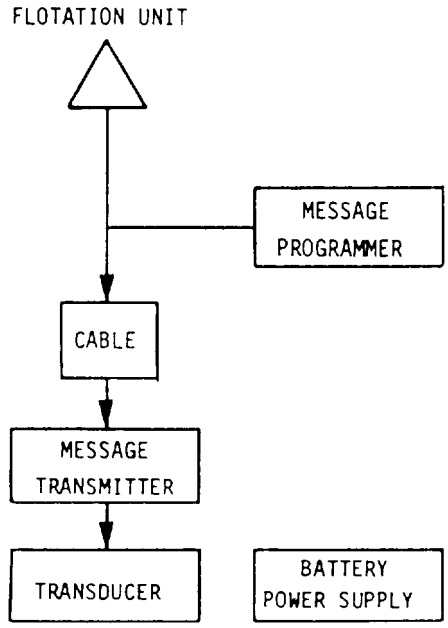


Figure 5-10.-DLC sonobuoy block diagram.

RECEIVER CHARACTERISTICS

The function of the sonobuoy receiver (SRX) is to receive the RF signals from deployed sonobuoys and detect the intelligence on the signals. The SRX also gives the intelligence to various onboard equipment for acoustic analysis and recording, and for navigation purposes. A typical SRX functional flow diagram is shown in figure 5-11. The sonobuoy data is amplified in the preamplifiers and input to the receiver assembly.

Sonobuoy Receiver Set

One commonly used SRX comprises 31 radio receivers that receive FM-modulated signals in the VHF range of 162.25 MHz to 173.5 MHz. Thus, simultaneous reception, demodulation (detection), and audio output of up to 31 RF channels are possible. These channels may each be any 1 of 31 preselected channels within the 162.25- to 173.5-MHz VHF range. Each audio output is provided in two levels—high audio and standard audio. These names refer only to the voltage level they provide for a given peak sinusoidal FM deviation of ±75 kHz.

The equipment is primarily intended for (but not limited to) installation in either fixed- or rotary-wing aircraft. Although capable of being used as an independent unit, the equipment is normally used in conjunction with some combination of several types of sonobuoys, a signal processor, and an acoustic data processor (ADP).

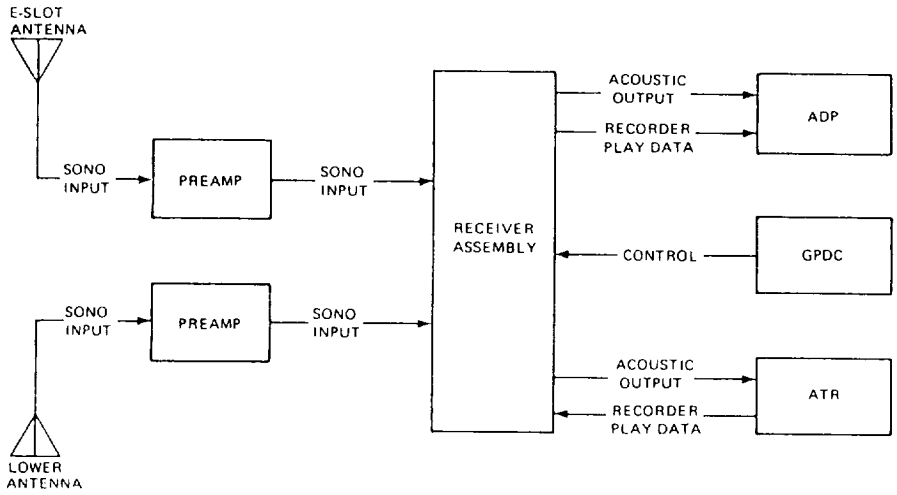


Figure 5-11.-SRX functional flow diagram.

The radio set control contains channel selector switches for each of the 31 receivers in the receiver assembly, output level meters for each selected receiver, and a digital display of the selected channels. The receiver assembly contains 31 separate but identical receiver modules, a common power supply, and a main electrical equipment chassis.

The radio set control is a small unit consisting of parts and hardware assembled together, and is a single module in itself. The receiver assembly is modularly constructed so that the 31 receiver modules and the power-supply module plug into the chassis. Each of the modules and chassis are further modularized so that printed-circuit assemblies and other small assemblies plug in or easily connect to the modules and chassis. Few parts are discretely assembled to the module chassis or main chassis.

The SRX set is controlled during operational use by the radio set control or dual channel control indicator (DCCI). This unit is mounted so it is accessible to the operator for carrying out the various operational functions. The receiver assembly unit is normally mounted in a remote area away from the radio set control. This is possible since there are no operational controls on the receiver assembly.

Newer sonobuoy receiver groups provide the capability of simultaneously receiving 20 sonobuoy signals. This is accomplished through use of 20 subassemblies, each of which may be independently and automatically tuned to any 1 of 99 sonobuoy RF channels now in use, and those that are in development for future deployment.

GENERATOR-TRANSMITTER GROUP.— The generator-transmitter group, AN/ASA-76, provides simultaneous processing and display of four channels of omnidirectional range information. The system is also a CASS. Processing control can be accomplished by either the generator-transmitter group or the digital data computer. Four channels of range information are supplied to the generator-transmitter group from the sonobuoy receivers through the SONO interconnection box. Range data is preprocessed by the generator-transmitter group, and is routed to the DIFAR system for processing and display. Command functions and sonic tones are developed within and transmitted by the generator-transmitter group to the CASS sonobuoy.

ON-TOP-POSITION INDICATOR.— The on-top-position indicator (OTPI) receiver provides reception of any 1 of the 31 sonobuoy frequencies. In conjunction with the UHF-DF (direction finder) group, the OTPI provides relative bearing to the sonobuoy. The relative bearing is displayed on the No. 1 needle of the

aircraft's horizontal situation indicator (HSI) or equivalent indicator (fig. 5-12).

Sonobuoy signals received on the UHF-DF loop antenna are lobe switched and routed to the OTPI receiver. Receiver audio is routed to an electronic control amplifier that provides drive signals to the antenna. When the received lobes are equal, the drive signals are reduced to zero and the antenna stops. A synchro generator, geared to the antenna drive motor, provides relative bearing to the selected sonobuoy. The OTPI operates in conjunction with the ARA-25 or ARA-50 direction finder group.

Sonobuoy Reference System

The sonobuoy reference system (SRS) is used to determine the positions of deployed sonobuoys relative to aircraft position. Through the use of angle-measuring equipment (AME), the SRS provides direction data to the aircraft's central computer to determine the positions of sonobuoys that are within line-of-sight of the aircraft. The information is used to update the sonobuoy positions on the tactical plot without the aircraft having to fly directly over the sonobuoys. The sonobuoy reference system consists of the components illustrated in figure 5-12. Once an

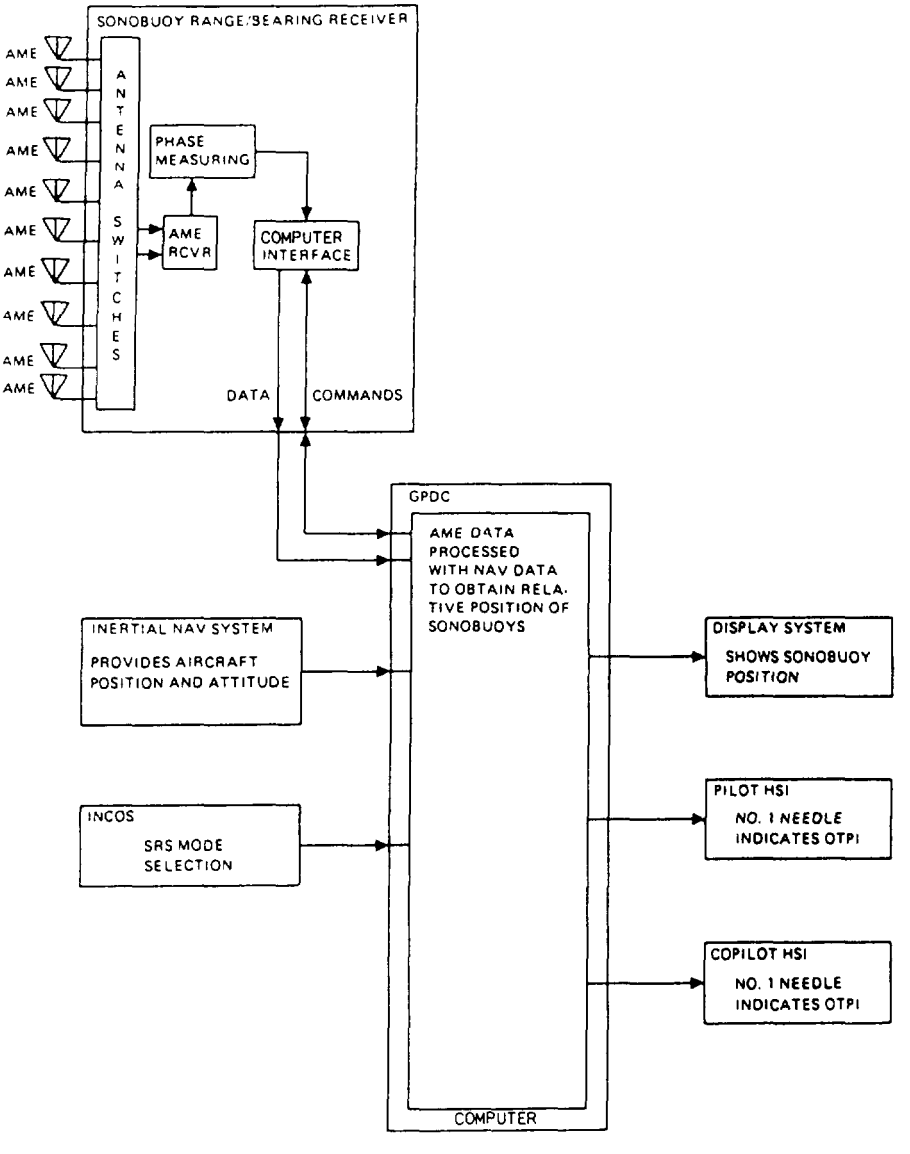


Figure 5-12.-Sonobuoy reference system (SRS) functional diagram.

eligible sonobuoy has been selected, the SRS receiver is then commanded to measure and send to the computer bearing data for the selected channel. For each antenna pair (fig. 5-13), the SRS receiver measures the difference in time-of-arrival for the selected RF signal between one antenna and the other.

In general, the antenna array is positioned so that the signal from the sonobuoy will reach one antenna before the other, and thus provide the SRS receiver-converter a signal phase difference that it can measure. (See figure 5-14.) If the incoming signal reaches both antennas at the same time, as

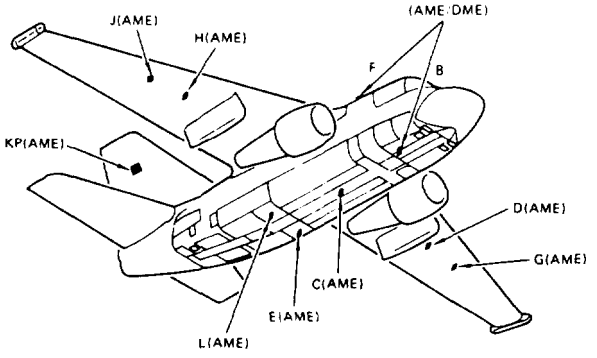


Figure 5-13.-SRS antenna locations.

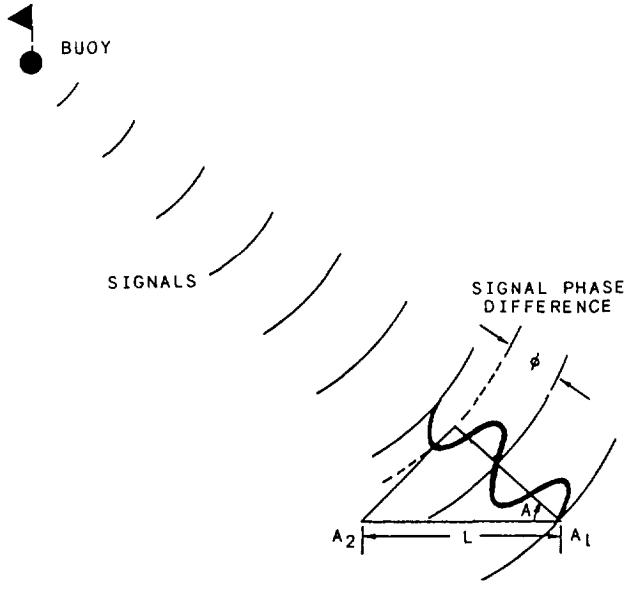


Figure 5-14.-SRS antenna array positioned so that phase difference of incoming signal can be measured.

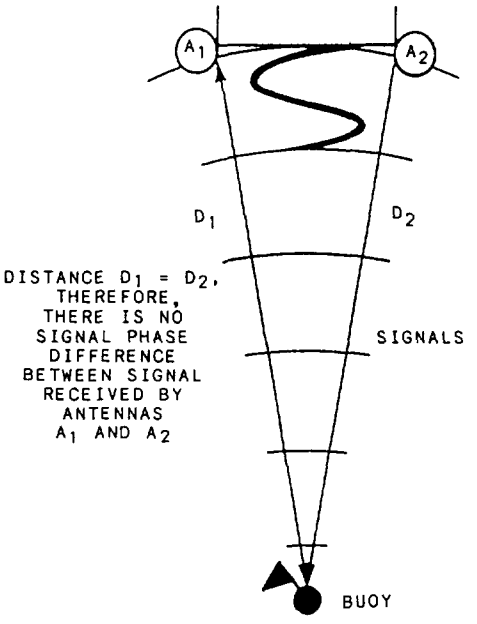


Figure 5-15.-Incoming signal arrives at both antennas of SRS array simultaneously. No signal phase difference is present to be measured.

shown in figure 5-15, the signals at the antenna outputs will be in phase with one another. Hence, there will be no phase difference for the SRS to measure. Figures 5-14 and 5-15 show that the signal phase difference at any given moment establishes the angular relationship between the antenna array baseline and the line-of-sight direction to the sonobuoy.

Before proceeding further, certain terms that will be used in this SRS discussion will be defined and their relationship to SRS will be given. To begin with, the imaginary line between the two antennas of the array is known as a baseline. The length of this baseline is expressed in wavelengths (A). Thus, the baseline value of an antenna array (expressed as a number of wavelengths or complete cycles) depends upon the frequency of the sonobuoy radio signal.

When the baseline length of an antenna array is equivalent to or less than one-half the wavelength of the incoming signal, only two possible directions from the aircraft to the sonobuoy can provide the same phase angle difference measurement. The two directions (bearings) are symmetrical to the baseline, as

shown in figure 5-16. This produces what is known as an ambiguity, a situation where there is more than one answer or solution to a problem. A 0.5-wavelength antenna provides the type of bearing ambiguity described above, which is known as a mirror-image ambiguity. The SRS resolves this type of ambiguity (that is, arrives at one answer) by using a two 0.5-wave- length antenna array configuration, with the arrays positioned 90 degrees to each other, as shown in figure 5-17. The common solution provided by antenna arrays L (solution = LOS) and T (solution = LOS) indicates the correct relative bearing of the buoy signal transmission to the aircraft.

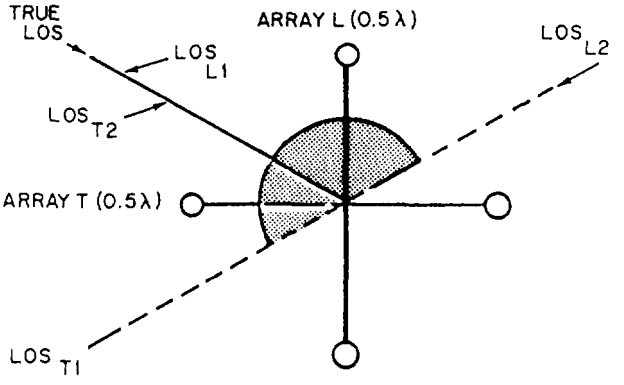


Figure 5-17.-Mirror image ambiguity is resolved when SRS measurements are made with two antenna arrays of 0.5λ or smaller that are positioned at a 90-degree angle to each other.

GENERAL MAINTENANCE

A typical sonobuoy receiver group system (fig. 5-18) like the AN/ARR-72 includes the SG-791/ARR-72(V) acoustic sensor signal generator (ASSG), which simulates sonobuoy signals aboard the aircraft. O-level maintenance personnel use the ASSG as a piece of built-in test equipment for the ARR-72 miniature sonobuoy receiver system. The ASSG provides the necessary modulated signals for verifying proper operation of the AN/ARR-72 system. The ASSG aids in troubleshooting by sending RF output signals to three points in the aircraft system. In the EXTERNAL mode, a signal is routed to the antenna via the preamp and provides end-to-end testing. The PREAMP mode routes a signal directly to the preamp and eliminates the antenna from the system. The RCVR mode bypasses the antenna and preamp and sends a signal directly to the receiver input. The organizational technician troubleshoots the system to a repairable component or faulty ASSG and orders it through supply. It is very important that the I-level technician understands the critical necessity for proper ASSG operation.

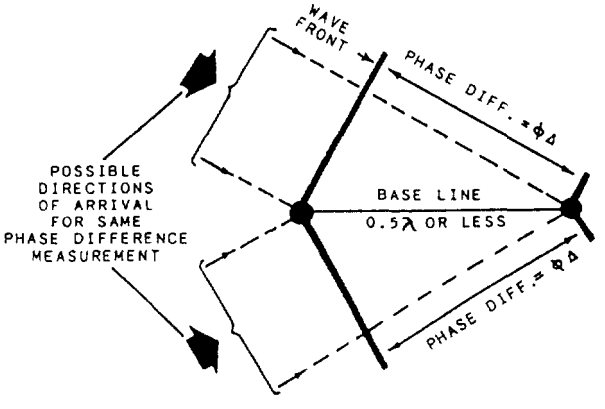


Figure 5-16.-Mirror image ambiguity occurs when SRS measurements are made with one antenna array that is 0.5λ or smaller.

SG-791/ARR-72 Acoustic Sensor Signal Generator (ASSG)

The maintenance philosophy on the ASSG includes I-level repair and alignment. The ASSG is a typical signal generator that uses voltage-controlled oscillators (VCO), summing amplifiers, and modulator circuitry for signal generation. The specific mode of operation determines which circuits are enabled to provide the desired output from the ASSG.

If an ASSG cannot be properly aligned to provide the required output signals within the tolerances, you will have to use your bench to further troubleshoot the system. Some of the circuits and circuitry within the SG-791 includes noise generators, oscillators, summing amplifiers, modulators, and crystals. The AN/ARR-72 MIM for the SG-791, NAVAIR 16-35 SG791-1, will provide the information for alignment and further troubleshooting.

The ASSG generates RF signals and includes normal LOFAR signal generation, broadband (BB) LOFAR, DEMOD, active RO/FR, and BT signal generation. The ASSG also includes circuitry for external modulation (EXT MOD), which allows external modulation of the RF

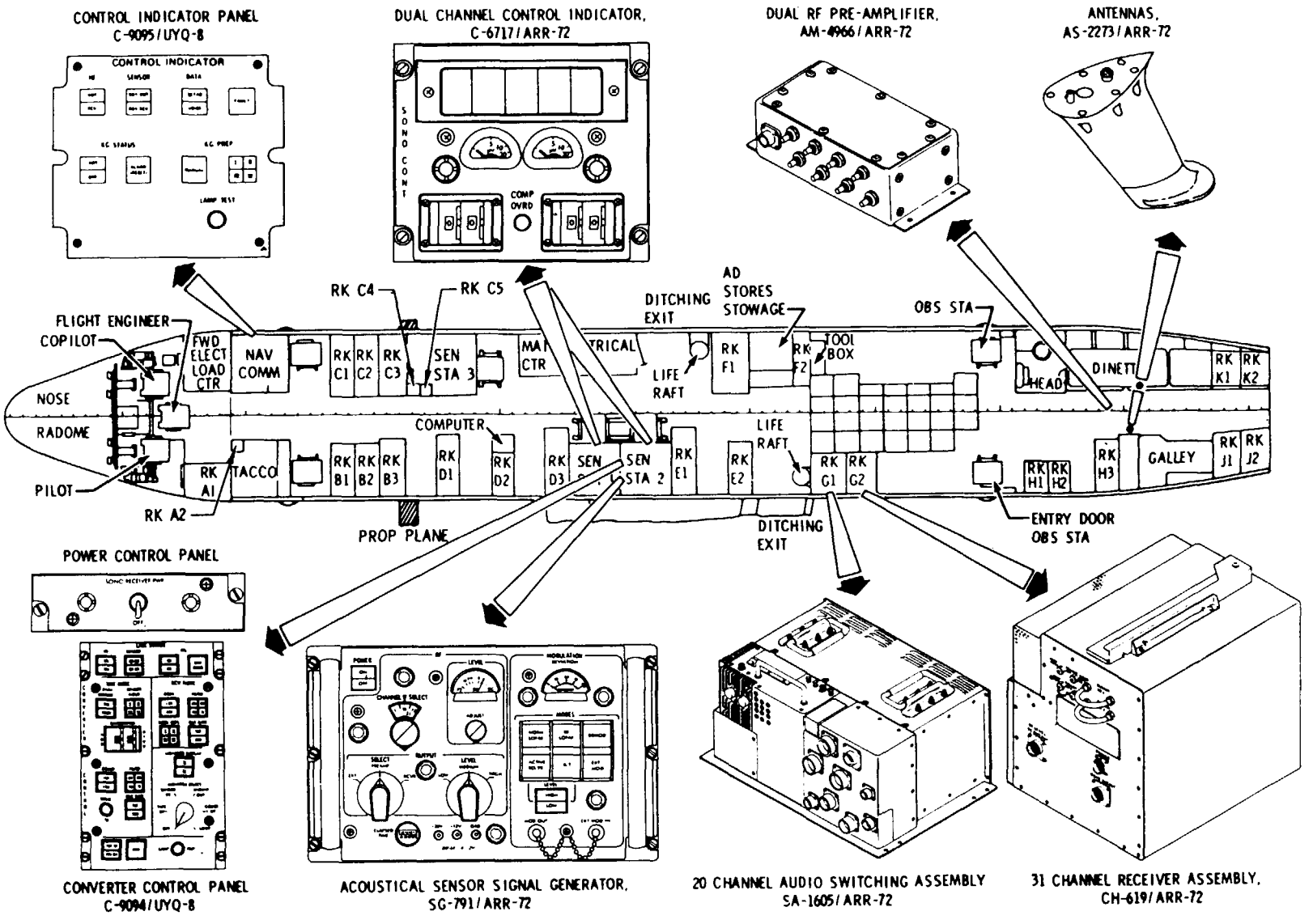


Figure 5-18.—Sonobuoy receiver group.

signals. Figure 5-19 is a block diagram of normal LOFAR signal generation.

Signal Flow

The normal LOFAR signal is a complex signal composed of broadband noise (which simulates background noise) 100 Hz and 225 Hz signals. The noise is generated through noise diode and transistor circuitry (A1Q1) and goes through amplifier stages A1Q2, Q3, Q4. The noise signal then goes through a 5 kHz filter, where it receives more amplification from A1Q5 and Q6 and couples to A1Q7 of the summing network A1Q7, Q8 and Q9. The 100-HZ oscillator (A1Q15 and Q16) generates the 100-HZ signal and applies it to amplifier A1Q17. A1R47 provides level adjustment and couples the signal to A1Q8 of the summing network. The 200-Hz oscillator (A1Q18 and A1Q19) signal receives amplification through A1Q20, is level adjusted through A1R58, and goes to A1Q8 of the summing network. The noise and 100-HZ and 200-Hz signals are summed and become the normal LOFAR signal outputs selected by depressing Z2A.

AN/ARR-72 Miniature Sonobuoy Receiver System

Normally the AIMD will receive a module from a receiver stick. The ATE shop will conduct troubleshooting and repair. The receivers use typical circuitry found in typical heterodyne receivers, as explained in the Navy Electricity and Electronics Training Series (NEETS) and in chapter 3 (Communications) of this manual. The modules are interchangeable with the exception of the RF oscillator. You can troubleshoot the electronics and circuits within the system with basic electronics circuitry knowledge, the MIM, and an appropriate ATE or test bench.

SIGNAL PROCESSING

A signal processor, spectrum analyzer (SA), or acoustic data processor (ADP) provide

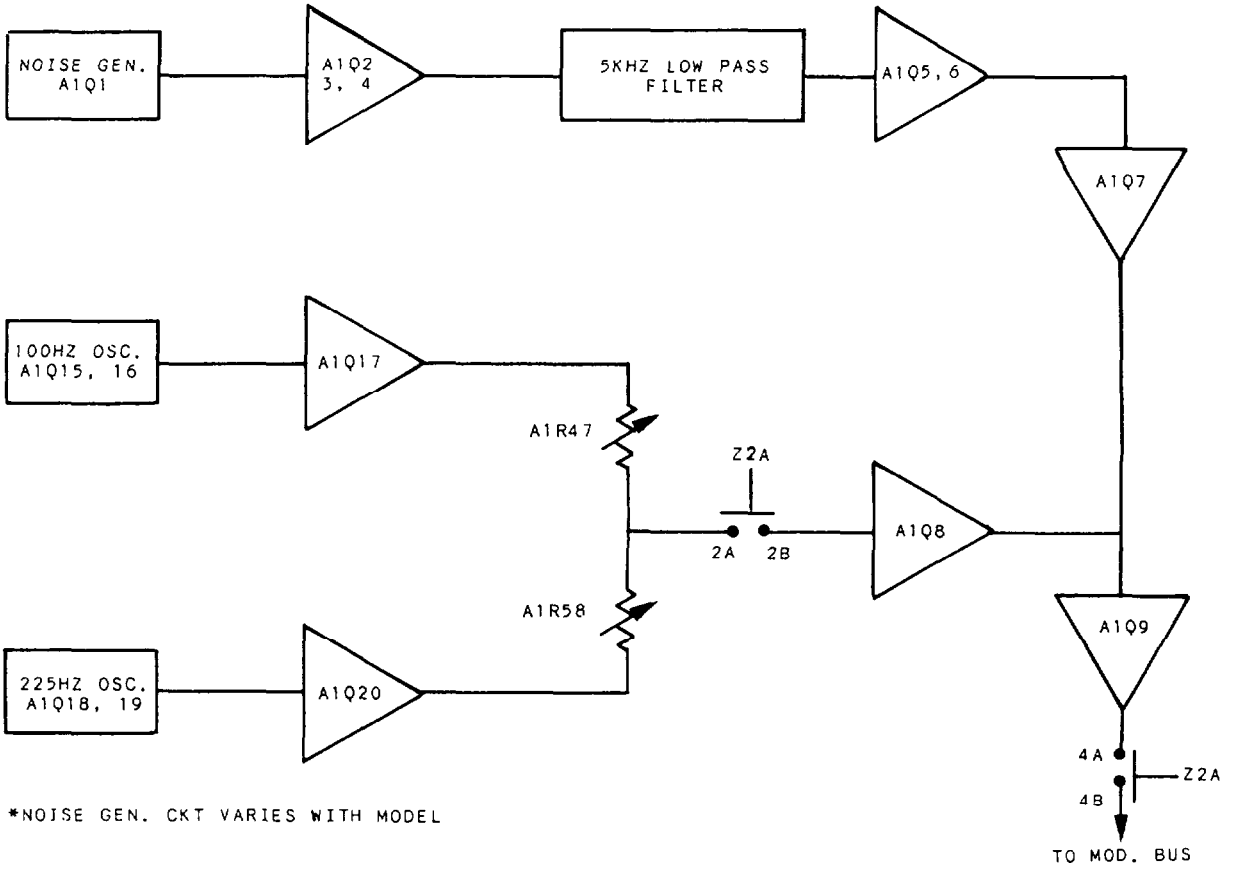


Figure 5-19.-Normal LOFAR signal generation block diagram.

basically the same functions when considered as part of an acoustic system. The processor aboard an SH-2F uses an RO-114 graphic recorder system, while the SH-60B and S-3B use the AN/UYS-1. The P-3C uses a version of the AN/AQA-7 recorder group or AN/UYS-1 systems. All of these systems have basically the same mission and perform some of the same functions.

The AN/UYS-1 spectrum analyzer (SA) (fig. 5-20) processes sonobuoy signals from CASS, RO, LOFAR, DIFAR, and DICASS sonobuoys

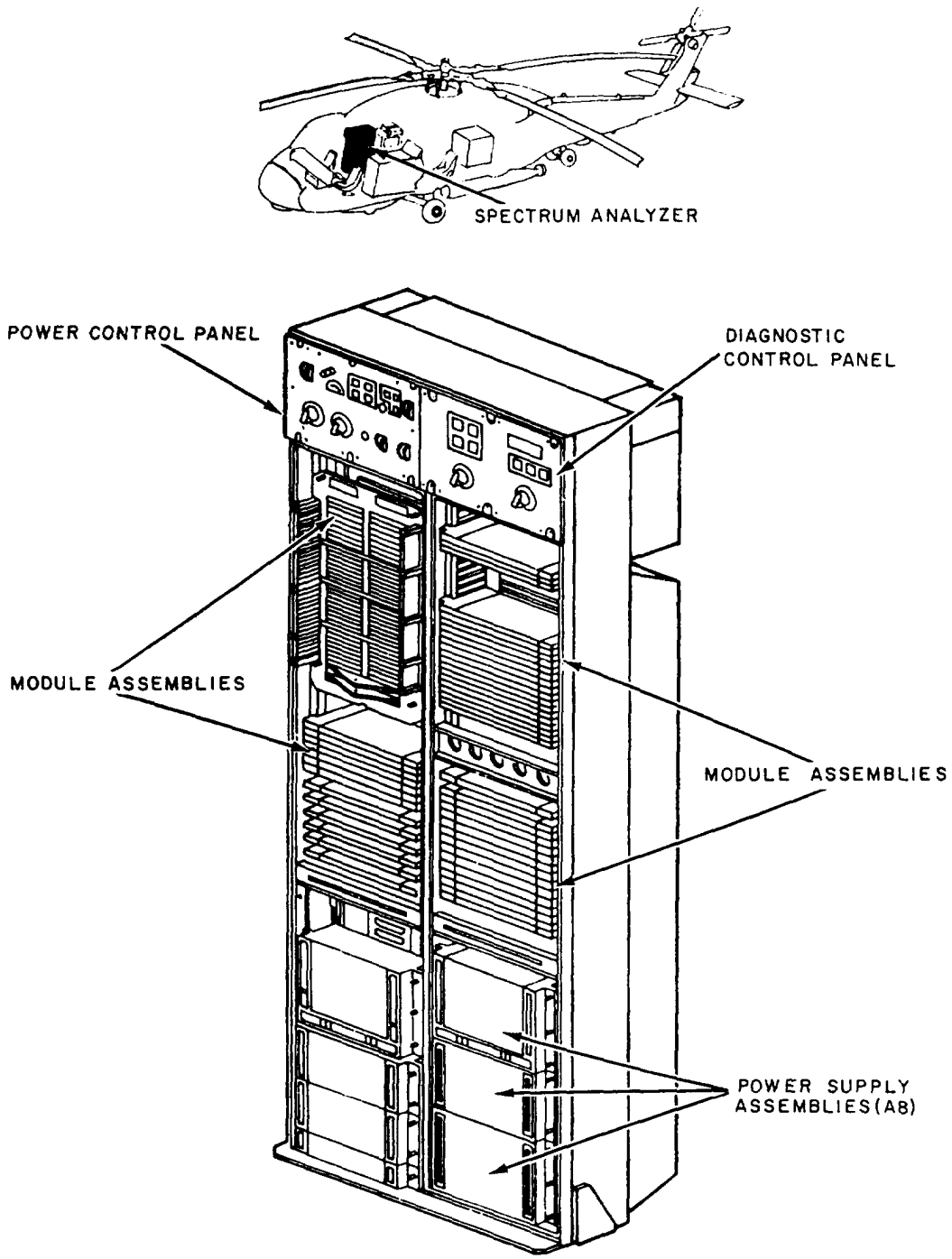


Figure 5-20.-AN/UYS-1spectrum analyzer and its location in SH-60B helicopter.

through the receiver. After processing, the signals are either displayed, recorded, or further analyzed, depending on the aircraft platform and acoustic suites. The SA consists of functional subunits, which contain all the major items needed for unit performance. Each functional subunit contains replaceable electronic modules. The functional subunits include:

1. Control processor
2. Input signal processor
3. Storage controller
4. Bulk store
5. Arithmetic processor
6. Power supply
7. Input/output (I/O)
8. Proteus digital channel
9. Diagnostics and power control panel

The SA is a high-speed signal processor especially designed for extracting target information from both active and passive sonar inputs. In most platforms the job assigned to the SA is determining source signal frequency, amplitude, bearing, Doppler, range, and similar parameters.

TAPE RECORDER (ATR)

One of the analog tape recorders used in ASW acoustic platform suites is the AN/AQH-4(V) sound recorder-reproducer set used in the P-3 community. This recorder provides a permanent record of audio data for analysis of contacts. Figure 5-21 shows the AN/A H-4(V) system interface diagram. The basic unit is a 14 track or channel recorder-reproducer with channel 1 for time code data and channel 2 for ICS recording. The AN/AQH-4(V) basic system contains two major subassemblies—the A1 tape transport assembly and the A2 card cage assembly. The Al contains five plug-in shop replaceable assemblies (SRA) that include a A1A1 power supply, A1A2 transport assembly, A1A3 capstan servo module, A1A4 counter module, and a A1A5 control module. The A2 assembly contains 14 plug-in modules that includes five A2A1 direct record or

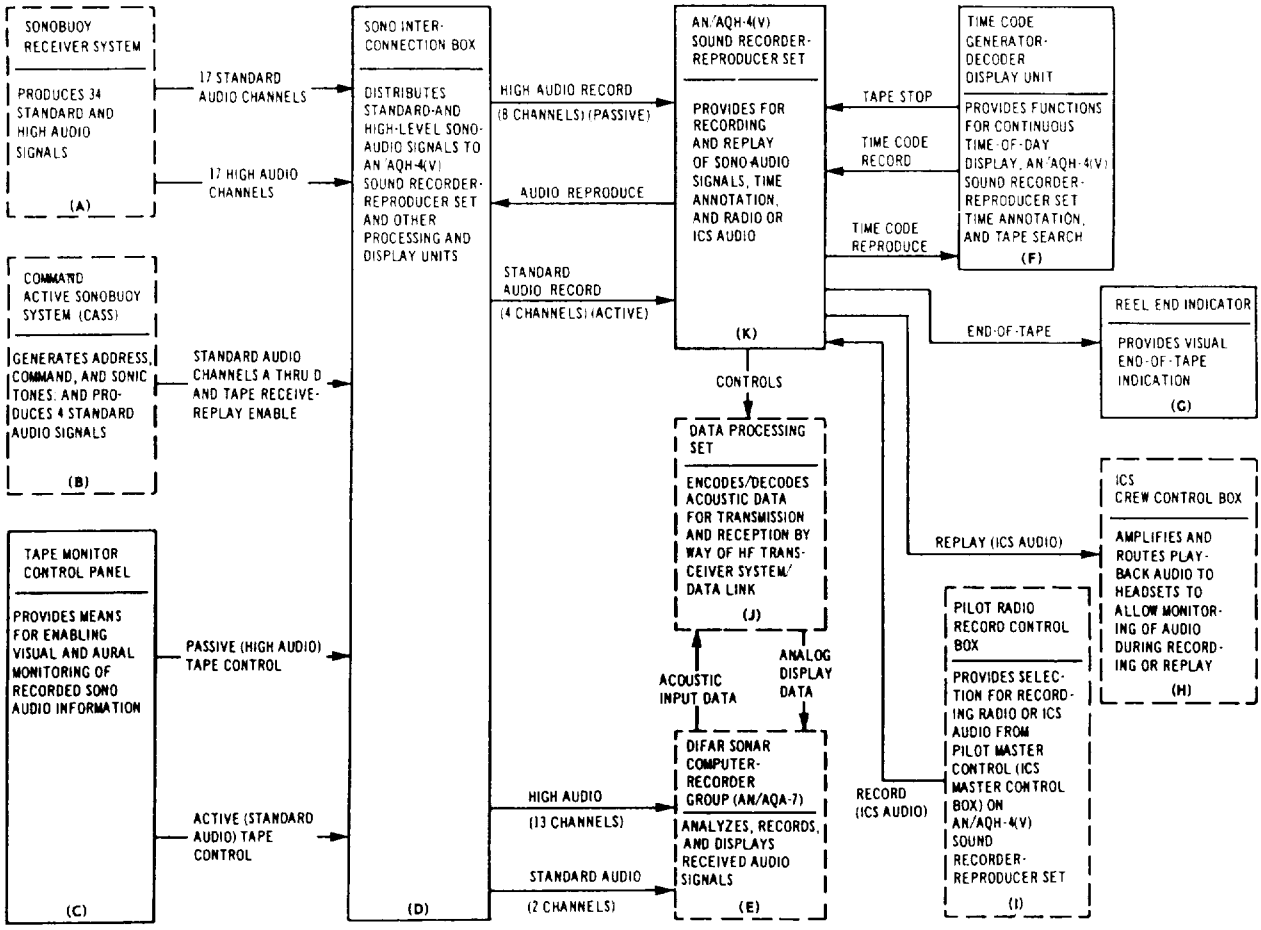
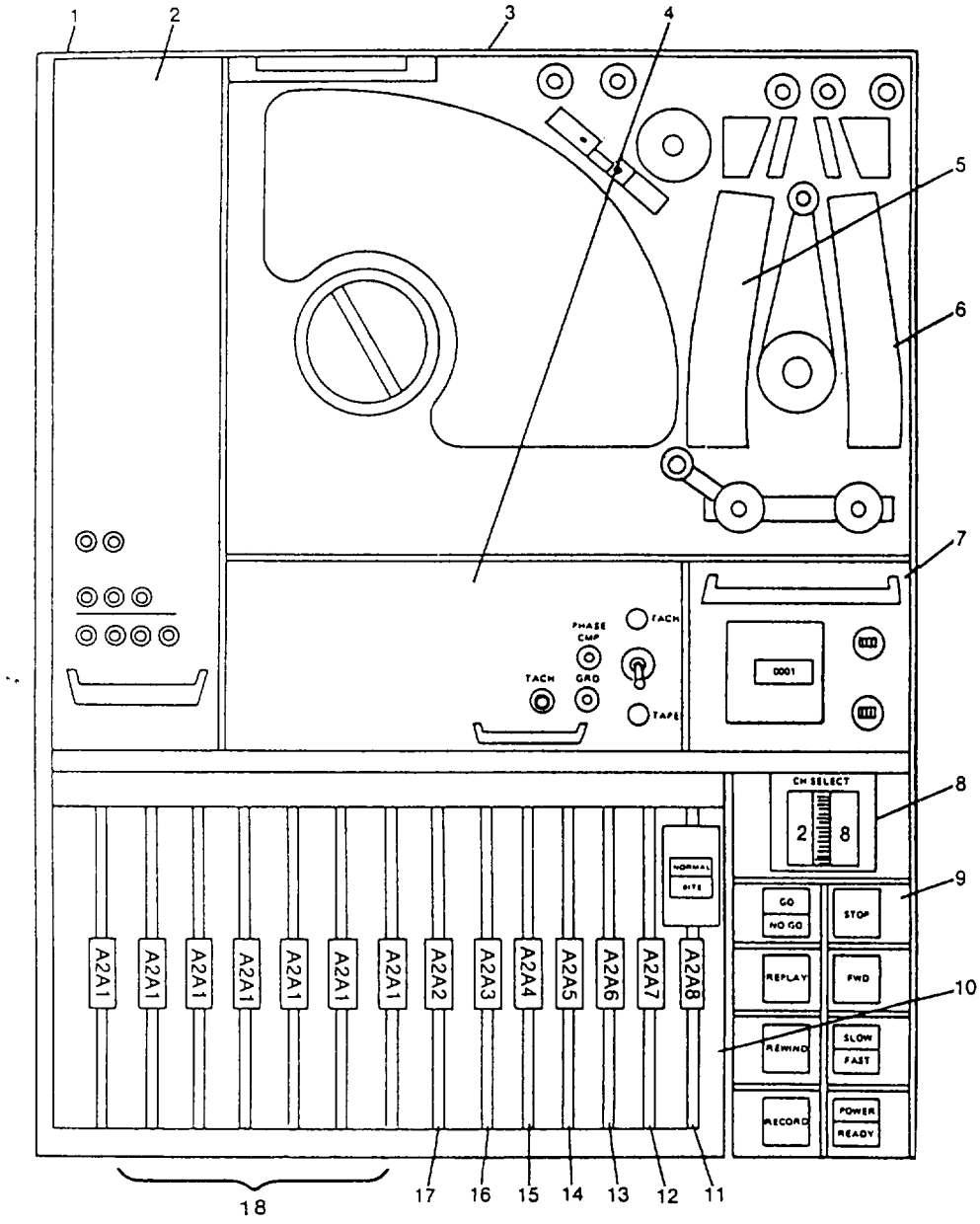


Figure 5-21.-Sound recorder-reproducer system interface diagram.

reproduce modules, eight A2A2 FM record or reproduce modules, and one A2A3 ICS record or reproduce module. Maintenance procedures and detailed signal flow and troubleshooting procedures are contained in NAVAIR 16-30AQH4-1.

The 28-track AN/AQH-4(V)2 recorder-reproducer set is the unit we will discuss.

The major components of the AN/AQH-4(V)2 (fig, 5-22) are the A1 and A2 subassemblies. The A1 sound recorder/reproducer assembly houses



1. Sound Recorder/Reproducer Assembly A1
2. Power Supply A1A1
3. Transport Assembly A1A2
4. Capstan Servo Module Assembly A1A3
5. Record Head Assembly A1PU1
6. Reproduce Head Assembly A1PU2
7. Counter Module Assembly A1A4
8. CH Select Switch A1S1
9. Control Module Assembly A1A5
10. Card Cage Assembly A2
11. Bite Switch PWB A2A8
12. Record Bite PWB A2A7
13. Reproduce Select PWB A2A6
14. FM Direct PWB A2A5
15. FM/Time Code PWB A2A4
16. 2-Channel FM Reproduce PWB A2A3
17. 7-Channel Record PWB A2A2
18. 3-Channel FM Record PWB A2A1

Figure 5-22.-AN/AQH-4(V) sound recorder-reproducer set components location.

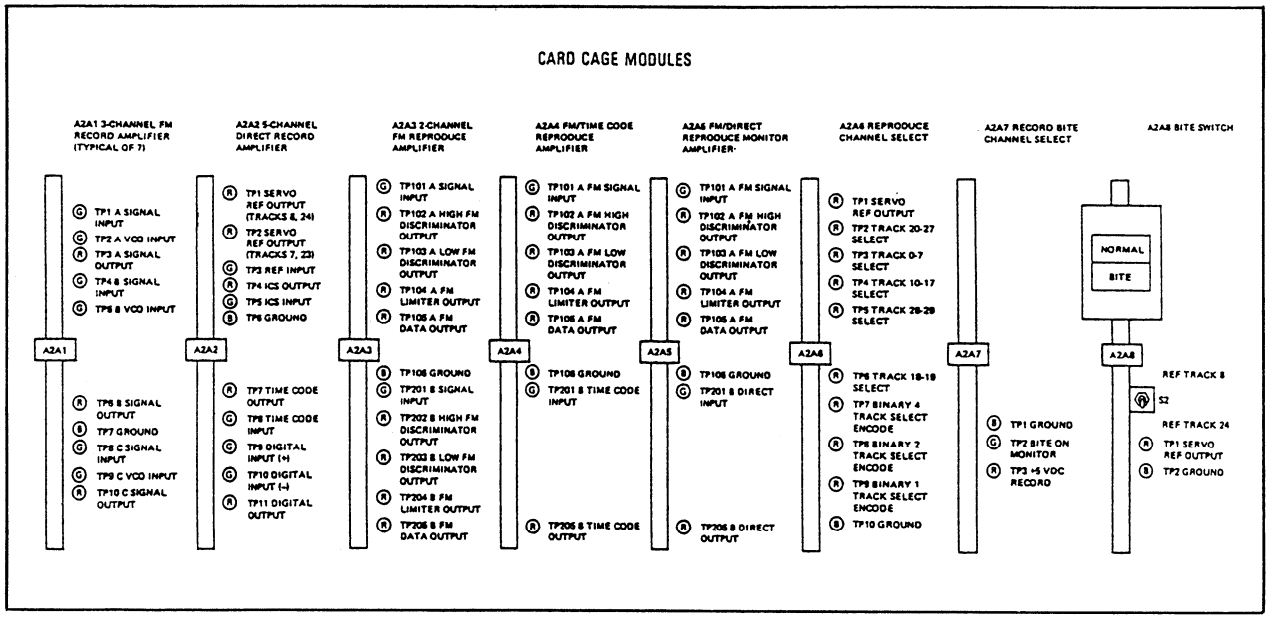


Figure 5-23.—AN/AQH-4(V)2 sound recorder-reproducer set test data.

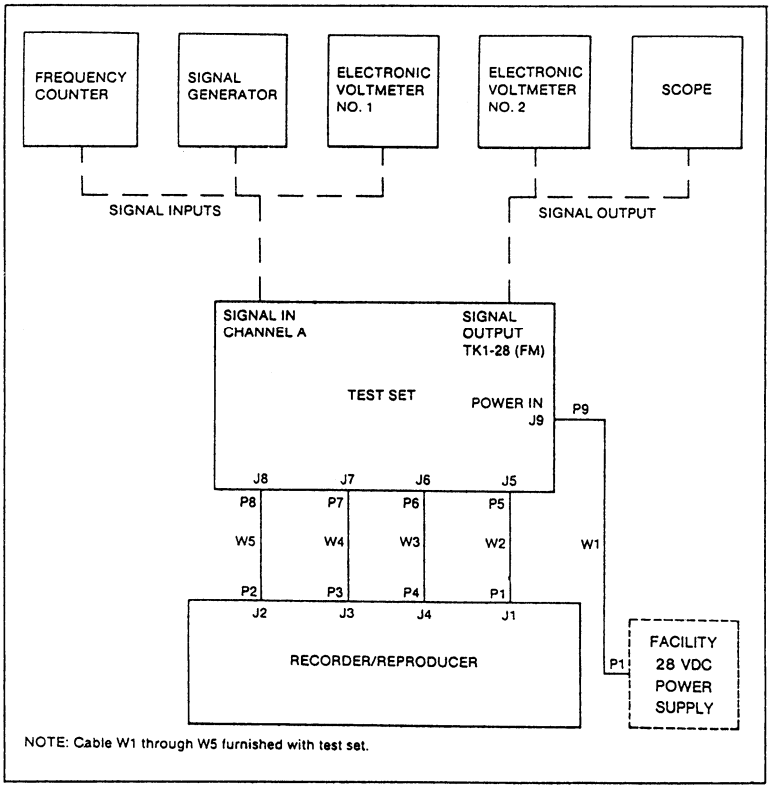


Figure 5-24.—Operational check bench setup.

the same modules as the AN/AQH-4(V); however, they are not completely interchangeable with one another. The MIM, NAVAIR 16-30AQH4-3, contains information on the modules and their interchangeable characteristics. The A2 card cage assembly still houses 14 circuit cards with 28 record/reproduce channel amplifiers. Twenty-one amplifiers are FM data channels and four are servo reference channels. The four remaining channels are ICS, time code, and digital channels. The A2 subassembly contains the A2A1, 3-channel FM record amplifier (7 each), the A2A2, 5-channel direct record amplifier, and the A2A3, 2-channel FM reproduce amplifier. The A2 unit also houses the A2A4 FM/time code reproduce amplifier, the A2A5 FM/direct reproduce monitor amplifier, the A2A6 reproduce channel select, the A2A7 record bite/channel select, and the A2A8 bite switch. Figure 5-23 shows the AN/AQH-4(V)2 A2 card cage modules and test point identification. Figure 5-24 is the operational check bench setup. Complete trouble-shooting and maintenance procedures are too lengthy and beyond the scope of this TRAMAN. As an example, we have provided a short functional explanation for the A2A1 3-channel FM record amplifier and the A2A2 5-channel direct record amplifier.

A2A1 3-Channel FM Record Amplifier

The A2A1 (fig. 5-25) is typical of seven identical circuit cards. Each card provides amplification for three channels of audio data for a total of 21 channels. The module accepts one to three channels of the 21 possible FM input signals through J3. The input signal' 'A" data at TP-1 is differentially coupled through U1 differential input amplifier circuitry to U4 switch. U4 selects between the data signal and bite signal, depending on the bite control signal, and routes the appropriate signal to the voltage-controlled

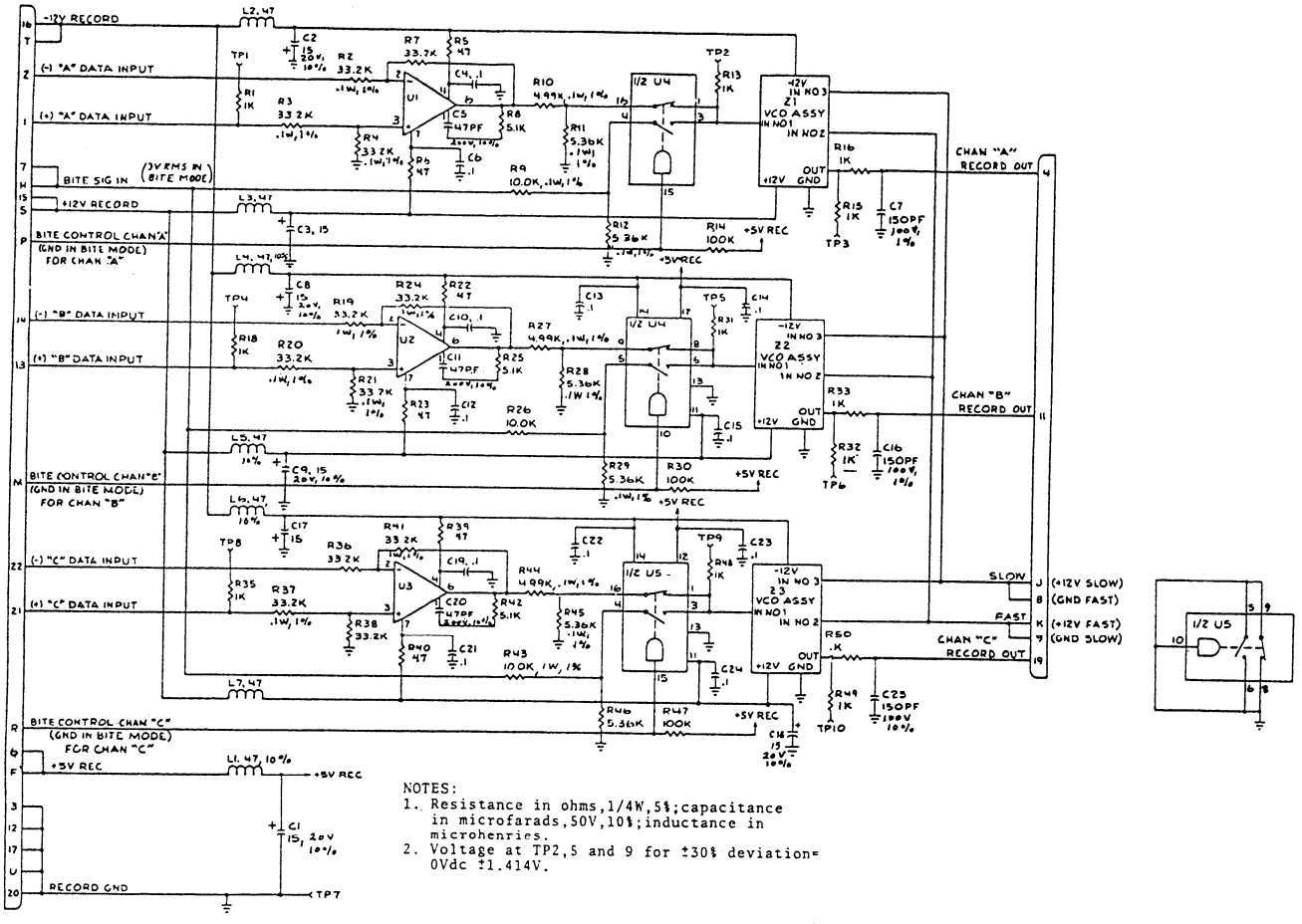


Figure 5-25.—3-Channel FM Record PWB Assembly A2A1 schematic diagram.

oscillator circuitry. The VCO provides carrier deviation, speed selection, and outputs to the record and bias adjust assembly.

A2A2 5-Channel Direct Record Amplifier

The A2A2 (fig. 5-26) interfaces ICS, time code, digital, and servo reference signals to the record heads. The input signals are from J3, BITE switch and servo reference signals from the A1A3, and outputs to the record and bias adjust assemblies. The ICS signal from J3 goes to pin 9 of U1, to operational amplifier U3, and output to the record and bias adjust assemblies A2A9 and A2A10. The A2A9 and A10 allow record current adjustment through variable resistances of the record amplifier outputs, and routes these signals to the A1A6A1 and A1A6A4 record mixer/bias adjust assemblies. The A1A6A1/A4 mixes the record signals with a 4 MHz bias signal from bias oscillator A1A2A1, and outputs them to record head A1PU1. The A1PU1 record head assembly contains two heads with 14 tracks each. One head is for odd tracks and one head for even tracks, providing a total of 28 tracks. The reproducer heads are identical to the record heads.

DIRECTIONAL LISTENING

The directional listening control (DLC) selects channel I, II, III, or IV for monitoring and provides either a cardioid pattern or figure-8 directional listening pattern. The operator uses the bearing dial to determine a targets bearing. The bearing dial controls a resolver and synchro transmitter that causes an audio null and sends a three-phase signal representing the bearing to the signal data converter (SDC).

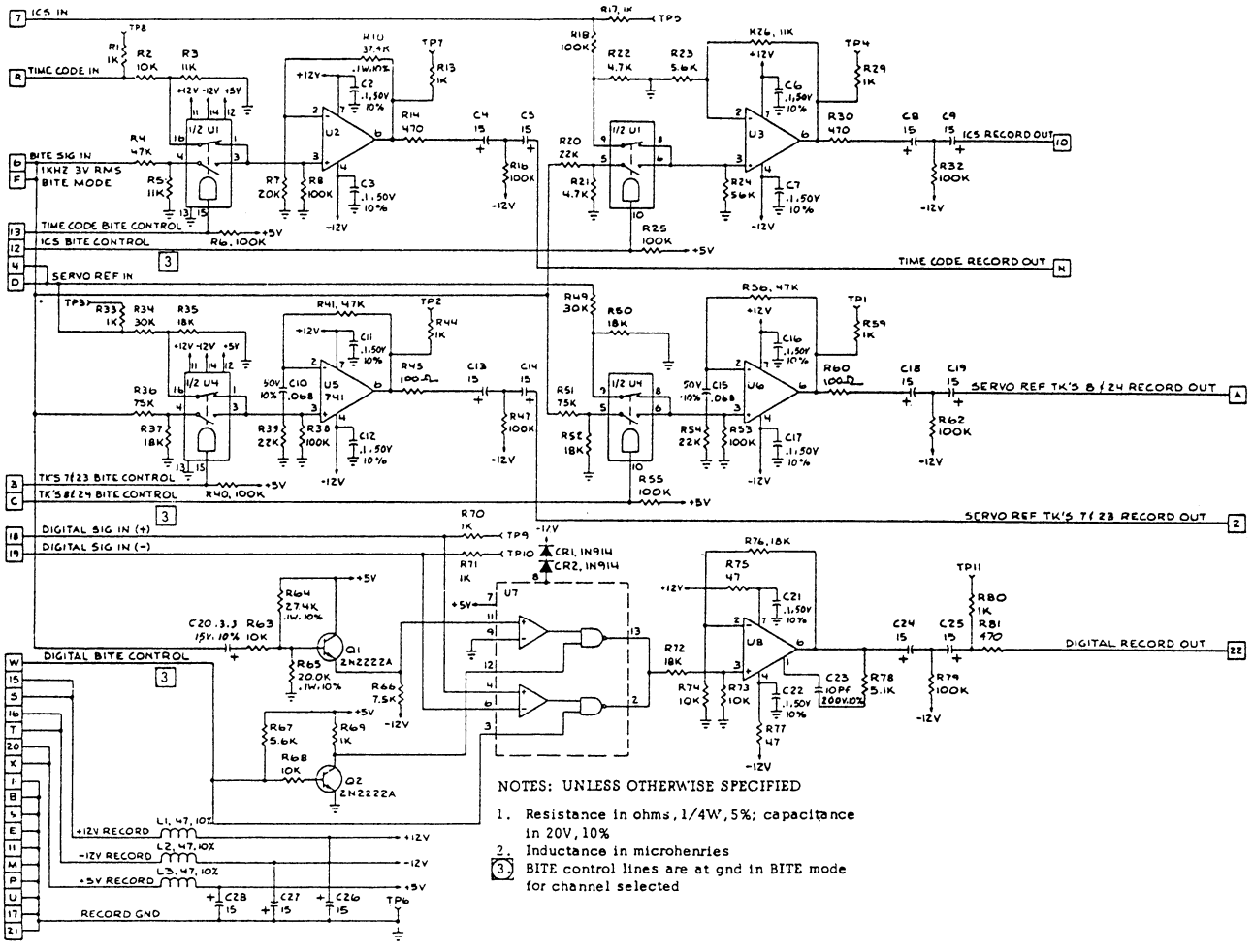


Figure 5-26.—5-Channel Direct Record PWB Assembly A2A2 schematic diagram.

SONAR PRINCIPLES

Learning Objective: *Identify factors that affect the behavior of a sound beam in water, and recognize Doppler, hydrophone, and transducer principles of operation.*

Airborne sonar equipment, commonly known as dipping sonar, is carried aboard SH-2F, SH-3, and SH-60B helicopters. Sonobuoys, which are sound-activated devices, are also dropped into the ocean from helicopters and from shipboard fixed-wing ASW aircraft, such as the S-3A *Viking,* and from patrol planes, such as the P-3C *Orion.* The floating sonobuoys, which are actually expendable sonar sets, are then monitored by radio. For photographs of and additional information on the ASW aircraft, refer to *Airman,* NAVEDTRA 10307 (series).

The operating principle of sonar is similar to that of radar, except that sound waves are used instead of radio-frequency waves. When sound wave energy strikes an object, the object reflects a portion of that energy back to its source. Since the speed of the sound wave and the time it takes to travel out and to return are known, range can be calculated. By knowing the direction of the reflected sound echo, the sonar equipment operator can obtain bearing information. In-depth information is available in the *AT2(O)* TRAMAN, NAVEDTRA 12330. Review of the *AT2(O)* manual will help you to understand sonar principles and troubleshooting procedures.

FACTORS AFFECTING THE SOUND BEAM

The sound waves of interest to the sonar operator are those that leave the sonar transducer in the form of a beam and go out into the water in search of a submarine. If the sound beam finds a target, it will return in the form of an echo. Any signal strength lost during the beam's travel through the water is known as transmission loss. Some of the factors determining transmission loss are discussed in the following paragraphs.

Absorption and Scattering

Some of the sound wave is absorbed in passing through water. The amount of sound lost this way depends on the sea state. Sound absorption is high when winds are great enough to produce whitecaps and in wakes and strong currents, such as riptides. Absorption is greater at high frequencies than at low frequencies.

Sound waves are weakened when they reach a region of seawater that contains foreign matter, such as seaweed, silt, animal life, or air bubbles. This foreign matter scatters the beam and causes loss of sound energy. The practical result of scattering is reduced echo strength, especially at long range.

Reflection

Echoes (reflection) occur when sound strikes an object or a boundary between transmission mediums in such a manner as to reflect the sound or to throw it back to its origin. Reflection occurs where two mediums are of sufficiently different densities and the sound wave strikes at a sufficiently large angle. Reflection occurs because mediums of different densities transmit sound at different speeds. The speed of sound through seawater is about four times faster than of sound through air. Water is 800 times denser than air. Therefore, when the sound beam strikes the boundary between the sea and the air, practically all of the beam will be reflected downward from the sea surface.

Similarly, when a sound wave traveling through seawater strikes a solid object such as a submarine, the difference in the densities and the sound velocity in the two mediums is such that all but a small part of the beam is reflected. That portion of the beam that strikes surfaces of the submarine perpendicular to the beam will be reflected directly as an echo. (See figure 5-27.)

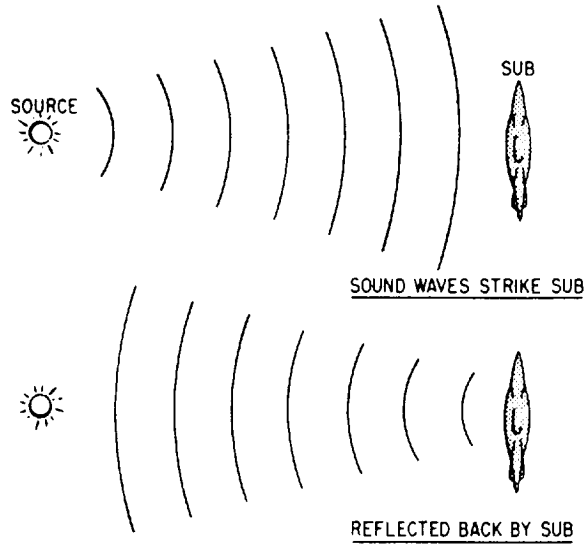


Figure 5-27.-Reflection of sound by a submarine.

In water less than 600 feet deep, the sound may also be reflected from the bottom. Other factors being equal, the transmission loss will be least over a smooth, sandy bottom and greatest over soft mud. Over rough and rocky bottoms, the sound is scattered, resulting in strong bottom reverberations.

Reverberation

When sound echoes and re-echoes in a large hall, the sound is reverberating. Reverberations are multiple reflections. Sound waves often strike small objects in the sea, such as fish or air bubbles. These small objects cause the waves to scatter. Each object produces a small echo, which may return to the transducer. The reflections of sound waves from the sea surface and the sea bottom also create echoes. The combined echoes from all these disturbances are called reverberations. Since they are reflected from various ranges, they seem to be a continuous sound. Reverberations from nearby points may be so loud that they interfere with the returning echo from a target.

There are three main types of reverberation or backward scattering of the sound wave.

1. There is reverberation from the mass of water. Causes of this type of reverberation are not completely known, although fish and other objects contribute to it.
2. There is reverberation from the surface. This is most intense immediately after the sonar transmission; it then decreases rapidly. The intensity of the reverberation increases markedly with increased roughness of the sea surface.
3. There is reverberation from the bottom. In shallow water this type of reverberation is the most intense of the three, especially over rocky and rough bottoms.

Divergence

Just as the beam from a searchlight spreads out and becomes weaker with distance, so does sound. The farther the target is from the sonar transducer, the weaker the sound waves will be when they reach it. This is known as spreading or divergence.

Refraction

If there were no temperature difference in the sea, the sound beam would travel in a straight line, since the speed of sound would be roughly the same at all depths. However, the speed of sound is not the same at all depths. The speed of sound in seawater increases from 4,700 feet per second to 5,300 feet per second as the temperature increases from 30°F to 85°F. Because of the varying temperature differences in the sea, the sound beam does not travel in a straight line, but follows curved paths. This results in bending, splitting, and distortion of the sound beam.

When a sound beam is bent, it is refracted. A sound beam is refracted (bent) when it passes from between areas of different temperatures (fig. 5-28). A sound beam will bend away from an area of higher temperature (higher sound velocity) toward an area of lower temperature (lower sound velocity). This is similar to the refraction of light, as discussed in *Aviation Electronics Technician (AT3),* NAVEDTRA 12329.

Speed of the Sound Beam

As mentioned previously, sound travels much faster in seawater than in the atmosphere. Near sea level, sound travels through the atmosphere at approximately 1,080 feet per second. In seawater, sound travels at 4,700 to 5,300 feet per second.

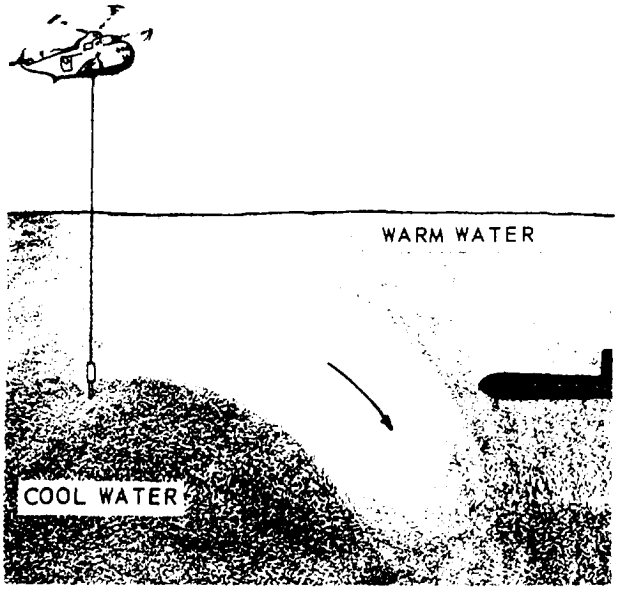


Figure 5-28.-Bending of sound beam between areas of different temperatures.

The three main characteristics of seawater that influence the speed of sound through the water are as follows:

1. Salinity (the amount of salt in the water)
2. Pressure (increases with depth)
3. Temperature (the effect of which is calculated in terms of slopes, or gradients)

The overall effect of increasing the salinity is to increase the speed of the sound. This means that when sound passes through water that varies in salinity, it travels faster in the saltier water. Such a change in salinity is considerable at the mouths of rivers emptying into the sea. In the open sea, the change in salinity is too small to significantly affect the rate of travel of the sound beam and may be ignored.

Since sound travels faster in water under pressure, the speed of sound in the sea increases proportionally with depth. This effect is also rather small.

Temperature is by far the most important of the factors affecting the speed of sound in water. The speed of sound will increase with increasing temperature at the rate of 4 to 8 feet per second per degree of change, depending on the temperature.

The temperature of the sea varies from freezing in the polar seas to more than 85°F in the tropics. Also, the temperature may decrease by more than 30°F from the surface to a depth of 450 feet. Thus, temperature is the most important factor because of its extreme variations. Remember, the speed of sound increases as the temperature of the water increases.

Depth and Temperature

Except at the mouths of great rivers where salinity may be a factor, the path followed by a sound beam will be determined by the pressure effects of depth and by temperature.

The pressure effect is always present and always acts in the same manner, tending to bend the sound beam upward. Figure 5-29 illustrates the situation when the temperature does not change with depth. Even though the temperature does not change, the speed of the sound increases with depth. The speed increase is due entirely to the effect of pressure. Notice in figure 5-29 that the sound beam bends upward.

Figure 5-30 shows what happens when temperature increases steadily with depth. When the surface of the sea is cooler than layers beneath

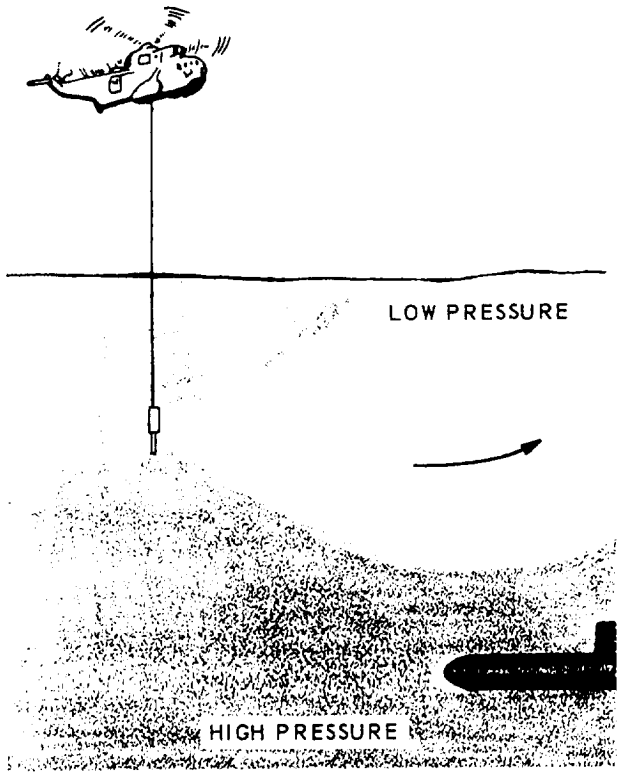


Figure 5-29.-Bending of sound beam away from high-pressure area.

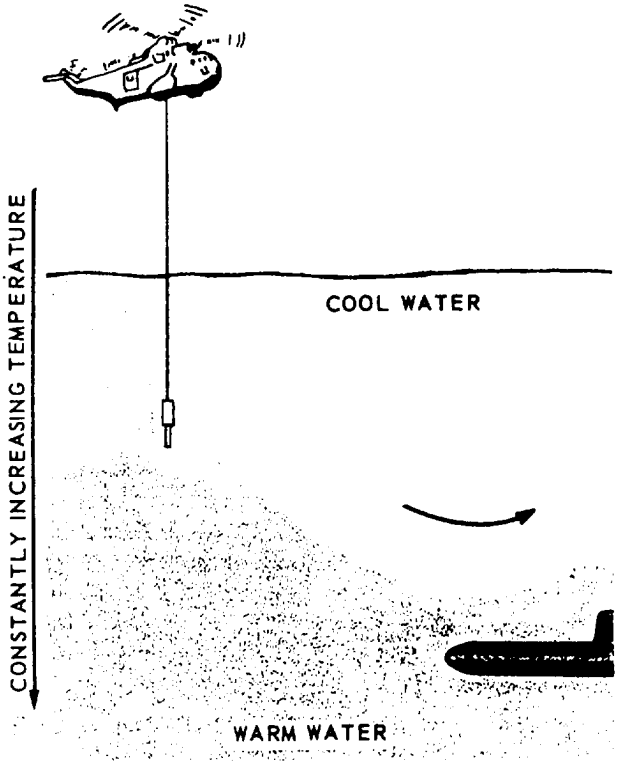


Figure 5-30.-Short ranges on a deep submarine.

it, the temperature increases with depth, and the water is said to have a positive thermal gradient. This is an unusual condition, but when it does happen, it causes the sound beam to be refracted sharply upward.

When the sea grows cooler as the depth increases, the water has a negative thermal gradient. Here the effect of temperature greatly outweighs the effect of depth, and the sound beam is refracted downward. This was shown in figure 5-28.

If the temperature is the same throughout the water, the temperature gradient is isothermal (constant temperature) (fig. 5-31). The surface layer of water in the figure is isothermal, but beneath this layer, the temperature decreases with depth. This causes the sound beam to split and bend upward in the isothermal layer and downward below it.

Remember, when no temperature difference exists, the sound beam bends upward due to pressure. When the temperature changes with depth, the sound beam bends away from the warmer water,

Under ordinary conditions, the sea has a temperature structure similar to that shown in figure 5-32. This temperature structure consists of three parts:

1. A surface layer of varying thickness with uniform temperature (isothermal) or a relatively slight temperature gradient
2. The thermocline, a region of relatively rapid decrease in temperature
3. The rest of the ocean, with slowly decreasing temperature down to the bottom

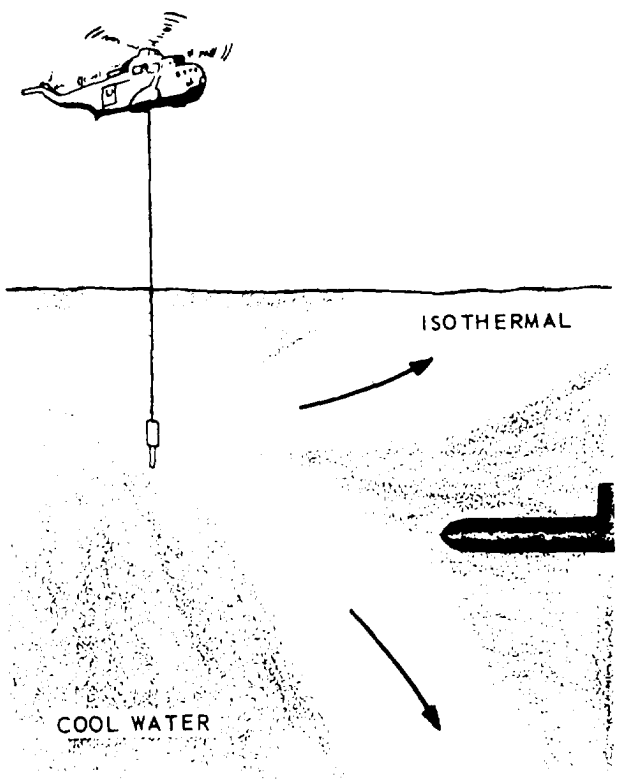


Figure 5-31.-Isothermal conditions.

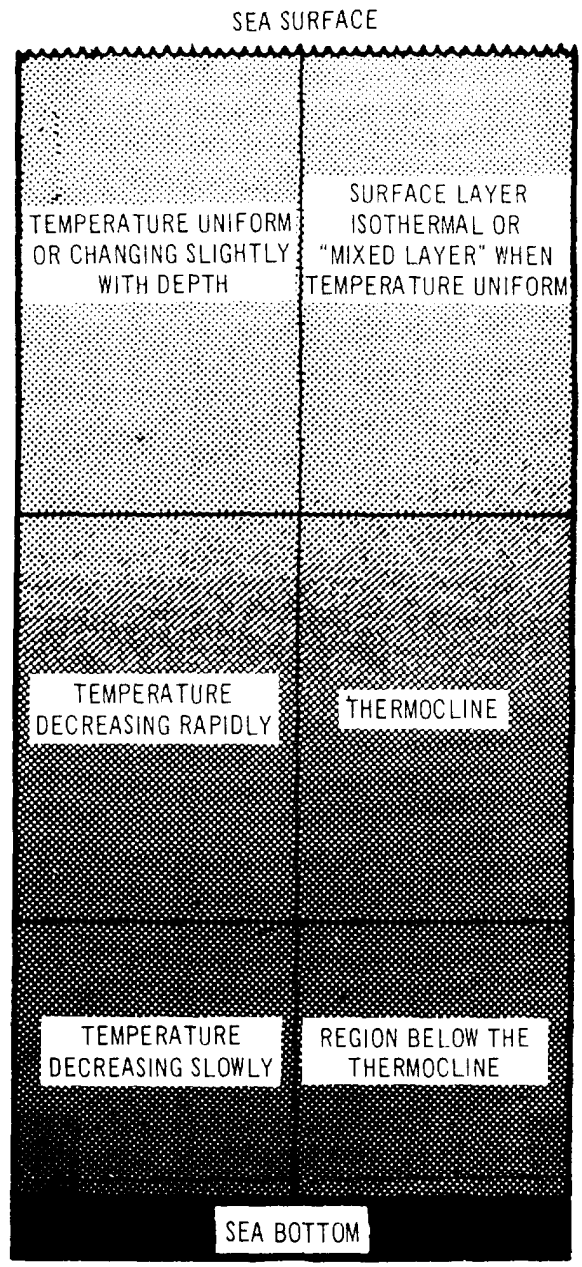


Figure 5-32.-Temperature structure in normal sea conditions.

If this arrangement changes, the path of the sound beam through the water will change.

Layer depth is the depth from the surface to the top of a sharp negative gradient. Under positive gradient conditions, the layer depth is the depth of maximum temperature. Above layer depth, the temperature may be uniform or a weak positive or negative gradient may be present.

Layer effect is the partial protection from echo ranging and listening detection, which a submarine gains when it submerges below layer depth.

Reports from surface vessels indicate that effective ranges on submarines are greatly reduced when the submarine dives below a sharp thermocline, and that the echoes received are often weak and sound "mushy."

RELATION OF DOPPLER EFFECT TO SONAR

Sonar equipment deals with three sounds. One is the sound actually sent into the water. The second is the sound of reverberations that return from all the particles in the water—seaweed, fish, and so forth. The third is the most important sound of all—the sound of the echo from the submarine.

The sound sent into the water (the actual PING) is seldom heard. All but a few sonar equipments are designed so that this signal is not picked up by the loudspeaker or headset. This means there are actually only two sounds to deal with in the discussion of Doppler effect in sonar.

Reverberations are echoes from all the small particles in the water. Consider just one of these particles for a moment. A sound wave from the transducer hits the particle and bounces back just as a ball would bounce back if thrown against a wall. If the particle is stationary, it will not change the pitch of the sound. The sound will return from the particle with the same pitch that it had when it went out.

If the sonar transducer is stationary in the water and sends out a ping of 10 kilohertz, the particles all send back a sound having the same pitch. Now suppose that the transducer acquires forward motion and a ping is sent out dead ahead. It is just as if the transducer were the oncoming train and the particles were occupants of the car, listening to the approaching sound. Remember that as the train came forward, the pitch of the whistle sounded higher to the occupants of the car, In the same way, the particles "'hear" a higher note and reflect this higher note. Therefore, the sonar equipment will detect a higher note than the one sent out. If the transducer in this example is pointed dead astern, a lower note than the one sent out will be heard.

If the transducer is aimed perpendicular to the direction of motion, the particles in the water will echo the same note sent out because the transducer is neither going toward the particles nor away from them (fig. 5-33).

Now consider the echo from the submarine, as illustrated in figure 5-34. Again, the transducer is shown stationary. When the submarine is neither going toward nor away from the stationary

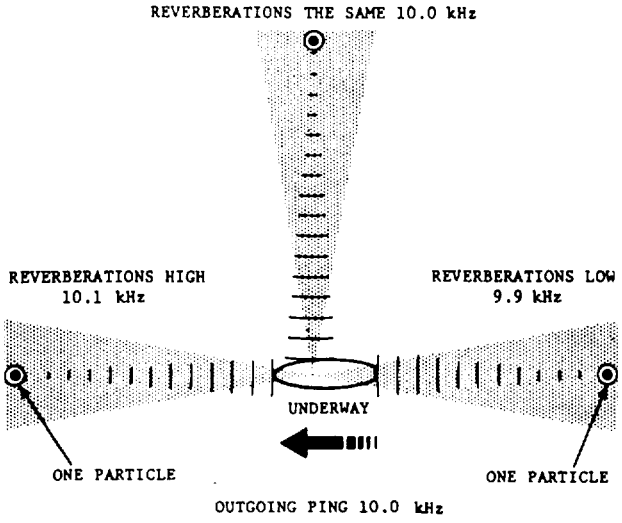


Figure 5-33.-Transducer installed on a moving ship. Pitch is high ahead, low astern.

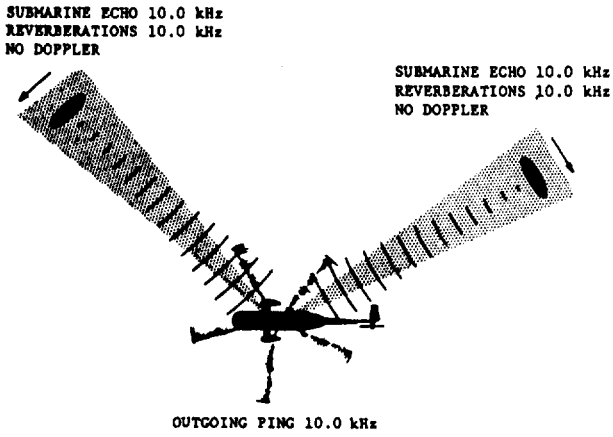


Figure 5-34.-Transducer supported by a helicopter. Doppler effect is absent when submarine is stationary or moves at right angles to sound beam.

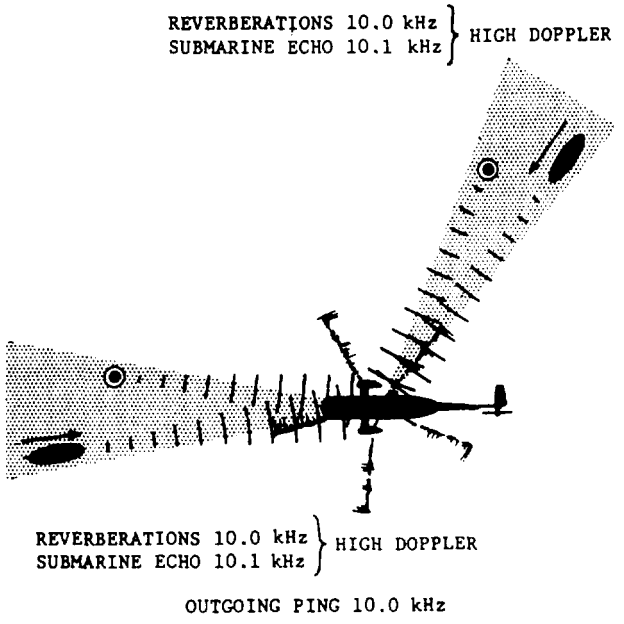


Figure 5-35.-Comparison of echo frequency and reverberation frequency when submarine moves toward transducer.

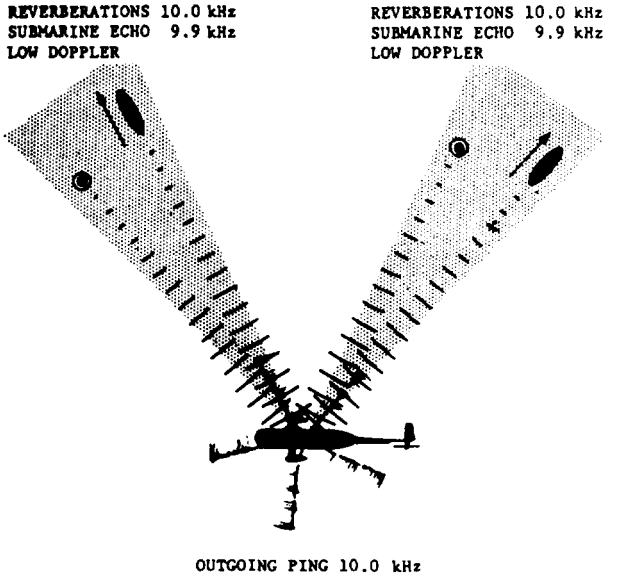


Figure 5-36.-Comparison of echo frequency and reverberation frequency when submarine moves away from transducer.

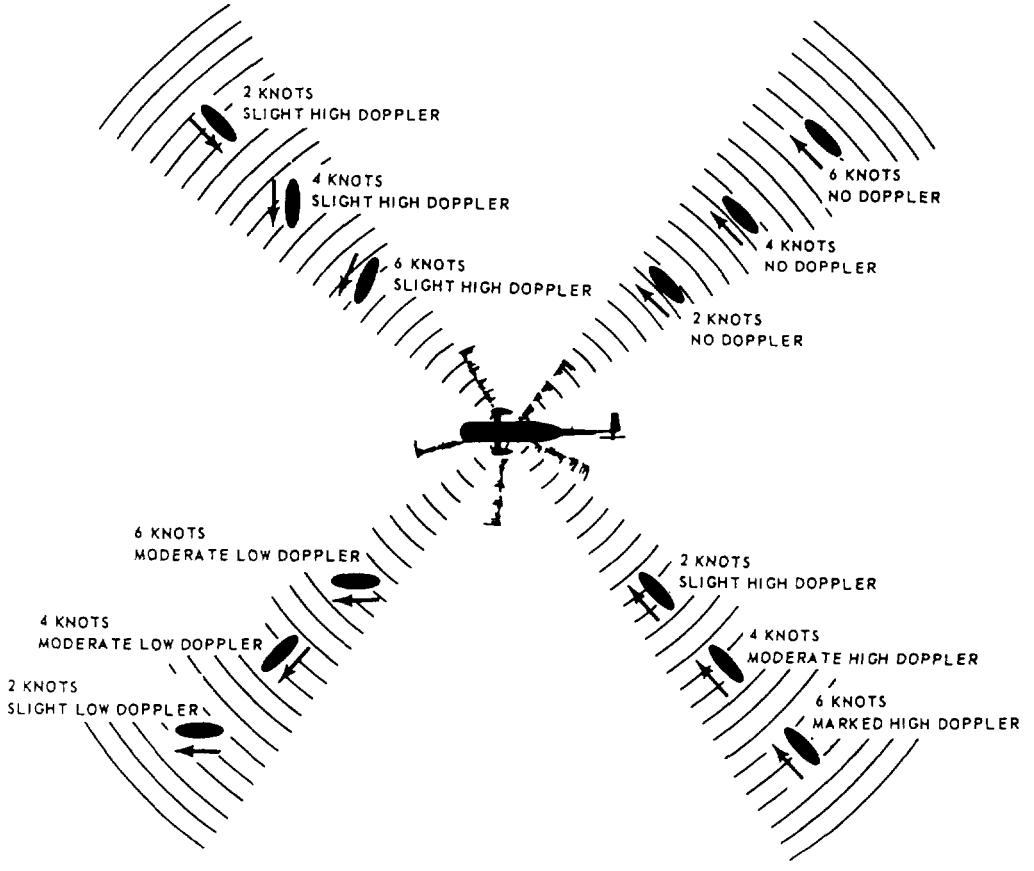


Figure 5-37.-Varying degrees of Doppler effect due to differences in course and speed of submarines.

transducer, it must either be stopped or crossing the sound beam at a right angle. If it is in either condition, it reflects the same sound as the particles in the water. Consequently, the submarine echo has exactly the same pitch as the reverberations from the particles.

Now suppose that the submarine is going toward the sound source, as illustrated in figure 5-35. It is as though the submarine is the train heading toward the car that is blowing its horn at the crossing. The horn sounds higher as the train approaches the car. In the same manner, the sound beam sounds higher to the submarine as it approaches the sound source.

The submarine reflects an echo of higher pitch than caused by the particles in the water, which are not moving. When the echo from the oncoming submarine is higher in frequency than the echoes from the reverberations, the Doppler is said to be high. The opposite form of Doppler shift will occur when the submarine is heading away from the transducer. In this case, the pitch of the echo from the submarine is lower than the pitch of the reverberations (fig. 5-36).

The degree of Doppler indicates how rapidly the submarine is moving relative to the transducer, For example, a submarine moving directly toward the transducer at 6 knots returns an echo of higher frequency than one moving at only 2 knots. Also, a submarine moving at 6 knots directly toward the transducer returns a higher echo than it would if it were heading only slightly toward the transducer. Refer to figure 5-37. This figure illustrates 12 submarines traveling at various speeds (knots) and courses (indicated by arrows) with respect to a stationary hydrophone supported by helicopter. Notice in the figure how the Doppler of each submarine is influenced by its speed and direction.

Doppler also makes it possible to distinguish the difference between a wake echo and a submarine echo. Relatively speaking, the submarine's wake is stationary. Therefore, its wake returns an echo whose frequency is different from that of the Doppler shifted submarine echo.

TRANSDUCER AND HYDROPHONE PRINCIPLES

A transducer functions as an underwater loudspeaker during sound transmission and as an underwater microphone during sound reception. The hydrophone, however, is used only to receive sound.

Transducers used in sonar operate on either the magnetostriction principle, the piezoelectric principle, or the electrostrictive principle. Figure 5-38 shows the construction of two transducers. The one shown in view A is a piezoelectric transducer and has many crystals mounted on the diaphragm. View B shows the magnetostriction transducer, which has many nickel laminations. The nickel laminations of the magnetostriction transducer and the crystals of the piezoelectric transducer are placed so close together that they cause the backing plate to act as one large vibrating surface. This arrangement

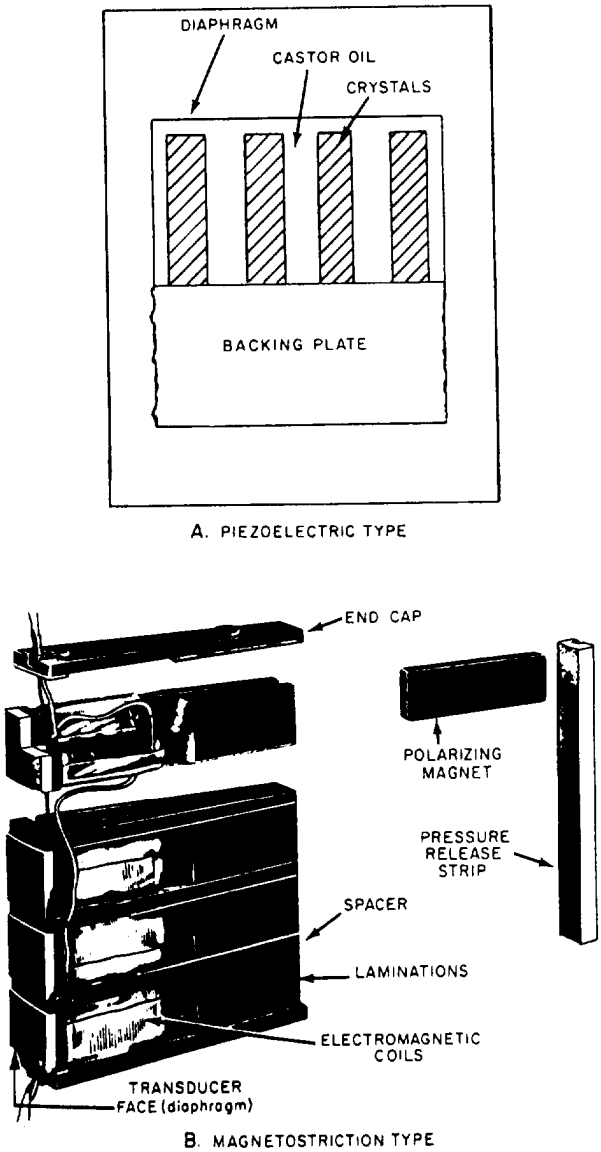


Figure 5-38.-Construction details of two types of sonar transducers.

is a contributing factor in producing the sound energy needed for sonar operation.

The magnetostriction transducer can be filled with a moisture-free gas to ensure long trouble-free operation. The piezoelectric transducer uses a watertight case that has a rubber dome for the diaphragm. The case is filled with a special moisture-free oil. This oil, which has nearly the same sound transmission characteristics as seawater, acts as a medium between the crystals' vibrations and the diaphragm. Both types of transducers must be made watertight to prevent corrosive action.

Like the transducer, the hydrophone operates on either the magnetostriction principle or the piezoelectric principle.

Magnetostriction Transducer

The property that causes certain metals to shrink or contract when placed in a magnetic field is called "magneto striation." The stronger the magnetic field, the greater the contraction. Magnetostriction is most pronounced in nickel and nickel alloys. For this reason, nickel tubes are used exclusively in magnetostriction transducers.

As mentioned previously, a transducer is similar to a radio loudspeaker. A loudspeaker has a diaphragm that is caused to vibrate. These vibrations produce sound waves. Since the diaphragm of a transducer is operated under the surface of the water, it must be much heavier in construction than a diaphragm operating in air. Therefore, it needs a large driving power.

One type of magnetostriction transducer contains nickel tubes that are rigidly attached to a heavy diaphragm. These tubes are placed in an alternating magnetic field. This alternating field causes the nickel tubes to contract and expand, which causes the diaphragm to vibrate and produce sound waves in the water. The frequency of the sound waves depends upon the frequency of the alternating magnetic field around the nickel tubes.

Another type of magnetostriction transducer does not employ nickel tubes. Instead, the elements of the transducer have nickel laminations pressed into a thermoplastic material. Permanent magnets are so mounted that they provide a magnetic field for polarizing the nickel. The transducer illustrated in view B of figure 5-38 is of this type, and it is normally used with scanning sonar equipment.

The directivity of the transducer determines the accuracy of the bearing information. The sound beam for a given frequency can be made narrower in azimuth by increasing the width of the transducer diaphragm. If a higher operating frequency is used, the transducer can be made smaller and still have the desired directivity. However, there are limitations. If the frequency is increased too much, propagation losses caused by absorption become objectionable.

Piezoelectric Transducer

The piezoelectric transducer operates in the same manner as the magnetostriction type except that crystals are used instead of nickel tubes. If a mechanical stress is applied to certain crystalline substances, an electrostatic voltage is produced. Conversely, an electric field applied to a crystalline substance causes a mechanical stress (expansion or contraction of the crystal). This property of a crystal is called piezoelectric effect. There are various types of crystals, such as quartz, Rochelle salts, and tourmaline, that have this property. Quartz crystals are not used in sonar transducers because the frequency response of a quartz crystal is too narrow.

Crystals can be damaged by either moisture, shock, or high temperature. Since high power produces high temperatures, the crystal transducer has a disadvantage in that it cannot handle as much power as the magnetostriction transducer. One advantage of the crystal transducer is that it can operate over a much wider frequency range than can the magnetostriction transducer. Additionally, the crystal type is sensitive and very efficient.

Electrostrictive Transducer

When an electric field is applied across a dielectric, the dielectric is deformed. This phenomenon of change in dimensions is called "electrostriction," and is independent of the direction (sign) of the electric field and proportional to the square of the field intensity. In the crystal transducer, the electrostrictive effect is present, but it is much smaller than the piezoelectric effect and so is ignored. However, with barium titanate, a ceramic, the electrostrictive effect is large in comparison with the piezoelectric effect.

Ceramic for transducers has the advantage over crystals in that ceramic can be molded to any desired shape. This property is particularly desirable for making cylindrical scanning or omnidirectional transducers. The converse of electrostriction, the change in electric potentials

when the material is stressed, takes place only when a constant polarization potential is present. The ac sonar signal is thus superimposed on the larger dc polarization, and the material dimension will vary directly with the magnitude of the resultant potential. The polarization of electrostrictive materials is thus analogous to the magnetic polarization required for magnetostrictive materials. Without the polarization, mechanical stresses will not produce an electric potential.

Nearly all transducers now being built are of the ceramic type. Ceramic compounds have high sensitivity, high stability with changing temperature and pressure, and relatively low costs.

Hydrophone

The hydrophone is a microphone especially constructed to pick up sounds while submerged. It is used only as a receiving device. The transducer is used for both transmitting and receiving sound waves. (The transducer acts as a hydrophone when it is receiving an underwater signal.) The hydrophone can be designed so that its receiving characteristics are highly directional. This enables the hydrophone to provide accurate target bearing information. The hydrophone is also designed to operate over a wide frequency range.

TYPICAL AIRBORNE SONAR SYSTEM

Learning Objective: *Identify the major components of a typical airborne sonar system, and recognize operating features and modes of operation of the sonar system, including those of the sonar data computer.*

A modern airborne sonar system has many of the operating characteristics and features of its predecessors. Use of solid-state components and digital computer techniques, however, has improved the sensitivity, accuracy, and reliability of the typical airborne sonar system. The operating features, major components, and modes of operation of a typical airborne sonar set will be presented in the following paragraphs.

OPERATING FEATURES

The airborne sonar detecting-ranging set is a lightweight, echo-ranging, dipping-sonar set capable of detecting and classifying moving and stationary underwater objects within 20,000 yards. The sonar detecting-ranging set (sonar set) will provide information concerning exact range, bearing, speed, and configuration of targets. Automatic compensation for changes in water temperature and transducer rotation is provided in the set.

The sonar set has four primary modes of operation: active, passive, communicate, and aspect. In addition, it has a recording bathythermograph (BT) mode. The BT mode provides the operator with information useful in determining the optimum search depth. The sonar set also has built-in test equipment (BITE) test modes that provide confidence checks of the equipment.

Six selectable ranges of 1, 3, 5, 8, 12, and 20 kiloyards, three selectable frequencies of 9,230, 10,000, and 10,770 hertz, and two pulse lengths of 3.5 and 35 milliseconds are provided for active search operation. The 3.5-millisecond pulse is used for close ranges, better target resolutions, and in high reverberation areas.

When operating in the SHORT or LONG range, the sonar set will transmit a 3.5- or 35-millisecond pulse into the water. The transmission frequency is selected by the FREQUENCY switch. The sonar set transmits low power (250 watts) in the 1-kiloyard range and high power (2500 watts) in all other ranges.

Returning target echoes are visually displayed on a circular cathode-ray tube (CRT) that presents relative range and bearing of the target from the helicopter position referenced to magnetic north. A moveable cursor circle is provided to the sonar operator for selecting the target echo to be investigated. Audible presentation of targets, depending on their position on the CRT, is provided in the left or right earphone by a different nonharmonic tone for each sector of left or right bearing.

In the PASSIVE mode of operation, underwater noise sources are displayed on the CRT as noise spokes. The passive mode is used to search for possible targets without alerting the target.

Upon reception of a target echo, denoted by an audible tone in one of the earphones and a brightened spot on the CRT, the sonar operator positions the cursor circle over the selected target. This is done by manipulating the CURSOR POSITION bearing control and range switch. When the cursor circle is centered over the target spot, precise bearing and range information is presented on the BEARING digital display in degrees and on the RANGE-YARDS digital

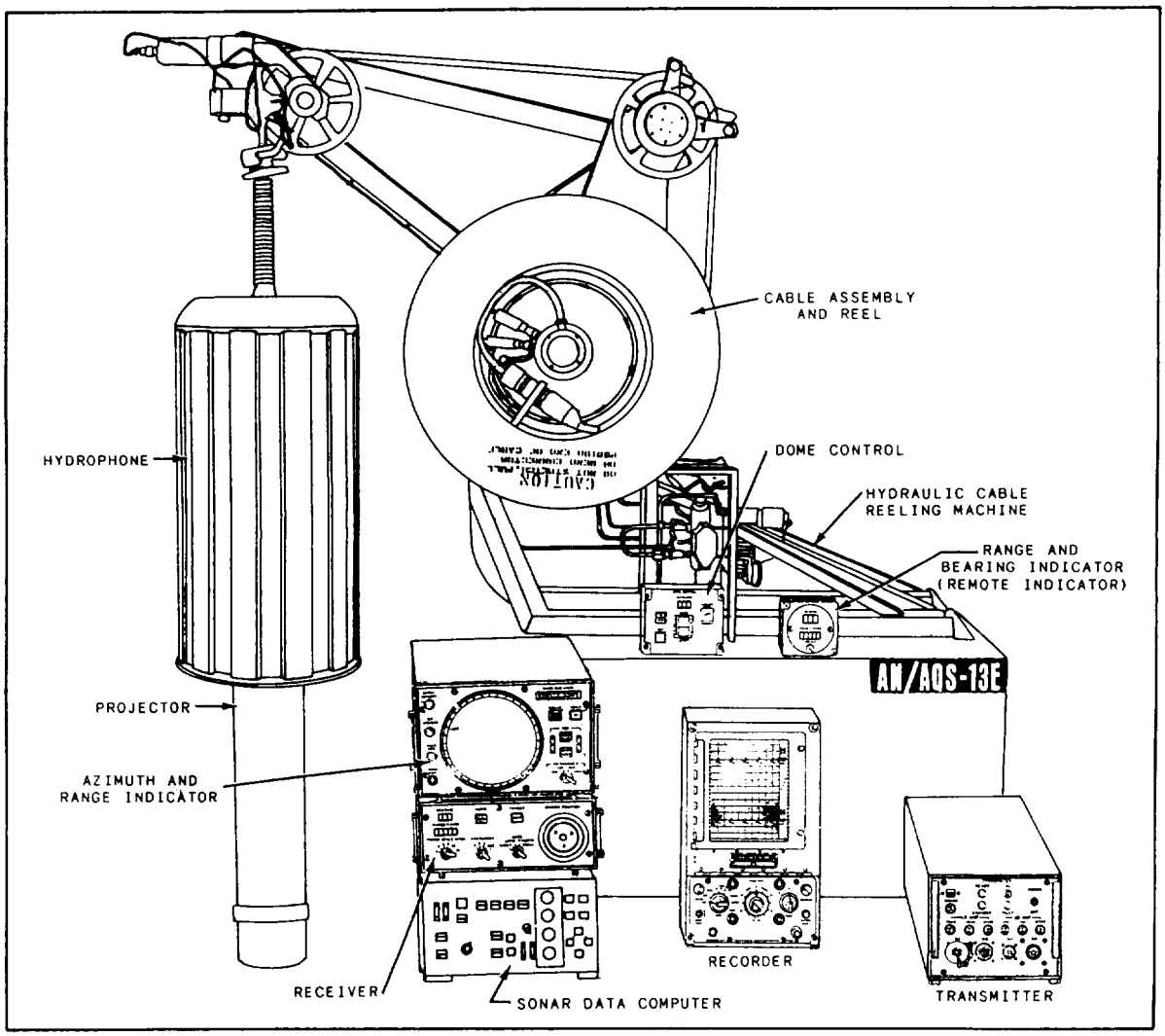
display in yards. The identical bearing and range information is available on the pilot's bearing and range indicator.

In the COMM (communication) mode of operation, underwater voice communications can be conducted between similarly equipped helicopters, submarines, or surface vessels. Voice communications is established by the operator by pressing a foot switch to transmit and releasing it to receive.

In the recorder bathythermograph (BT) modes, the sonar set is operated in the PASSIVE mode to provide temperature gradient recordings in one of two BT modes (25°F to 75°F or 45°F to 95°F). Temperature gradient recordings are obtained to depths of 450 feet. and are measured using the recorder cursor. The chart paper is automatically controlled to align the water temperature indication at the correct transducer depth indication. Additionally, targets displayed on the CRT can be recorded on a dry, electrosensitive, strip chart by the recorder.

Test switches on the sonar set initiate special test signals within the system to permit the operator to test all system components, either on the ground or in flight. Front panel advisory indicators illuminate to indicate a go or no-go status, and the CRT and the range rate meter indicate the test results.

The sonar hydrophone and projector combination is called the DOME. The dome is fastened to the end of a 500-foot cable, and is raised and lowered by the hydraulic cable reeling machine. When the operator depresses the



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Figure 5-39.-Major components of a modern airborne sonar set.

LOWER switch on the dome control, the reeling machine lowers the dome at the rates of 8 feet per second in air and 5 feet per second in water. When the operator presses the RAISE switch on the dome control, the dome raises at a speed of 11 feet per second up to 40 feet of depth, then 5 feet per second to the trail position. When the operator presses the SEAT switch, the dome raises from the trail position to seat position at 1 foot per second.

Emergency raising of the dome is provided in the event that the helicopter or the reeling machine hydraulic system become inoperative.

Two explosive charges, which drive the guillotine blade to cut the 500-foot cable, are installed on the reeling machine. Under emergency conditions, the pilot can jettison the dome and lowered cable by firing the two guillotine charges.

MAJOR COMPONENTS

The complete sonar system as installed in a helicopter consists of 13 separate components and 3 mounting bases, Photographs of the major units of the system are shown in figure 5-39. Refer to this figure as you read about each unit in the following paragraphs.

Azimuth and Range Indicator

The function of the azimuth and range indicator (fig. 5-40) is to display the target echoes on a PPI and to provide direct readout of bearing and range information.

The following components are mounted on the front panel:

1. A single gun, 7-inch cathode-ray tube to provide the display
2. A CURSOR INTENSITY control for regulating the brightness of the cursor circle on the CRT
3. A CRT INTENSITY control for controlling overall CRT brightness
4. A VIDEO GAIN control for regulating the gain of targets
5. An AUDIO GAIN control for setting the level of audio
6. A RANGE RATE-KNOTS meter for displaying the opening or closing speed of selected targets
7. An MTI THRESHOLD switch for selected speed threshold or targets to be displayed on the CRT
8. A DISPLAY switch for selecting either sonar or sonobuoy signals for display
9. A POWER switch for energizing or de-energizing the sonar set, except the dome control
10. TEST switches and indicators for initiating built-in test equipment (BITE) functions

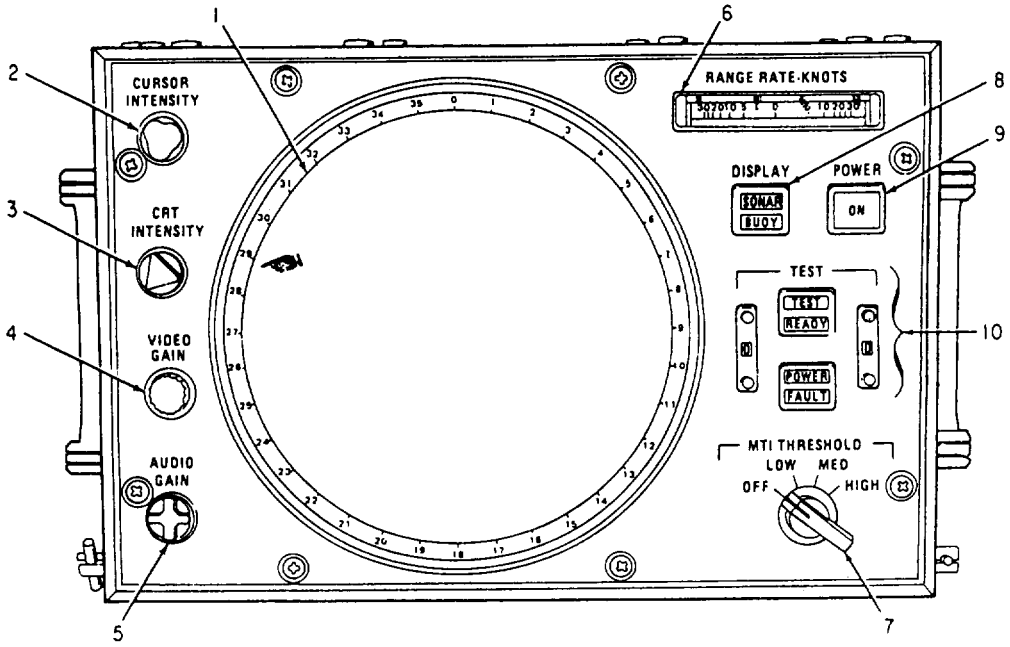


Figure 5-40.-Sonar azimuth and range indicator.

Bearing and Range Indicator

The bearing and range indicator (fig. 5-41), which is actually a remote indicator, is mounted on the pilot's instrument panel.

The remote indicator presents the pilot with digital bearing and range information (in degrees magnetic and yards) when the sonar operator sets the TARGET switch on the sonar receiver to VERIFY.

Cable Assembly and Reel

The special-purpose cable is pretensioned on its reel. The cable is 500 (±5) feet long, contains 30 shielded conductors encased in a braided steel strength member, and is protected by a waterproof, insulated, outer covering of polyurethane. Colored bands are spaced along the length of the cable to indicate the amount of cable payout. Table 5-1 lists the location and color of the printed bands.

The cable assembly and reel is bolted to the hydraulic cable reeling machine (fig. 5-39), and is connected electrically to the reeling machine slip ring assembly. The outboard end of the cable is provided with a strain relief connector that provides mechanical and electrical connections to the sonar hydrophone.

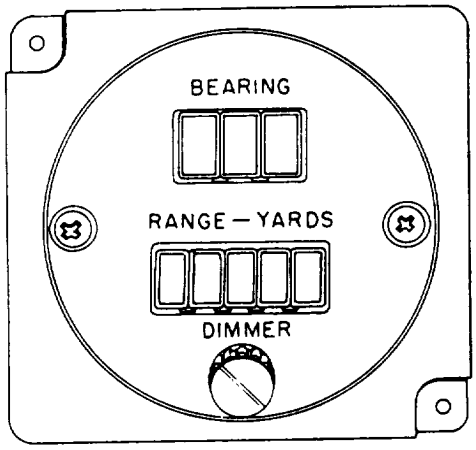


Figure 5-41.-Bearing and range indicator (remote indicator).

Table 5-1.-Length Color Code for Sonar Cable

BAND LOCATION	BAND COLOR
4 feet 6 inches	White
7 feet 6 inches	Yellow
21 feet	Red
50 feet	Green
100 feet	White
150 feet	White, Green
200 feet	White, White
250 feet	White, White, Green
300 feet	White, White, White
350 feet	White, White, White, Green
400 feet	White, White, White, White
450 feet	White, White, White, White, Green

Dome Control

The front panel of the dome control chassis, shown in figure 5-42, contains the following:

1. A three-digit DEPTH-FEET display for indicating dome depth
2. A TRAIL/UNSEATED indicator to advise the operator of the dome position
3. A RAISE/LOWER switch for energizing the reeling machine hydraulic raise or lower sequence
4. A SEAT switch/indicator for raising the dome from the trail position to the seat position and indicating when the dome is in the seat position
5. An AUXILIARY RAISE switch (with a safety cover) for raising the dome in the event of a hydraulic system malfunction

Hydraulic Cable Reeling Machine

The hydraulic cable reeling machine (fig. 5-39) uses a hydraulic motor to raise and lower the

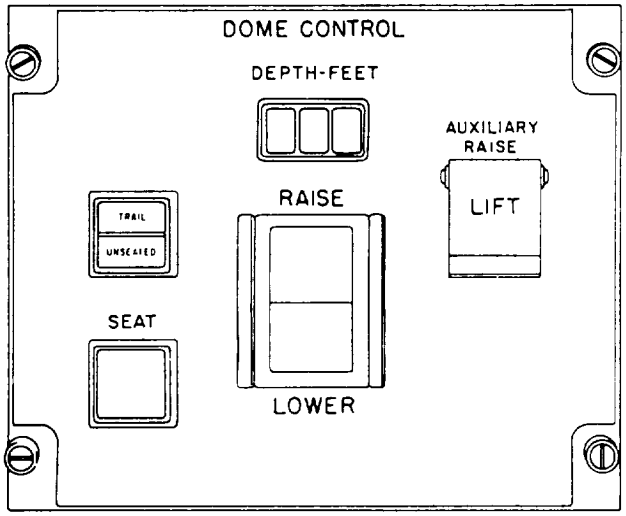


Figure 5-42.-Dome control.

dome. The sequence of raising or lowering is accomplished by energizing solenoids on the hydraulic control package, which program hydraulic pressure to release or retrieve the dome.

The reel rotates to pay out or retrieve the special-purpose electrical cable. As the cable is retrieved or payed out, a level wind assembly mounted at the top of the frame moves laterally to wind or unwind the cable evenly. The level wind is chain-driven from the gearbox assembly.

A standard one-half inch, square-drive speed wrench, which is supplied with the reeling machine, can be used as a handcrank to manually release or retrieve the cable. When the handcrank is used, the electrical circuits are disabled.

Recorder

The recorder (fig. 5-43) operates in four modes of operation: a low (25°F to 75°F) and high 45°F to 95°F) bathythermographic (BT) mode, a range

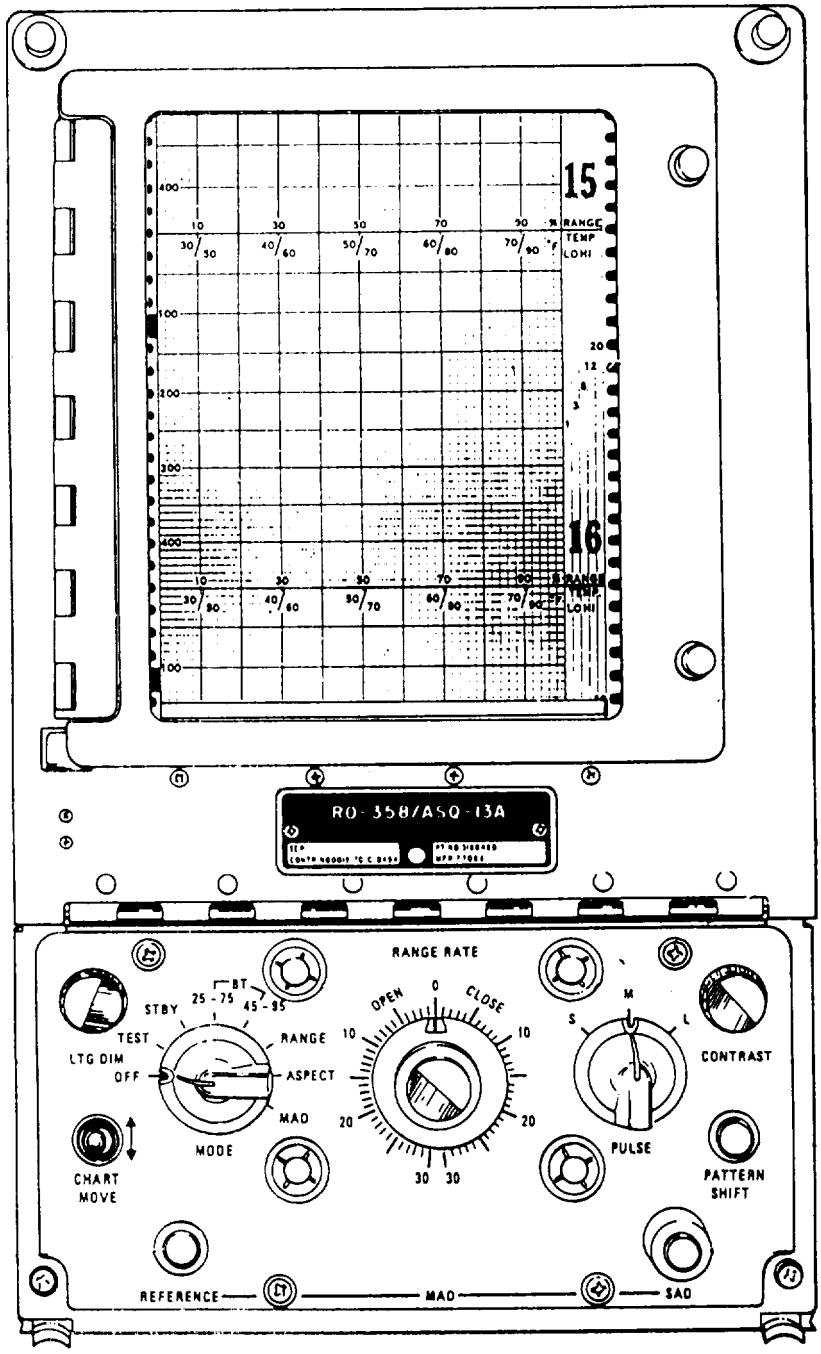


Figure 5-43.-Recorder.

mode, an aspect mode, and a self-test mode. In the BT mode, the recorder accepts temperature and depth signals from the receiver, and plots them as temperature gradients on the chart paper. In the range mode, the recorder plots video targets on the range scale in a strip chart display. In the aspect mode, the recorder accepts video signals from the receiver and plots video targets on an expanded range scale in a strip chart display. In the self-test mode, the recorder functions internally to apply the aspect transmitter keying pulses to the video input circuits.

Sonar Hydrophone and Sonar Projector

The underwater transmitting and receiving element is composed of a projector (transmitting array) and a hydrophone (receiver array). As mentioned before, the combination of hydrophone and projector, which are joined electrically and mechanically, is referred to as the DOME.

The projector assembly of the dome contains a projector, a flux gate compass, and a pressure potentiometer. The projector is composed of six matched ceramic rings (barium titanate) and a tuning transformer. The projector converts the electrical pulses from the sonar transmitter to acoustic pulses that are radiated in an omnidirectional pattern through the water. The flux gate compass forms a portion of the display stabilization loop, providing an output to indicate the sonar dome azimuth deviation from magnetic north. The pressure potentiometer provides an output to indicate depth of the dome in water. The projector is covered with a black neoprene boot that is filled with oil.

The hydrophone assembly consists of 16 stave assemblies bolted to a cork-lined fiber glass barrel, an end bell, a temperature sensor, and an electronic package. The staves, filled with oil and hermetically sealed during assembly, convert the received acoustic pulses to low-level ac signals. These signals are amplified and applied through the special-purpose electrical cable to the receiver located in the helicopter. The stave housings are stainless steel, each containing 12 matched ceramic rings with trimming capacitors, and mounted on a printed-circuit board. The output of each stave is applied to a preamplifier, located in the electronic housing. A temperature sensor for measuring temperature of the water is located in the electronic housing end bell.

The dome requires no adjustments. All inputs and outputs are made with a special electrical connector located on top of the electronic housing.

Sonar Receiver

The following components are mounted on the front panel of the receiver (fig. 5-44).

1. A RANGE SCALE-KYDS switch for selecting the operating range
2. A MODE switch for selecting the operating frequency of the sonar set
3. A CURSOR POSITION control for controlling the position of the cursor circle on the CRT in azimuth
4. A CURSOR POSITION switch for controlling the range of the cursor circle on the CRT
5. A three-digit BEARING display that indicates cursor circle bearing in degrees from magnetic north
6. A five-digit RANGE-YARDS display that indicates the range of the cursor circle in yards
7. An AUDIO switch/indicator for selecting audio from all eight sectors, or only the sector selected by the cursor circle position
8. A TARGET switch/indicator for applying bearing and range information to the pilot's bearing and range indicator

Sonar Transmitter

The following components are mounted on the front panel of the transmitter (fig. 5-45).

1. A POWER circuit breaker for applying 115 volts ac, three-phase power to the transmitter
2. A running time meter
3. An HV indicator to signal the presence of high voltage

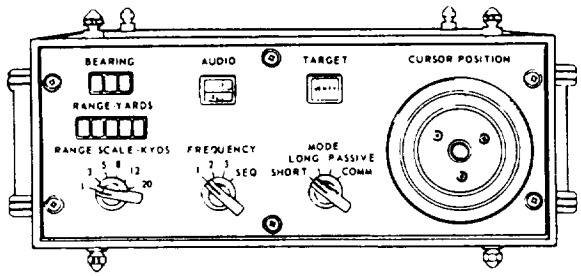


Figure 5-44.-Receiver.

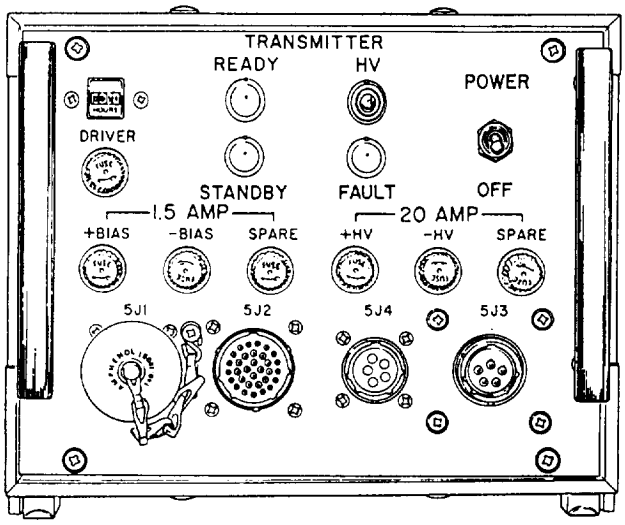


Figure 5-45.-Transmitter.

4. A READY indicator to signal when the transmitter is ready for operation

5. STANDBY and FAULT indicators to signal a malfunction in the transmitter

MODES OF OPERATION

Sonar set operating modes were mentioned earlier in the chapter. They are discussed in more detail in the following paragraphs.

The sonar set provides three operational modes: echo ranging (LONG or SHORT), PASSIVE, and COMM. A fourth mode, TEST, is used to determine that the sonar set is in operational status. Three recording modes are also provided: low (25°F to 75°F) or high (45°F to 95°F) BT (bathythermograph), RANGE, and ASPECT. A fourth recording mode, TEST, is used to determine that the recorder is in operational status.

Echo-Ranging Mode

The sonar set produces recurrent 3.5-milisecond (SHORT) or 35-milisecond (LONG) acoustic pulses that are radiated through the water from the projector portion of the dome. Returning target echoes are received by the hydrophone and processed into a left and right half-beam for each sector. Target bearing is determined by the phase difference existing between the left and right half-beams formed for each sector scanned. Bearing of the target is resolved from the edge of each of the eight 45-degree sectors scanned. Target range is determined by the elapsed time between transmission of a given pulse and the return of a target echo. Bearing and range are presented simultaneously as a single target pip on the CRT. Variations of the speed of sound in water due to temperature of the water surrounding the dome are compensated for automatically.

An audio signal is developed for each returning target echo. These audio signals are applied to the helicopter intercommunication system in such a manner that signals representing the left four sectors of the CRT are applied to the left earphone, and signals for the right four sectors are applied to the right earphone. A different nonharmonic tone is generated for each of the four sectors in each CRT half when the AUDIO switch is in the ALL position. In the ONE position, the audio representing the CRT sector in which the cursor is positioned is applied to both earphones.

The nature of the object causing the echo can be determined by the outline and intensity of the target display on the CRT as well as by the quality and intensity of the audio. The opening or closing speed of the target within the cursor circle is displayed automatically on the RANGE RATE-KNOTS meter.

Passive Mode

In the PASSIVE mode, active echo-ranging is disabled, and underwater sounds may be received and displayed on the CRT. Bearing information is presented in this mode of operation and appears in the form of a noise spoke on the CRT. Audio is presented in the same manner as in the echo-ranging mode.

Communication Mode

The COMM mode is used for two-way underwater voice communication with other appropriately equipped helicopters, ships, or submarines operating within range. Voice communication operation is activated by placing the MODE switch to COMM. Voice transmission is accomplished by depressing a foot switch and speaking into the microphone. Releasing the foot switch permits monitoring voice signals from other similar underwater communication systems.

When the AUDIO switch is set to ONE, reception of underwater voice signals is accomplished by placing the cursor circle in the CRT sector in which the noise spoke appears and by regulating the AUDIO GAIN control.

Test Modes

The test modes check the operational status of the system as a whole and the various components of the system as individual units. These test modes use internally generated stimuli.

During normal operation, the test circuits sample major system functions and voltages. If a sampled function exceeds preset limits, the FAULT indicator illuminates for the length of time that the fault exists.

Recorder Bathythermographic Mode

The recorder bathythermographic (BT) mode is used to obtain graphs of temperature gradients appearing beneath the surface of the surrounding water to depths of 450 feet, Temperature and depth signals obtained from the dome are processed by the receiver and dome control. These signals are applied to the recorder circuits when the recorder MODE selector switch is positioned to the BT mode position. The recorder chart drive circuits automatically position the chart paper to provide correct chart registration. Recorded scale marks on the chart paper denote the temperature scale being used for each temperature recording. The recorder plots temperature on the vertical axis and depth on the horizontal axis of the moving chart.

Recorder Range Mode

The recorder RANGE mode is used to obtain continuous strip-chart displays of target echo ranges. Range scale control signals from the receiver RANGE SCALE-KYDS switch are accepted by the recorder sweep circuits to correlate the range sweeps. As the chart paper moves, range scale marks are recorded on the chart paper to denote the range scale being used for each range recording. Target echo video signals are applied to the styluses when they appear in time, as related to the range sweep. The video signals are recorded each time a stylus passes over the range position of a target. The chart advances a small increment for each stylus sweep.

Recorder Aspect Mode

The recorder ASPECT mode is used to obtain continuous strip-chart displays of target echo signals. Timing and control signals, generated within the recorder, slave the receiver timing circuits to alternate sweep ramps between transmit and receive cycles. During each transmit sweep ramp, a train of short keying pulses is generated and pulse width is regulated in the recorder. This pulse train is applied to the receiver. During each receive sweep ramp, the train of received target echo video pulses is applied to the recorder styluses. Target echo signal level is neither limited nor affected in the system. This permits varying intensity recordings (highlights) of target structural characteristics for optimum target classification.

Recorder Test Mode

The sonar operator uses the recorder TEST mode to check the operational status of the recorder. The recorder TEST mode effectively checks the operation of the recorder stylus drive, stylus write, and chart drive operations. In addition, all front panel controls on the recorder can be checked by the operator for operational compliance and accuracy.

SONAR DATA COMPUTER

A recent improvement in the airborne sonar field has been the use of the sonar data computer (SDC) with sonar sets. This computer is mated to the existing sonar set to provide processing and display of LOFAR, DIFAR, and CASS sonobuoy signals on the sonar set's CRT. The SDC is also used to provide a more accurate fix on the target by providing a digital readout of target range, speed, and bearing.

ASW NONACOUSTIC SYSTEM

Learning Objective: *Describe ASW nonacoustic system equipment characteristics, to include magnetic anomaly detection and magnetic compensation.*

The AN/ASQ-81 MAD system signal flow diagram shown in figure 5-46 represents the typical signal flow within a platform. The overall system includes the ASA-64/ASA-71, the ASA-65, and associated system components involved in the localization and detection of submarines. The ASA-64, Submarine Anomaly Detecting System (SAD) (coupled with the ASA71 system) recognizes submarine signals and marks them automatically in the presence of geologic, maneuver, geomagnetic, and equipment noise as detected by the ASQ-81. The ASA-65 system is

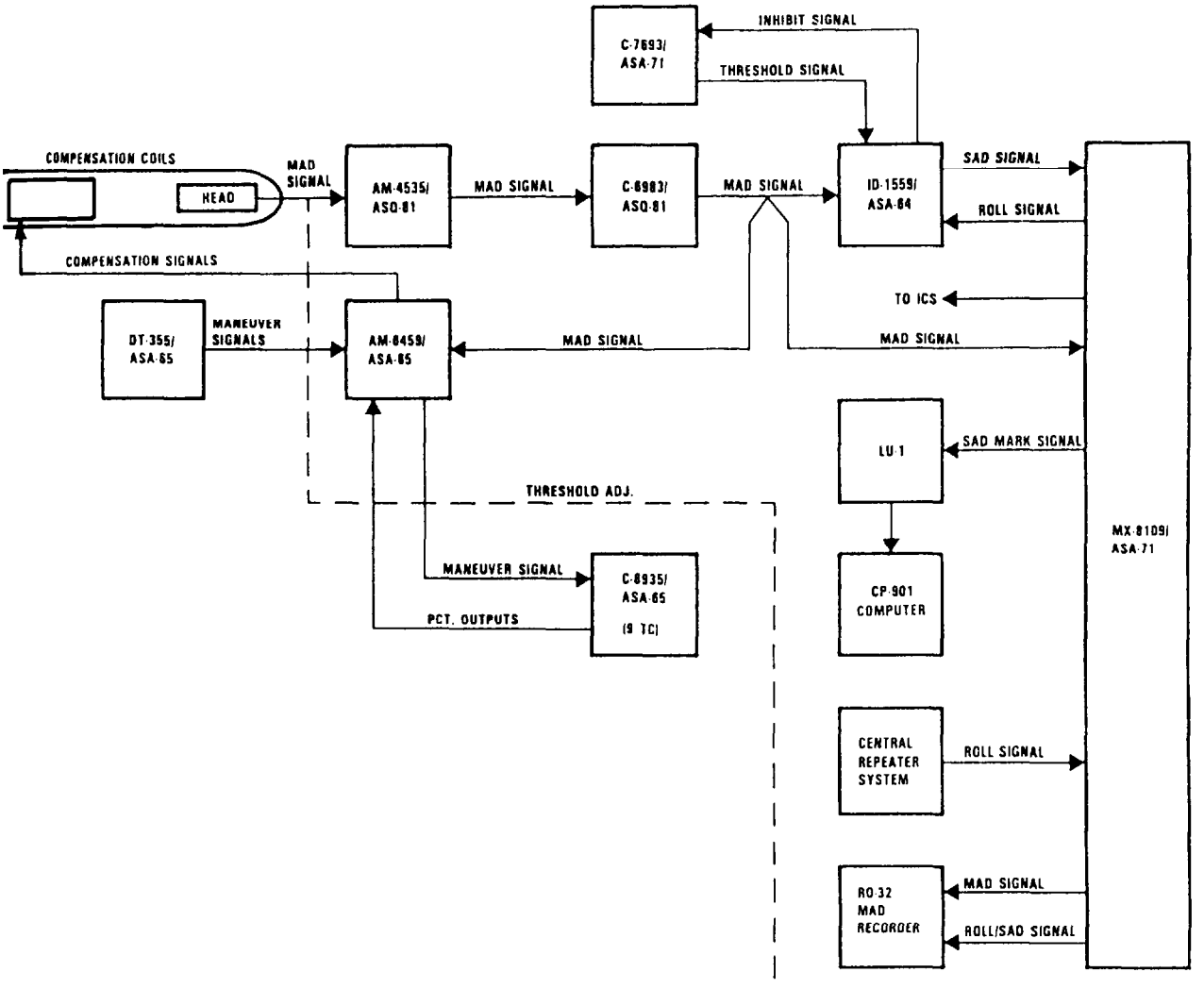


Figure 5-46.-ASQ-81 signal flow diagram.

the magnetic compensator group, which compensates for magnetic fields generated by the aircraft in flight. Before we can discuss the individual systems involved in the MAD problem, we must look at some principles and theory.

PRINCIPLES OF MAGNETIC DETECTION

Light, radar, and sound energy cannot pass from air into water and return to the air in any degree that is usable for airborne detection. On the other hand, lines of force in a magnetic field are able to make this transition almost undisturbed because the magnetic permeability of water and air are practically the same. Specifically, the lines of force in the earth's magnetic field pass through the surface of the ocean essentially undeviated by the change of medium (from water to air or vice versa) and undiminished in strength. Consequently, an object under the water can be detected from a position in the air above if the object has magnetic properties that distort the earth's magnetic field. A submarine has sufficient ferrous mass and electrical equipment to cause a detectable distortion (anomaly) in the earth's field. Detection of this anomaly is the function of magnetic anomaly detection (MAD) equipment.

Magnetic Anomaly

The lines comprising the earth's natural magnetic field do not always run straight north and south. If traced along a typical 100-mile path, the field twists at places to east or west, and assumes different angles with the horizontal. Angles of change in the east-west direction are known as angles of variation, while angles between the lines of force and the horizontal are

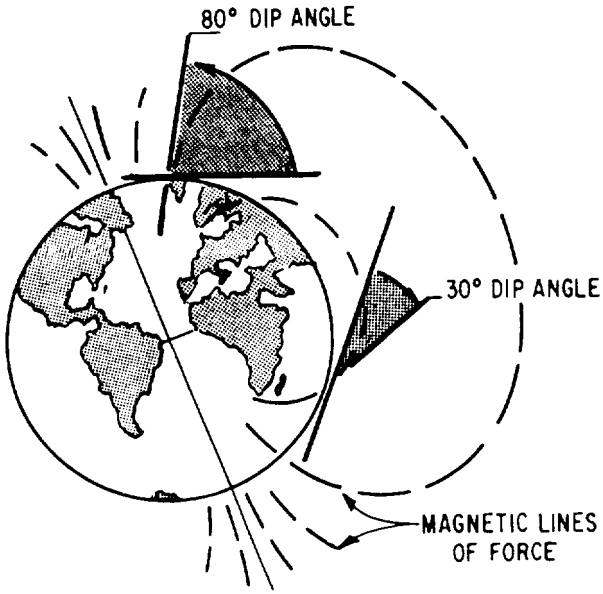


Figure 5-47.-Dip angles.

known as angles of dip (fig. 5-47). At any given point between the equator and the magnetic poles, the relationship of the angle between the earth's surface and the magnetic lines of force is between 0 degrees and 90 degrees. This angle is determined by drawing an imaginary line tangent to the earth's surface and to the line of force where it enters the earth's surface. The angle thus formed is called the dip angle.

If the same lines are traced only a short distance, 300 feet for instance, their natural changes in variation and dip over such a short distance (short-trace) are almost impossible to measure. However, short-trace variation and dip in the area of a large mass of ferrous material, though still extremely minute, are measurable with a sensitive anomaly detector. This is illustrated in figure 5-48. The dashed lines represent lines of force in the earth's magnetic field.

View A shows the angular direction at which natural lines of magnetic force enter and leave the surface of the earth. Note that the angles of dip are considerably steeper in extreme northern and southern latitudes than they are near the equator. View B represents an area of undisturbed natural

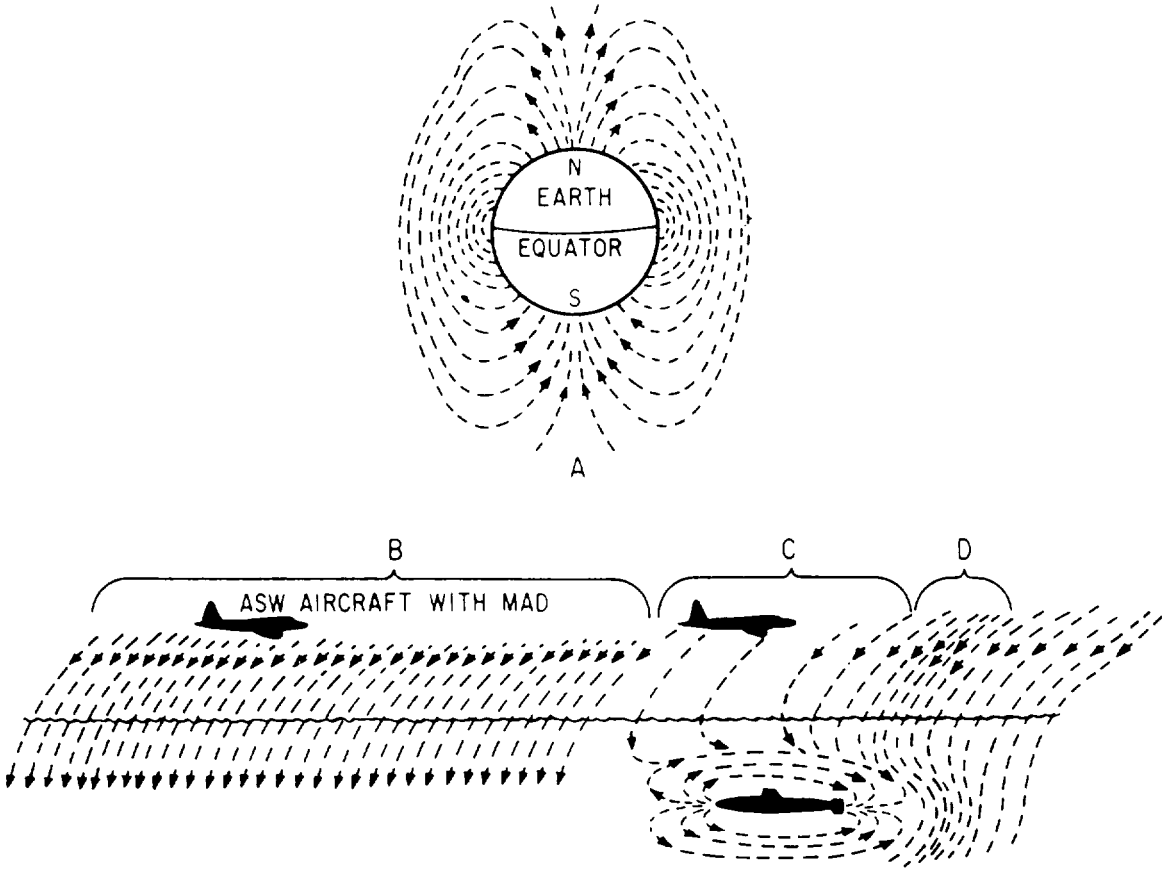


Figure 5-48.-Simplified comparison of natural field density and submarine anomaly.

magnetic strength. In views C and D, the submarine's magnetic field distorts the natural field as shown. The density of the natural field is decreased in view C and increased in view D. The natural angle of dip is also affected, but only very slightly.

Submarine Anomaly

The maximum range at which a submarine may be detected is a function of both the intensity of its magnetic anomaly and the sensitivity of the detector. A magnetometer is used as the detector in MAD equipment. Magnetometer theory is discussed later in this chapter.

A submarine's magnetic moment (magnetic intensity) (fig. 5-49), which determines the intensity of the anomaly, is dependent mainly on the submarine's alignment in the earth's field, its size, the latitude at which it is detected, and the degree of its permanent magnetization.

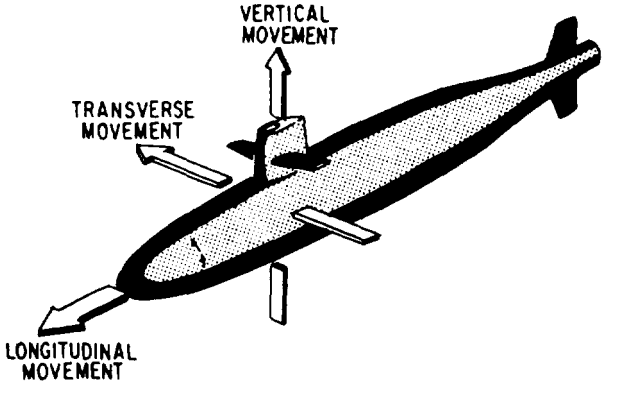


Figure 5-49.-Submarine's magnetic moment.

MAD equipment, in proper operating condition, is very sensitive, but the submarine's anomaly, even at a short distance, is normally very weak. The strength of a complex magnetic field (such as that associated with a submarine) varies as the inverse cube of the distance from the field's source. That is, if the detectable strength of a field source has a given value at a given distance and the distance is doubled, the detectable strength of the source at the increased distance will then be one-eighth of its former value. Therefore, from the foregoing, at least two facts should be clear. First, MAD equipment must be operated at a very low altitude to gain the greatest proximity possible to enemy submarines. Second, the searching aircraft should fly at a predetermined speed and follow an estimated search pattern. This ensures systematic and thorough searching of the prescribed area so that no existing anomalies are missed.

Anomaly Strength

Up to this point, the inferred strength of a submarine's anomaly has been exaggerated for purposes of explanation. Its actual value is usually so small that MAD equipment must be capable of detecting a distortion of approximately one part in 60,000. This fact is made apparent by pointing out that the direction of alignment of the earth's magnetic lines of force is rarely changed more than one-half of 1 degree in a submarine anomaly.

Figure 5-50, view A, represents a contour map showing the degree of anomaly caused by a

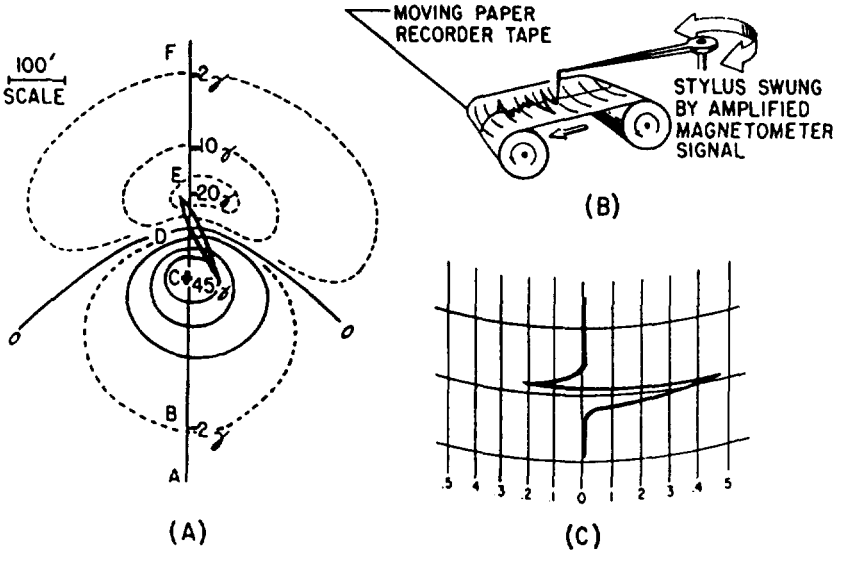


Figure 5-50.-(A) Degree of anomaly; (B) anomaly stylus; (C) sample anomaly record.

submarine. The straight line is approximately 800 feet in length and represents the flight path of a searching aircraft through the area of the submarine anomaly. If the submarine were not present, the undisturbed magnetic intensity in the area due to its assumed natural characteristics would be 60,000 gammas. (The gamma, symbolized by the Greek letter, is a measure of magnetic intensity.) All variations in the field, when the submarine is present, would then be above or below this natural intensity. Therefore, 60,000 gammas is the zero reference drawn on the moving paper tape shown in figure 5-50, view C.

Refer to view A of figure 5-50. Starting with the aircraft at point A, where the anomaly is undetectable, the earth's field concentration decreases to an intensity of –2 (59,998) at point B, Its intensity then increases until a peak value of +45 is reached at point C. From that point it decreases to zero at point D. Beyond point D, another zone of what amounts to magnetic rarefaction is encountered. The earth's field is less intense than its normal value. Consequently, anomalous values in this zone are considered as minus quantities. A peak minus intensity is reached at point E, and thereafter, the signal rises back to its normal, or undetectable, intensity at point F.

As the varying degrees of intensity are encountered, they are amplified and used to drive a swinging stylus, as shown in figure 5-50, view B. The tip of the stylus rides against the moving paper tape, leaving an ink trace. (Some recorders use electrosensitive paper tape and a chain-driven stylus. Some other recorders use a CRT-type display.) The stylus is swung in one direction for positive and in the opposite direction for negative. The magnitude of its swing is determined by the intensity of the anomaly signal. Figure 5-50, view C, is a sample of paper recording tape, showing the approximate trace caused by the anomaly in view A.

In the illustration just given, the search aircraft's altitude was 200 feet. At a lower altitude, the anomaly would have been stronger, and at a higher altitude, it would have been weaker.

MAGNETIC NOISE

For the purpose of this discussion, any noise or disturbance in the aircraft or its equipment that could produce a signal on the recorder is classified as magnetic noise.

Noise Sources

In an aircraft there are many sources of magnetic fields, such as engines, struts, control cables, equipment, and ordnance. Many of these fields are of sufficient strength to seriously impair the operation of MAD equipment. Consequently, some means must be employed to compensate for "magnetic noise" fields. The noise sources fall into two major categories—maneuver noises and dc circuit noises.

MANEUVER NOISES.— When the aircraft maneuvers, the magnetic field of the aircraft is changed, causing a change in the total magnetic field at the detecting element. The aircraft maneuver rates are such that the signals generated have their major frequency components within the bandpass of the MAD equipment.

Maneuver noises may be caused by induced magnetic fields, eddy current fields, or the permanent field.

The variations in the induced magnetic field detected by the magnetometer are caused by changes in the aircraft's heading. This causes the aircraft to present a varying size to the earth's magnetic field, and only the portion of the aircraft parallel to the field is available for magnetic induction.

Eddy current fields produce maneuver noise because of currents that flow in the aircraft's skin and structural members. When an aircraft's maneuver causes an eddy current flow, a magnetic field is generated. The eddy current field is a function of the rate of the maneuver. If the maneuver is executed slowly, the effect of the eddy current field is negligible.

The structural parts of the aircraft exhibit permanent magnetic fields, and, as the aircraft maneuvers, its composite permanent field remains aligned with it. The angular displacement between the permanent field and the detector magnetometer during a maneuver produces a changing magnetic field, which the detector magnetometer is designed to detect.

DC CIRCUIT NOISES.— The dc circuit noise in an aircraft comes from the standard practice in aircraft design of using a single-wire dc system with the aircraft skin and structure as the ground return. The resulting current loop from the generator to load to generator serves as a large electromagnet that generates a magnetic field similar to a permanent magnetic field. Whenever the dc electrical load of the aircraft is abruptly changed, there is an abrupt change in the magnetic field at the detector.

Compensation

Regardless of its source, strength, or direction, any magnetic field may be defined in terms of

three axial coordinates. That is. it must act through any or all of three possible directions—longitudinal, lateral, or vertical—in relation to the magnetometer detector.

Compensation for magnetic noises is necessary to provide a magnetically clean environment so that the detecting system will not be limited to the magnetic signal associated with the aircraft itself.

Experience has shown that the induced fields and eddy current fields for a given type of aircraft are constant. That is, from one aircraft to another of the same type, the difference in fields is negligible. These fields may be expected to remain constant throughout the life of the aircraft, provided significant structural changes are not made. In view of these factors, the aircraft manufacturer provides compensation for induced fields and eddy current fields.

Eddy current field compensation is usually achieved by placing the detector magnetometer in a relatively quiet magnetic area. In some aircraft the magnetometer (detecting head) is placed at least 8 feet from the fuselage. This is done by enclosing the detecting head in a fixed boom (fig, 5-51, view A), or in an extendable boom (fig. 5-51, view B). Helicopters tow the detector head by use of a cable (fig. 5-51, view C).

Induced magnetic field compensation is accomplished by using Permalloy strips. The aircraft is rotated to different compass headings, and the magnetic moment is measured. The polarity and the variation of the magnetic moment are noted for each heading, and Permalloy strips are oriented near the detector magnetometer to compensate for field changes due to aircraft rotation. Additional compensation is needed for the longitudinal axis, and is provided for by the development of outrigger compensators of Permalloy near the detecting element.

Permanent field compensation must be done in three dimensions rather than in two, and it is accomplished by three compensating coils mounted mutually perpendicular to each other (fig. 5-52, view A). The aircraft is rotated in 5-degree and 10-degree steps around its three axes. Adjustment of the field strength is accomplished by controlling the amount of direct current that flows through a particular coil. Figure 5-52, view B, shows a circuit for a single compensating coil.

Compensation for the dc magnetic field is accomplished by using electromagnetic compensating loops. The loops are arranged to provide horizontal, vertical, and longitudinal fields, and are adjusted to be equal and opposite to the dc magnetic field caused by the load current. The compensating loops are connected across a variable resistor for a particular distribution center, and are adjusted to allow current flow proportional to the load current for correct compensation. Different types of aircraft have several sets of compensating loops, depending upon the number of distribution centers. In newer aircraft, production changes have been made to use ground return wires to minimize loop size.

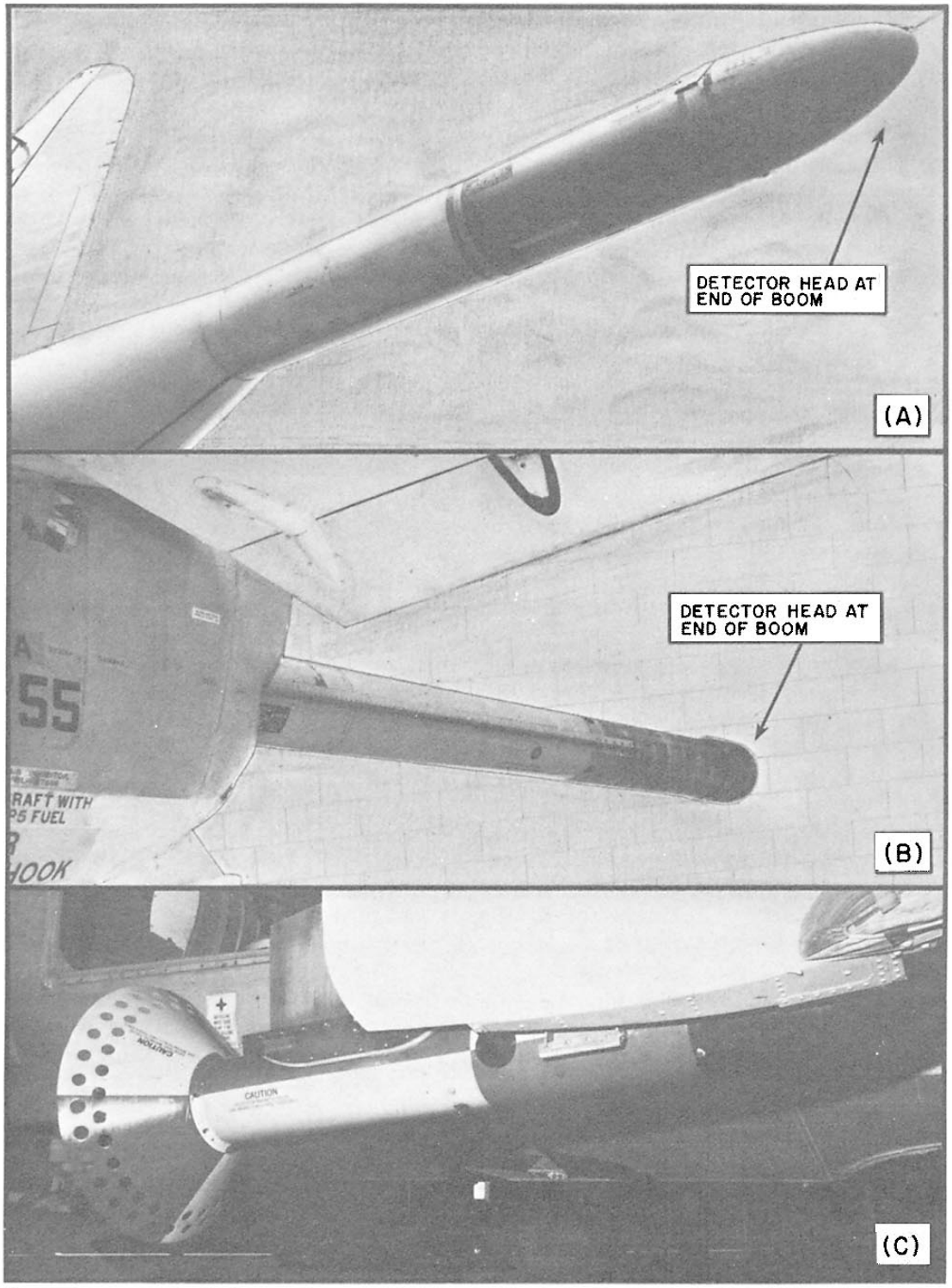
The procedure for adjustment of the dc compensation system makes use of straight and level flight on the four cardinal headings. For example, actuation of a cowl flap motor will cause dc field changes representative of those caused by any nacelle load. The load is energized, the size and polarity of the signal are noted, and the compensation control is adjusted. The load is re-energize, and the compensation control is adjusted again. Adjustments are continued until the resulting signals from the dc field are minimized.

Under ideal conditions, all magnetic fields tending to act on the magnetometer head would be completely counterbalanced. In this state the effect on the magnetometer is the same as if there were no magnetic fields at all. This state exists only when the following ideal conditions exist:

1. The aircraft is flying a steady course (no maneuvers) through a magnetically quiet geographic area.
2. Electric or electronic circuits are not turned on or off during compensation.
3. Direct current of the proper intensity and direction has been set to flow through the compensation coils, so that all stray fields are balanced.

To approximate these conditions, the compensation of MAD equipment is usually performed in flight, well at sea. In this way the equipment is compensated under operating conditions, which closely resemble those of actual ASW search flights.

From the foregoing, it should be clear that the objective of compensation is to gain a state of total balance of magnetic forces around the magnetometer. Thereafter, any sudden shift in one of the balanced forces (such as an anomaly in the earth's field force) upsets the total balance. This imbalance is indicated on the recorder. Unfortunately, a shift in ANY of the balanced forces will be indicated. Shifts in any of the forces other than the earth's natural field are regarded as noise.



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Figure 5-51.–(A) Stationary detector boom; (B) extendable detector boom; (C) cable-deployed towed detector .

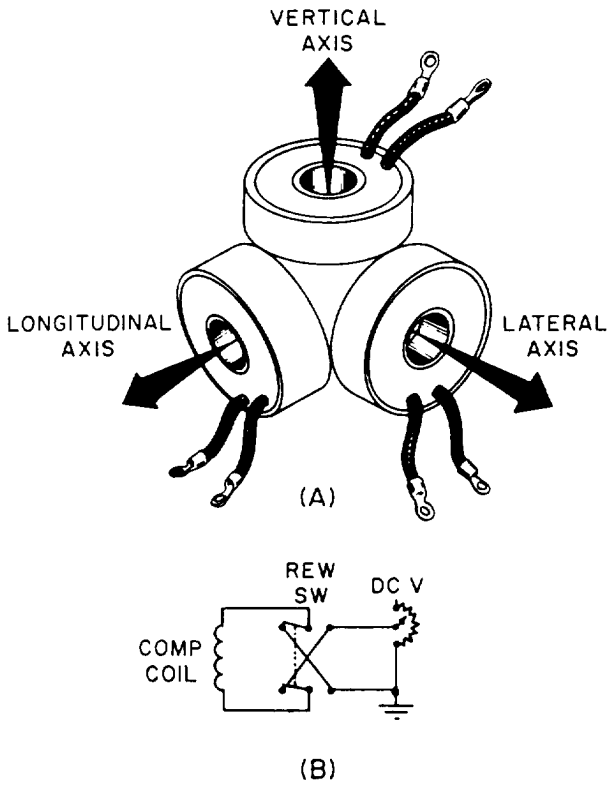


Figure 5-52.-(A) Arrangement of compensating coils; (B) compensating coil circuit.

SOLID-STATE MAGNETOMETER THEORY

Learning Objective: *Recognize the theory of operation of solid-state MAD magnetometers, including nuclear resonance, the Larmor frequency, the metastable state, and the use of helium in magnetometers.*

Magnetometers currently used in Navy MAD systems, such as the AN/ASQ-81 system, are solid-state devices. Their theory and operation are discussed in the following text.

MAGNETIC NUCLEAR RESONANCE

Magnetic nuclear resonance is based on the theory of atomic structure, which states that "the atom is considered to be a nucleus around which one or more electrons are orbiting." The nucleus has a positive charge because protons are part of it. The electrons have a negative charge that causes the atom to be neutral. The electrons and the nucleus have a spin and, because of the spin and the charge, a magnetic moment results.

The electron also is spinning about its own axis (much like the earth orbits around the sun once a year and spins about its axis once a day). This spin causes the electron to have a magnetic moment, much like a small magnet. and to exhibit the characteristics of a tiny gyro. As is the case with a mechanical gyro, a force applied to the electron causes it to precess, resulting in the wobble motion of the electron's spin axis, as shown in figure 5-53, view A. Furthermore, the

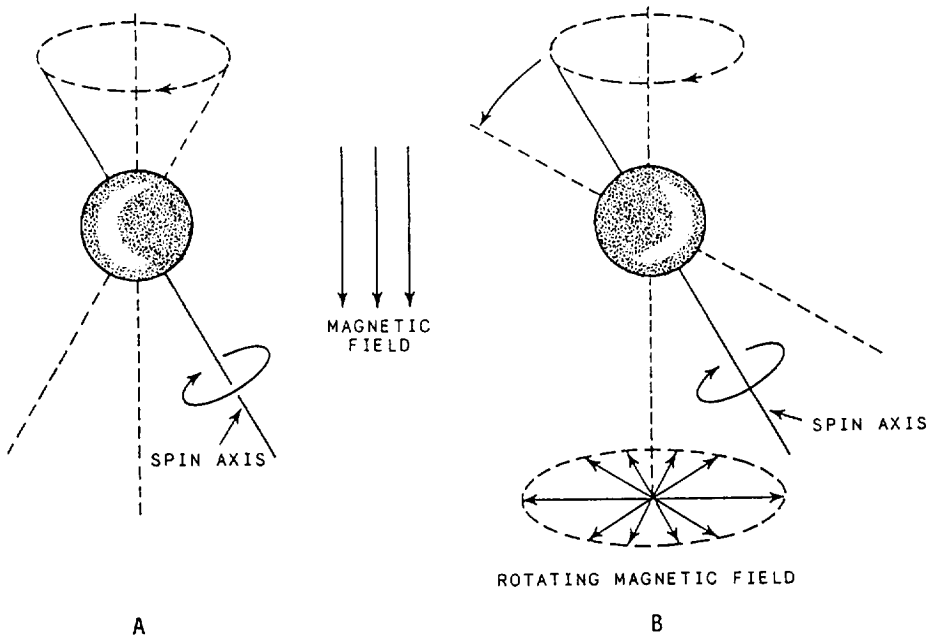


Figure 5-53.-Electron precession.

magnetic characteristics of an electron make it possible to substitute a magnetic field force (earth's magnetic field) for the mechanical force normally used to precess a conventional gyro. In addition to this, if a rotating magnetic field at a radio frequency is applied perpendicularly to the main magnetic field, the electron precesses further. This condition is depicted by the dash lines in figure 5-53, view B, as an increased wobble motion of the electron's spin axis.

When the frequency of rotation of the magnetic field is adjusted until it is the same as the natural frequency of the particular material in use (helium in this case), the deviation of the spin axis of the electrons tends to increase and paramagnetic resonance is achieved. Electron paramagnetic resonance is resonance in which the electron is the only particle shifting energy states. Resonance occurs when the angular velocity of the rotating magnetic field is approximately the natural spin axis wobble rate, or precession rate, of helium electrons. This wobble rate is also called the Larmor frequency. As the electrons are caused to precess more by the external RF magnetic field, the amplitude of the precession becomes so great that the electrons jump to a higher energy level, at which time light energy is absorbed by the helium. Light is used initially to increase the energy level of the electrons to a metastable energy state, which is a higher energy level with a much longer lasting duration than any other excited level. Helium gas is one of the elements or materials that can assume a metastable energy level.

The ASQ-81 solid-state MAD system uses a beam of low-frequency light to periodically increase the energy level of the electrons and orient the magnetic moments of the atoms and their electrons on a plane in the direction of the light beam. This procedure is known as optical pumping. The pumping action of the light energy causes the electrons to jump two energy levels; that is, from ground energy level (E_0) to excited energy level (E_2). Excited energy state 2 (E_2) has a very short lifetime without external excitation. Since the light is pumping, the electrons tend to fall back toward the stationary state (E_0), but must pass through the metastable excited state—energy state 1 (E_1).

If the pumping action is made continuous, the majority of electrons at any one time will congregate at the (E_1) or metastable excited state. This is necessary because if resonance of unaligned ground state helium were attempted with an external rotating magnetic field, only a small percent of the gas would change state and absorb light energy. This would make detection of resonance difficult. However, with the helium's atomic system aligned and excited when resonance occurs, a great majority of the atoms change state and an easily monitored amount of light is absorbed. The metastable state in the gas is necessary to have a relatively stable higher energy level in a much larger number of atoms. This produces a much greater change when resonance occurs.

As explained later, optical energy is absorbed and released by the electron during these energy level transitions. It should be noted here that resonance can apparently be achieved only in certain solids and liquids with loosely knit atomic structures and in gases such as helium.

This discussion so far has concerned the energy level change of the helium atom electrons by the use of an external magnetic field (supplied through coils) oriented 90 degrees to the earth's field. The external field rotates at the Larmor frequency of the electron, which was determined by the earth's magnetic field strength. Since the precession or Larmor frequency of electrons varies with magnetic field intensity at the rate of 28 Hz per gamma, monitoring the Larmor frequency changes is a convenient method for detecting and measuring changes in the earth's magnetic field.

However, a small change in the Larmor frequency of electrons is difficult to measure directly. It is more convenient and accurate to make an indirect measurement by monitoring the definite increase in light energy absorption that occurs at resonance. The atoms then radiate some of this light energy as the Larmor frequency shifts away from the externally applied field's frequency, and resonance is lost. Thus, variations in the earth's magnetic field strength are reflected by the changes in the quantity of light passing through the helium gas. Helium gas is contained in a transparent container called an absorption cell.

As shown in figure 5-54, view A, energy is absorbed in forcing electrons to jump from one

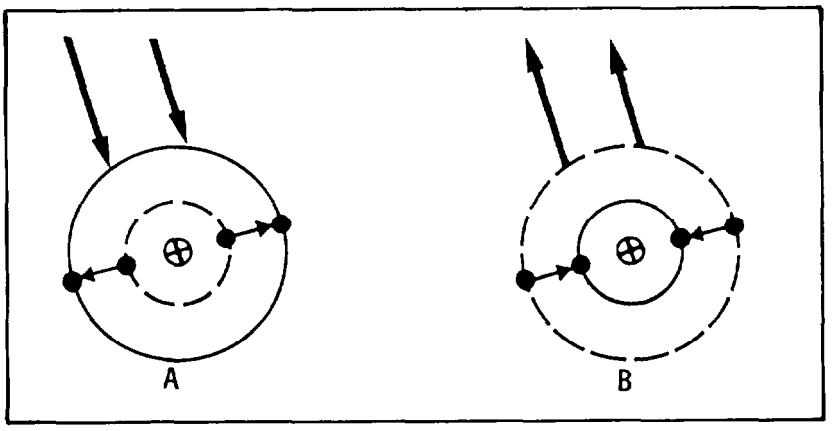


Figure 5-54.-Helium atom (A) absorbing energy and (B) radiating energy.

orbit to another orbit at a greater distance from the nucleus. Figure 5-54, view B, depicts the atom giving up energy as the electrons move back to their original orbit.

When the oscillator supplying the coils (resonance oscillator) reaches the Larmor frequency, light is absorbed. When the resonance oscillator comes off the Larmor frequency, light is emitted. Thus, the Larmor frequency (which represents the earth's magnetic field strength) can be tracked by knowing the resonant oscillator frequency when the absorption cell is absorbing energy.

As shown in figure 5-55, the light energy source for the magnetic resonance magnetometer is a helium lamp. The infrared (IR) detector, which is a very sensitive and trouble-free device, is used to track the light energy level changes.

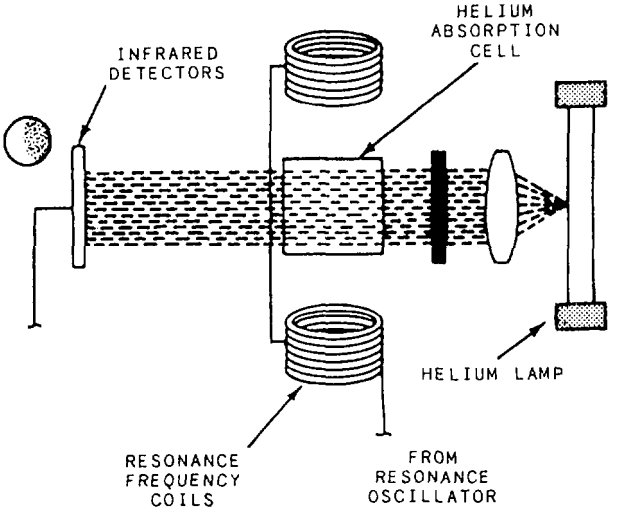


Figure 5-55.-Essential parts of helium magnetometer.

HELIUM MAGNETOMETER OPERATION

The necessary components for a magnetic resonance magnetometer are shown in figure 5-55. The external energy source, the helium discharge lamp, applies light energy to the transparent helium-filled absorption cell. Light energy is absorbed or given up by the helium atoms as the cell goes in and out of resonance, caused by slight changes in the earth's field. The IR detector, mounted on the other side of the absorption cell from the light source, picks up these energy level changes and produces an electrical output. The energy level is low when the cell is resonant and high when it is not resonant. The IR detector output signal can then be used to keep the resonance oscillator on or about the Larmor frequency of the helium in the cell, regardless of changes in the earth's magnetic field. A magnetic anomaly can then be detected when the resonance oscillator center frequency makes atypical swing within the bandpass relative to the flight envelope of the MAD-equipped type of aircraft.

If the resonance oscillator frequency were always maintained precisely at resonance, the IR output would simply be a very low dc level. This low dc level would be difficult to monitor and would not provide a phase change above and below resonance. For this reason the RF energy applied to the resonance coils is frequency

modulated by 430 Hz. The resultant variation in the magnetic field forces the helium atoms in and out of resonance around the null and provides a phase reference signal to a phase detection circuit. Figure 5-56 depicts the phase reversal above and below resonance. Note also that at resonance, which is the normal operating point, the IR detector acts much like a full wave rectifier. As a result of rectifying action, the IR output at resonance is a pulsating 860-Hz signal. The detector also acts like a discriminator, in that the FM swing of the resonance oscillator is converted to an AM output when near the resonant frequency of the helium atoms in the absorption cell. This is true because its output goes positive with each increase in light energy given off by the absorption cell.

SOLID-STATE MAD SYSTEM

Learning Objective: *Recognize the components and operating principles of the AN/ASQ-81 MAD system.*

The ASQ-81 has different configurations depending on platform. Some of the possible weapons replaceable assemblies (WRAs) include:

1. Detecting Set Control, C-9086/ASQ-81
2. Amplifier Power Supply, AM4535/ASQ-81
3. Towed Body TB-623/ASQ-81 (magnetic detector)

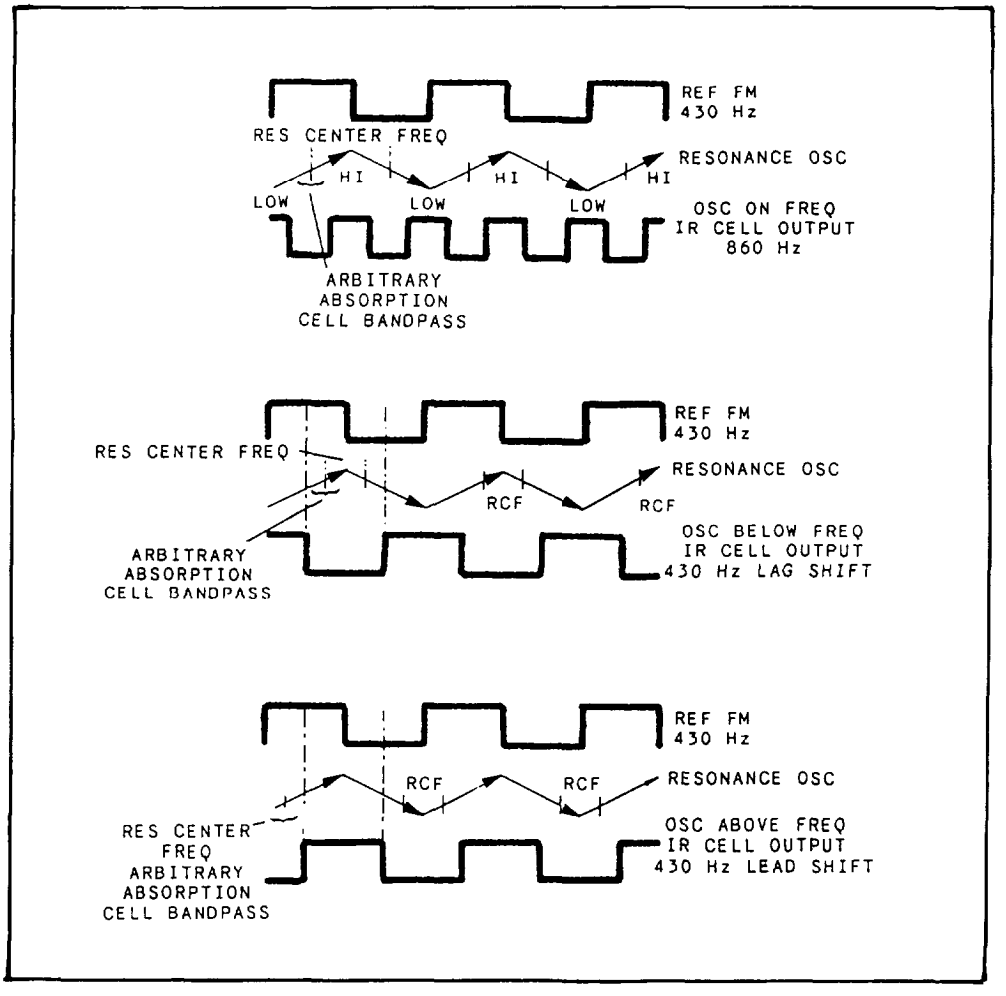


Figure 5-56.-Resonance waveforms from a phase detection circuit.

4. Reeling Machine Control, C6984/ASQ-81

5. Reeling Machine RL 305/ASQ-81

The AN/ASQ-81 system is typical of the solid-state MAD systems currently used on Navy ASW aircraft. The basic platform includes the magnetic detector, the amplifier-power supply, and the detecting set control. Figure 5-57 shows the controls and indicators as they appear on the equipment. The magnetic detecting set is a sensitive metastable helium magnetometer used to locate and classify submarines by detecting a disturbance or change in the normal earth's magnetic field.

MAGNETIC DETECTOR

The description of a helium magnetometer given in preceding paragraphs is for the basic sensing element. The actual arrangement of the magnetometer is shown in figure 5-58.

The detection element includes six separate helium absorption cells and six IR detectors, arranged in pairs, with the pairs oriented at 90 degrees to each other. This configuration ensures that one or more of the pairs is at least partially in line with the earth's field regardless of aircraft attitude or direction of flight. The signals from

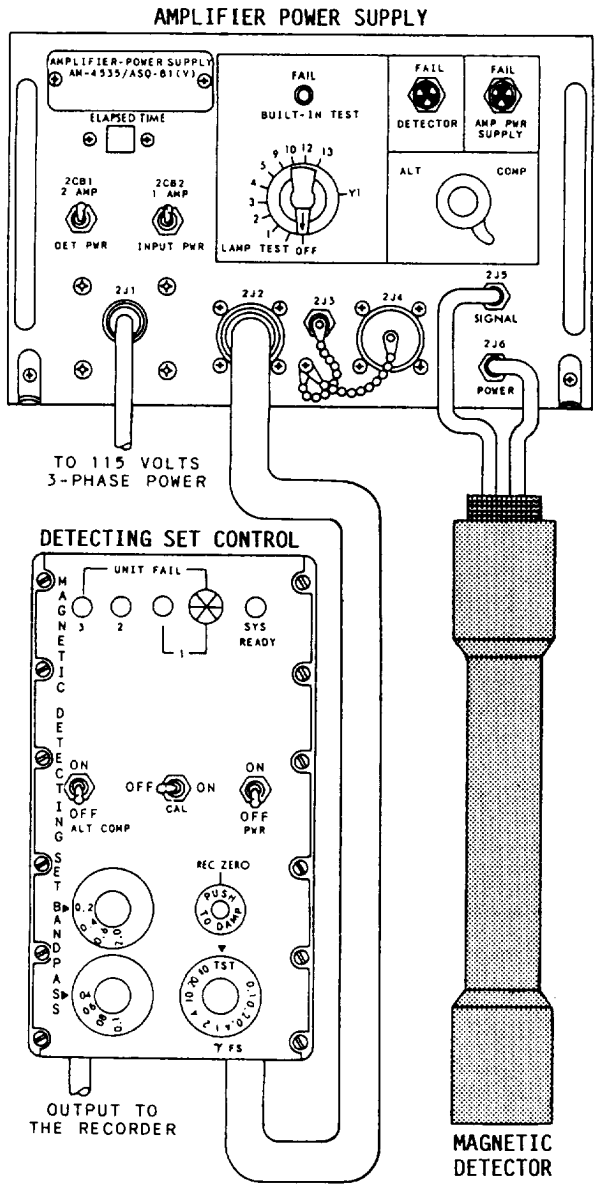


Figure 5-57.-MAD system units.

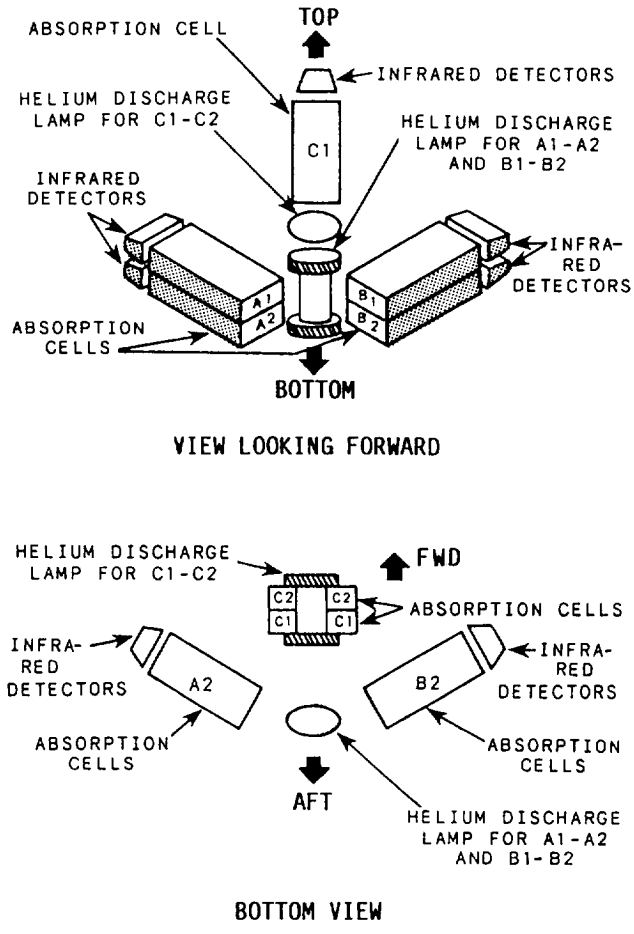


Figure 5-58.-Magnetometer axis orientation.

all three detector pairs are combined in a summing amplifier; thus, the final output to the amplifier-power supply is not affected by aircraft maneuvers.

Two helium discharge lamps provide light energy to the three absorption cell pairs in the magnetic detector. The lamps are ignited by a 52-k Hz, 1500-volt supply (fig. 5-59). After ignition the lamps are maintained in an ionized state by the 49.6-MHz output of the exciter-regulator.

In addition, the magnetic detector unit includes a pressure transducer and an altitude compensator circuit. The output of the pressure transducer varies the frequency of a 5.4-kHz oscillator in the altitude compensator at the rate of 1 Hz/ft of altitude change. The total swing of the oscillator is 5.0 to 5.8 kHz, which can compensate for a maximum of 400-foot rapid altitude variations. Additional circuitry in the amplifier-power supply converts the altitude-induced frequency change to a varying dc level. This dc level may, at the operator's discretion, be used to correct the final signal output.

CAUTION

When maintenance is performed on or near the magnetometer head, special non-ferrous tools MUST be used. Additionally, any parts (nuts, bolts, or screws) that are replaced MUST be made of a special non-ferrous metal. Always consult the MIM for the system being worked on before attempting repairs.

AMPLIFIER-POWER SUPPLY

A single coaxial cable carries all signals between the magnetic detector and the amplifier-power supply. The signals on the coaxial cable include the 5.4-kHz altitude signal, the 430/860-Hz IR signal, the 0.6-to 2.2-MHz resonance coil excitation, and the 52-kHz ignition signal. The 52-kHz ignition signal is monitored first by system BITE for its presence and, with the latest configuration, is checked approximately 5 minutes later (maximum warmup time) for its absence. If the signal is not present initially, the system timer stops and shows a detector head failure. Approximately 5 minutes later, the absence of the signal is checked. The absence of the signal indicates that the intensity of both helium lamps

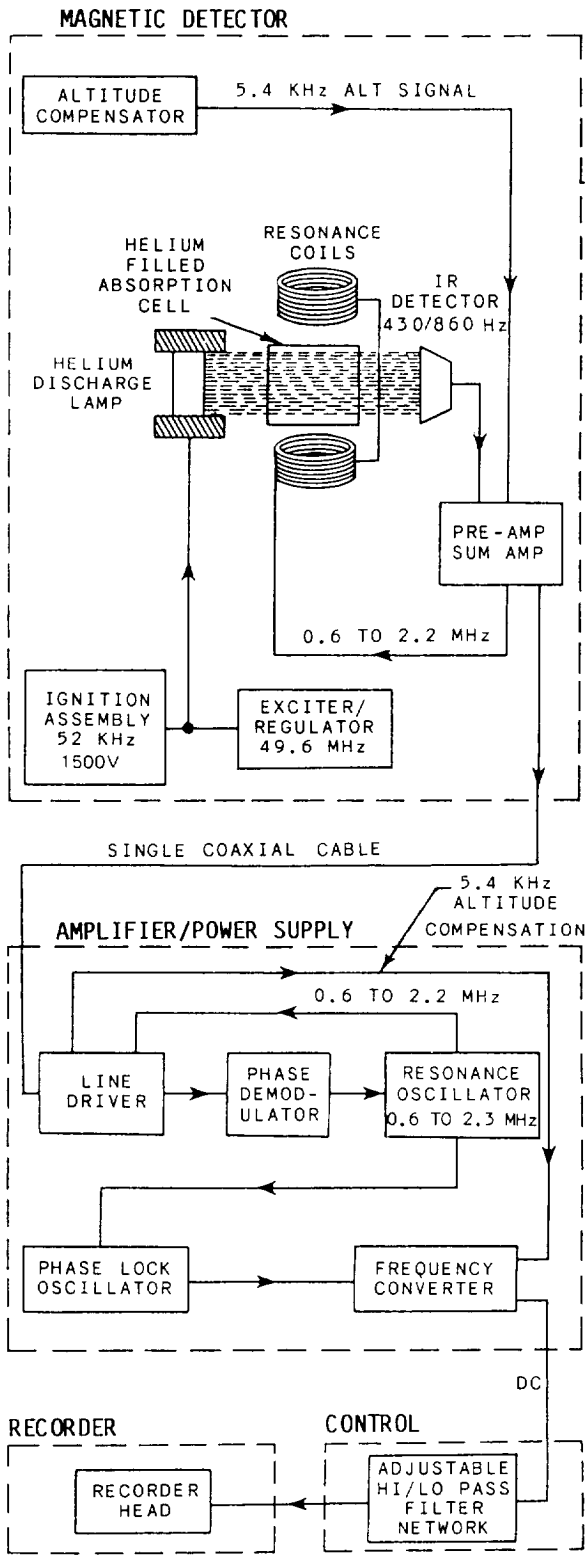


Figure 5-59.-Simplified block diagram of a MAD system.

was sufficient to cause the system to switch from ignition to the normal 49.6-MHz exciter signal. If this switchover did not occur, the system removes power to the detector head and illuminates the detector failure indicator. Bandpass filters at both ends of the coaxial cable route the frequencies to their respective circuits.

The IR detector error signal (anytime 430 Hz is present, this represents an error between the Larmor frequency and the resonance oscillator center frequency; 860 Hz indicates matched frequencies) is phase-detected in the phase demodulator to produce a variable positive or negative dc voltage. The variable dc is, in turn, used to change the resonance oscillator frequency. This closed loop action keeps the oscillator always at resonance as the earth's field strength changes or as an anomaly is detected. When neither 430 Hz nor 860 Hz is coming back from the magnetic detector, the system senses this and causes the resonance oscillator to sweep its entire range (0.6 MHz to 2.2 MHz). During the down sweep of the resonance oscillator, when 430/860 Hz is detected, the sweep is stopped and the loop is closed. The resonance oscillator output is routed to the resonance coils via the line driver and the preamplifier/summing amplifier. It is also applied to the phase lock oscillator assembly.

The purpose of the phase lock oscillator is to reproduce the resonance oscillator frequency, retain the magnetic anomaly, and eliminate the 430-Hz modulation signal. The oscillator is voltage controlled by a dc signal from the acquisition circuit until it locks on to the resonant frequency. After lock-on, the phase detector provides the control necessary for tracking the resonant frequency. This tracking is similar to the way the resonant oscillator tracks the Larmor frequency.

The frequency converter develops a variable dc signal proportional to the frequency shift of the phase lock oscillator. The frequency converter also generates a variable dc voltage proportional to the frequency shift of the 5.4-kHz altitude compensation signal input supplied by the magnetic detector unit. If the operator selects ALT COMP on the control unit, the two dc signals are combined by a summing network to compensate for magnetometer altitude change effects. Two driver-amplifiers provide a primary output to the detecting set control unit and an auxiliary output for test purposes.

The primary output is passed through a series of high- and low-pass filters in the control unit. The filters remove all extraneous frequencies and noise from the variable dc except the anomaly signal. The filtered dc output drives the pen of a recorder to produce a permanent record of the submarine anomaly.

MAINTENANCE

The maintenance philosophy for the AN/ASQ-81 and its components includes WRA turn-in to the local AIMD. Maintenance of the detector head is basically test and check if any capability at all. However, the amplifier, control box, and reeling machine can be repaired, depending on local repair capabilities. The intermediate maintenance manuals, NAVAIR 16-30ASQ81-1 and -2, contain the troubleshooting procedures and IPB.

MAGNETIC COMPENSATOR GROUP

Learning Objective: *Recognize the components and operating features of the AN/ASA-65 magnetic compensator group, including the three magnetic field terms associated with the compensation process.*

The AN/ASA-65 magnetic compensator group is used in conjunction with the AN/ASQ-81 MAD system to reduce the effects of unwanted magnetic disturbances (anomilies) during MAD subsystem operation. As previously mentioned, the MAD compensator generates magnetic fields in the MAD subsystem boom to nullify noise interference from unrelated sources.

MAGNETIC FIELD TERMS

Relative to MAD compensation theory, the word *term* refers to a magnetic field component. Permanent field terms are designated by a single capital letter such as *T, L,* or *V,* which stand for transverse, longitudinal, and vertical, respectively. These three axes, which are references to the three aircraft axes, remain fixed regardless of aircraft orientation with respect to the earth. Induced terms are designated by two capital letters, which may be any combination of *L, T,* or *V,* such as *LL, VL,* or *TT.* The first capital letter designates the inducing aircraft structure axis, and the second capital letter designates the resulting field axis as seen at the detector head. The inducing axis may be different from the resulting field axis at the detector head. Eddy-current terms are designated

by two lowercase letters, which may be any combination of the three basic components in the same manner as induced terms, such as *ll*, *vl*, or *tt*.

The total interfering field at the sensor can be resolved theoretically into 16 terms comprising three permanent, five induced, and eight eddy-current terms. The interference could be eliminated by generating an opposing field at the detector containing 16 terms, which cancel their interfering counterparts. In practice, not more than nine are required to compensate any aircraft satisfactorily because not all the induced and eddy-current terms are significant.

AN/ASA-65 COMPENSATOR COMPONENTS

AN/ASA-65 magnetic compensator group components involved in the magnetic compensation of the AN/ASQ-81 MAD system includes a few system component variations. The basic system contains a control-indicator unit, an electronic control amplifier, a magnetometer assembly, and compensation coils. However, the upgraded ASA-65(V)4/5 includes the addition of the compensator group adapter. As you read about the components, refer to figure 5-60 and observe the relationship between the various signals and components involved. Refer to figure 5-61 to see the physical appearance of the components comprising the AN/ASA-65 magnetic compensator group.

Control-Indicator Unit

The control unit contains all the controls and indicators required for system operation and all the elements necessary for term adjustment. The control unit provides the operator with a numeric indication of compensation potentiometer position.

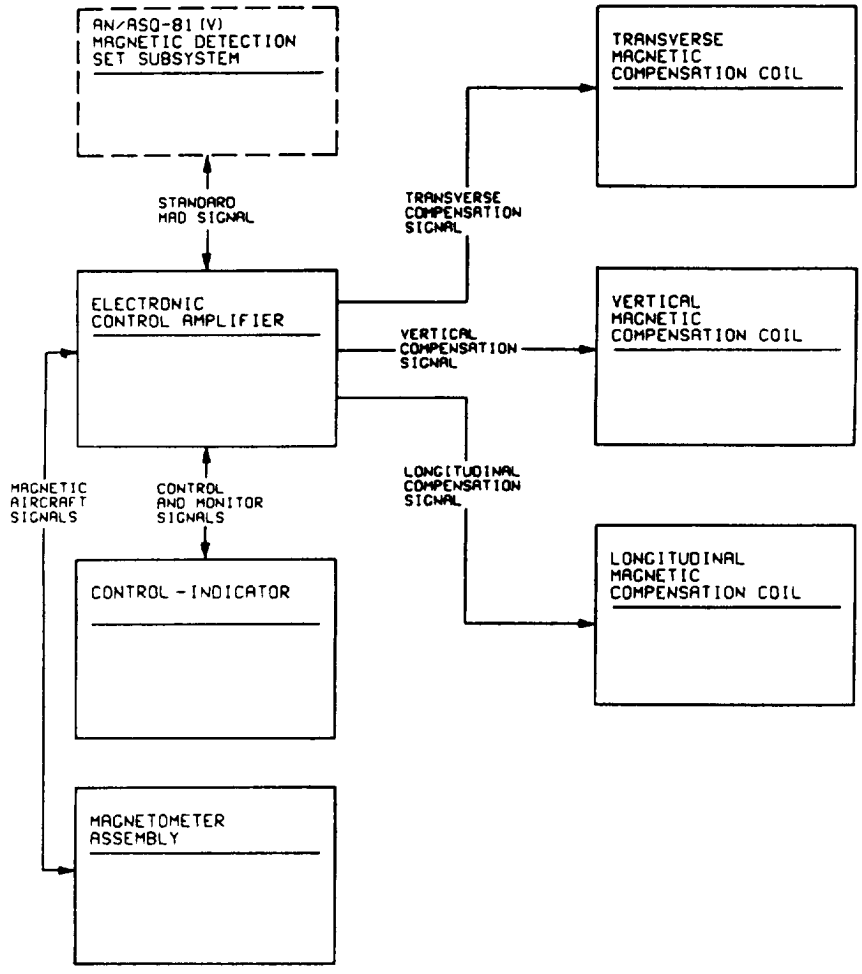


Figure 5-60.-Magnetic compensator group interface diagram.

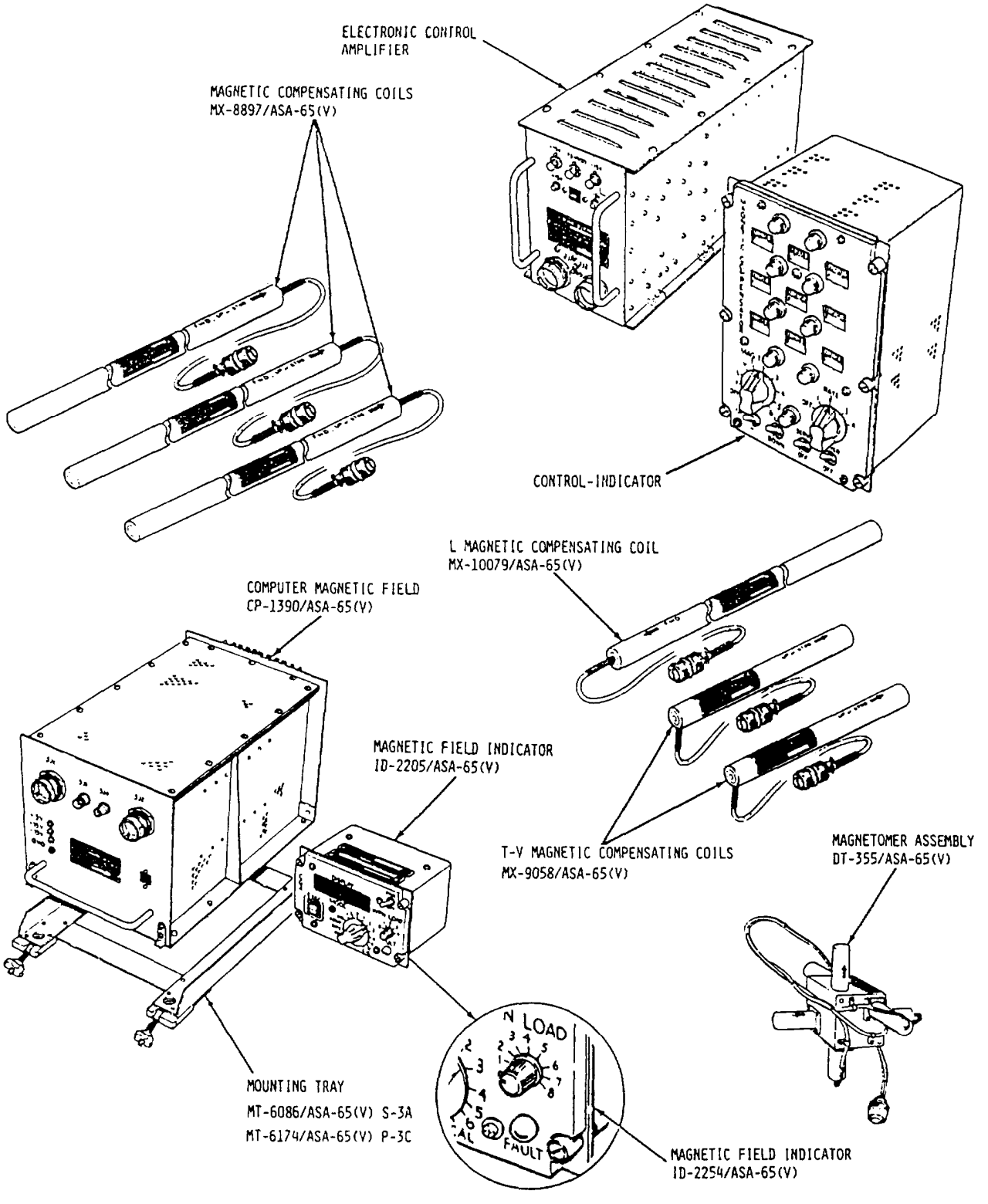


Figure 5-61.-Magnetic compensator groups AN/ASA-65(V)4 and 5.

Electronic Control Amplifier

The electronic control amplifier (ECA) processes standard magnetic anomaly detector (MAD) signals from the MAD subsystem, operator compensation adjustments, and maneuver signals from the magnetometer. The ECA provides compensation currents, which are sent to the MAD boom compensation coils. The ECA provides all necessary interconnections and receives all inputs to the system. It converts nine separate term adjustment outputs from the control-indicator to three current outputs, which energize coils of the cutput coil assembly.

Magnetometer Assembly

The magnetometer contains three coils oriented to sense magnetic strength in each of the basic longitudinal, transverse, and vertical aircraft axes. This results in three output signals, which are sent to the ECA as operating voltages for induced and eddy current compensation. In the upgraded ASA-65(V)4 and 5 configurations, signals also go to the magnetic field computer. The computer calculates the difference value of the terms necessary to reduce the maneuver-related interference to a minimum.

Transverse Magnetic Compensation Coil

The transverse coil generates a transverse magnetic field that opposes the aircraft-generated noise field for compensation. This field cancels or minimizes magnetic fields interfering with MAD operation. The transverse coil is located in the MAD boom.

Vertical Magnetic Compensation Coil

The vertical coil generates a vertical magnetic field that opposes the aircraft-generated noise field for compensation. This field cancels or minimizes magnetic fields interfering with MAD operation. The vertical coil is located in the MAD boom.

Longitudinal Magnetic Compensation Coil

The longitudinal coil generates a longitudinal magnetic field that opposes the aircraft-generated noise field for compensation. This field cancels or minimizes magnetic fields interfering with MAD operation. The longitudinal coil is located in the MAD boom.

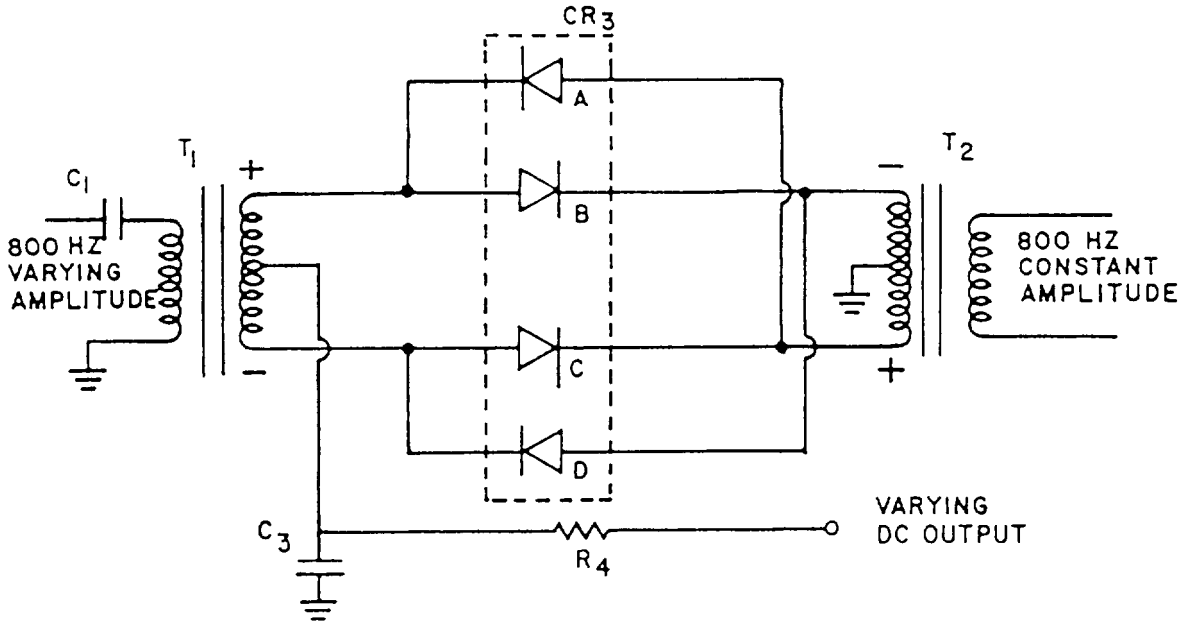


Figure 5-62.-Ring demodulator circuit.

ASA-65(V)4 and 5 Magnetic Field Computer

The addition of the compensator group adapter (CGA) completely computerizes the compensation calculation. Compensation is performed simultaneously on all nine terms by performing a maneuver pattern requiring only a few minutes. The computer provides the necessary interconnection between primary components of the CGA and receives 28 analog maneuver signals. These maneuver signals are digitized for data processing to provide LED drive voltages to the magnetic field indicator.

Magnetic Field Indicator

The indicator contains the controls and indicators necessary to operate the CGA to perform semiautomatic compensation. The unit also displays term difference figures, calibration voltages, and BITE codes. The MODE control selects the various operating conditions of the computer.

MAINTENANCE

The maintenance philosophy for the AN/ASA-65 includes WRA and SRA turn-in to AIMD. NAVAIR 16-30ASA65-1 and -2 are the MIMs for intermediate repair. The local AIMD ICRL will indicate repair capability for the specific components.

A good understanding of modulators and demodulators is necessary in troubleshooting the AN/ASA-65, especially a ring demodulator (fig. 5-62). To understand the operation of the balanced demodulator, assume the 800-Hz signals at T_1 and T_2 are equal in amplitude. With the coupling transformers polarized as shown, diode B will conduct during the first alteration of 800 Hz, and C will conduct during the second half cycle. The output developed across C_3 and R_4 filters out the signal and provides a constant dc output. However the input amplitude at T_1 is always varying, and therefore the output dc varies as the input amplitude does. Thus, the output is a varying dc representing the amplitude modulation of the 400 Hz at T_1. If either input signal is lost, the output will be zero.

The ring modulator, figure 5-63, provides a 400-Hz signal modulated with 0.1 to 0.5 Hz. Without the 0.1 to 0.5 Hz modulation signal, which represents the maneuver signal, diodes CR_3 and CR_4 conduct during one-half cycle and CR_5 and CR_6 during the other half cycle of the 400-Hz input. Due to the polarization of the transformers, no current flows through 2A5T1 primary. Once we supply the 0.1 to 0.5 Hz maneuver signal, it blocks the diodes and lets current flow through the primary of 2A5T1. This modulates the 400-Hz signal output. With the loss of either signal, there will be no output.

SUBMARINE ANOMALY DETECTOR (SAD)

Learning Objective: *Recognize the capability and purpose of the ASA-64/ ASA0-71 (SAD).*

SAD works in conjunction with MAD, and increases the capability to detect submarines

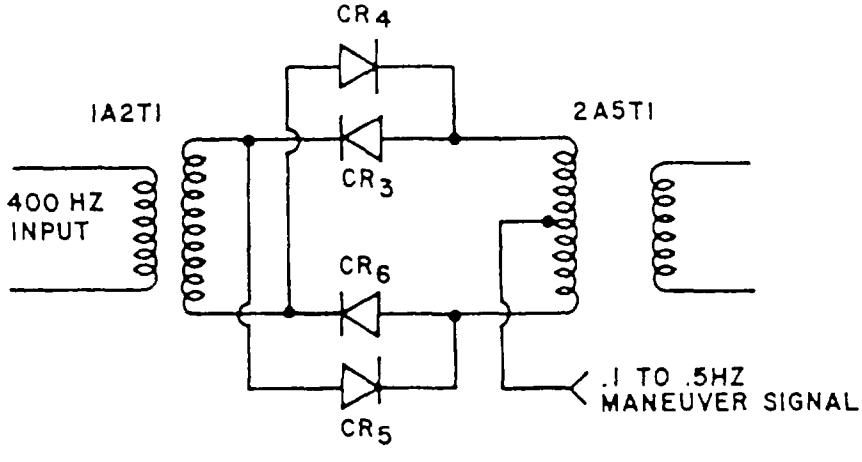


Figure 5-63.-Ring modulator.

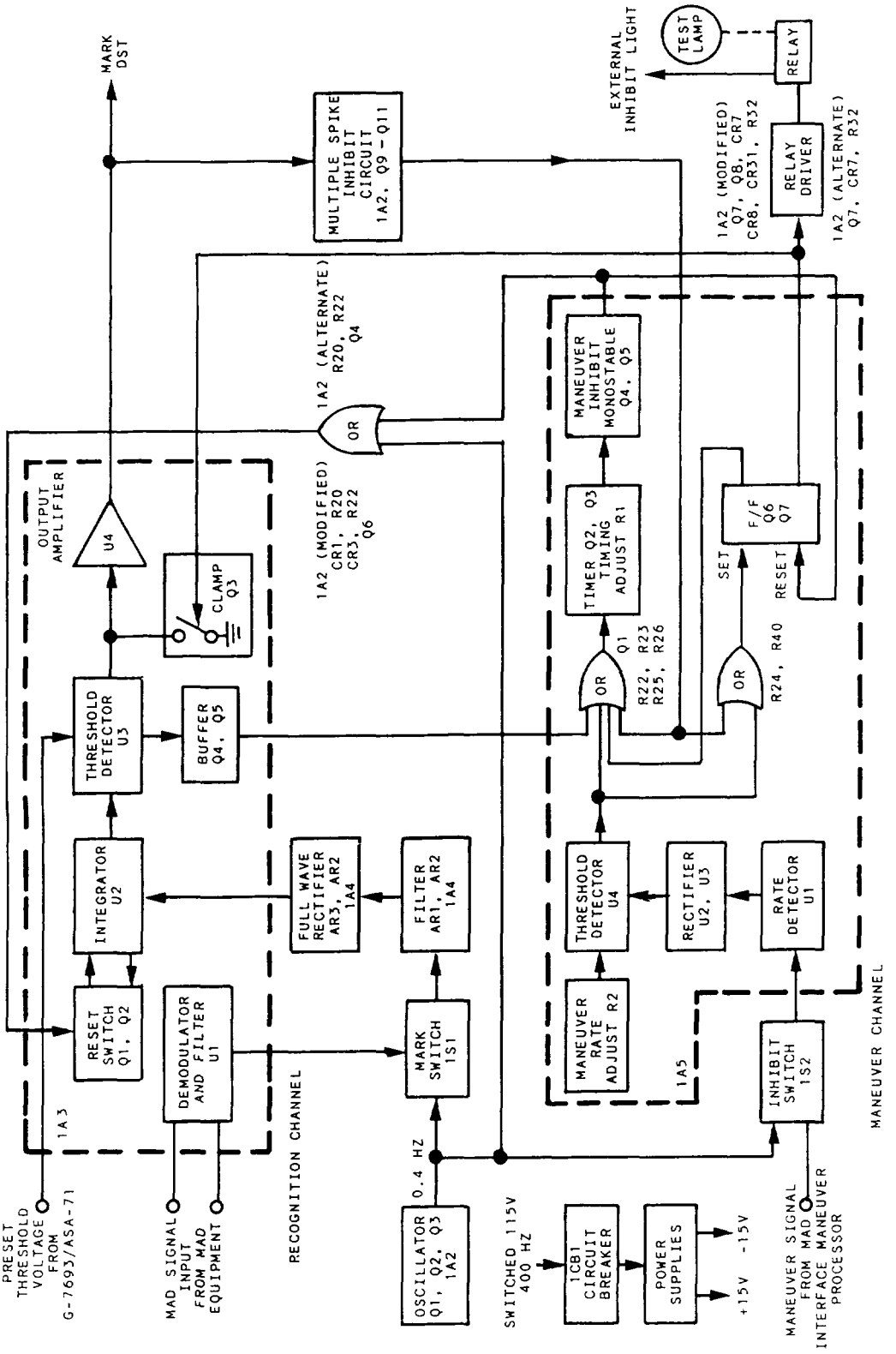


Figure 5-64.—ASA-64 system block diagram.

through evaluating MAD signals and separating submarine anomaly from magnetic noise. SAD also provides audio and visual indications that a submarine type of contact has been made. The ASA-64 SAD system block diagram (fig, 5-64) indicates some of the signal flow and shows some of the typical circuitry within the ASA-64. One of the outputs is a +15 volts dc to the threshold adjust on the C-7693/ASA-71, which then becomes the preset threshold input to the ASA-64 Threshold Detector U3. The system separates the submarine anomaly by selective filtering, full-wave rectification, short term integration, and correlation of aircraft maneuvers through the recognition and maneuver channels.

The MAD/SAD system interface is accomplished by the AN/ASA-71 selector control group. The group includes the C-7693/ASA-71 selector control panel (A302) and the MX-8109/ASA-71 selector control subassembly (A303). The A302 selector control panel provides controls for operating the A303 selector control subassembly. The A303 receives, processes, and distributes signals between the MAD/SAD system and the central repeater system, ICS, and ADP system. The A302 selects the signal to be routed through the A303, which is either the AUX, SAD, or MAD signal. Specific troubleshooting and maintenance procedures are contained in NAVAIR 16-30ASA64-1 and NAVAIR 16-30ASA71-1 MIMs.

CHAPTER 6

RADAR CIRCUITS

Chapter Objective: *After completing this chapter, you should be able to identify the operating requirements for radar synchronizers; recall the purpose of radar receivers and transmitters; recognize common types of radar indicators; and describe the operating principles and characteristics of an IFF system.*

Radar systems consist of a transmitter that sends out RF signals, a receiver that is located at the same site, and an indicator that gives a visual indication of echoes returned by the target. To accomplish this, many circuits are required. The circuits described in this chapter are those of a typical pulsed-type radar. The major components are synchronizers, transmitters, receivers, and indicators.

SYNCHRONIZERS

Learning Objective: *Identify the operating requirements for radar synchronizers, to include timers and markers.*

The purpose of a radar timer is to synchronize the sweep voltage or current for the indicator with the transmitter pulse. The specific function of the synchronizer is to produce the trigger pulse that starts the transmitter, the sweep circuits, range mark generators, blanking circuits, and gating circuits.

Either timing or control is the function of the majority of the circuits in radar. Circuits in a radar set accomplish one of these functions by producing a variety of voltage waveforms, such as square waves, sawtooth waves, trapezoidal waves, rectangular waves, brief rectangular pulses, and sharp peaks.

In sound systems and in radio, electronic circuits operate within the limits for which they are designed. In radar timing circuits, electronic circuits are often violently overdriven, frequently operating at points that range from well into the base current region to far beyond cutoff. Although all of these circuits are broadly classified as timing circuits, the specific function of any individual circuit might be timing, wave shaping, or wave generating.

Radar systems may be classified as either self-synchronized systems or externally synchronized systems. In a self-synchronized system, the timing trigger pulses are obtained from the transmitter. In an externally synchronized system, the timing trigger pulses are obtained from a master oscillator, which is usually external to the transmitter. The master oscillator may be a sine-wave oscillator, a stable (free-running) multivibrator, or a blocking oscillator.

When a blocking oscillator is used as a master oscillator, the timing trigger pulses are usually obtained directly from the oscillator. When a sine-wave oscillator or an astable multivibrator is used as a master oscillator, pulse-shaping circuits are required to form the necessary timing trigger pulses.

In a self-synchronized radar system, the repetition rate of the timing trigger pulses is determined by the repetition rate of the modulator (or transmitter) pulses. In an externally synchronized radar system, the repetition rate of the timing trigger pulses from the master oscillator determines the pulse repetition rate of the transmitter.

An indicator, such as a cathode-ray tube (CRT), is associated with every radar system. The indicator presents target data (range, bearing, and elevation) in visual form so that the target may be located. Trigger pulses from the timer (synchronizer) are frequently used to produce gate

pulses. When applied to the indicator, these gate pulses perform the following functions:

1. Initiate and time the duration of the indicator sweep voltage.
2. Intensify the CRT electron beam during the sweep period so that the echo pulses may be displayed.
3. Gate a range mark or range marker generator so that range marker signals may be superimposed on the indicator presentation. (The terms *marks* and *markers* are normally interchangeable.)

Figure 6-1 shows the time relationship of waveforms in a typical radar set. The timing

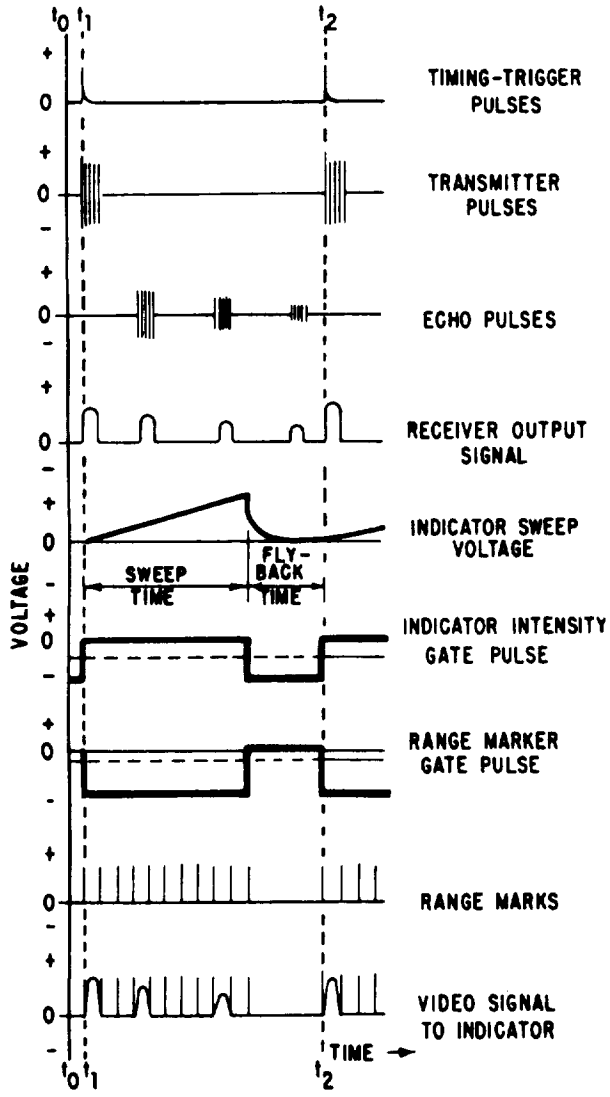


Figure 6-1.-Time relationship of waveforms.

trigger pulses are applied to both the transmitter and the indicator. When a trigger pulse is applied to the transmitter, a short burst, or pulse, of RF energy is generated. This energy is conducted along a transmission line to the radar antenna, from which it is radiated into space. If the transmitter energy strikes one or more reflecting targets in its path, some of the transmitted energy is reflected back to the antenna. Echo pulses from three reflecting targets at different ranges are illustrated in the part of figure 6-1 labeled "echo pulses." The corresponding receiver output signal is also shown.

The initial and final pulses in the receiver output signal are caused by the energy that leaks through the transmit-receive (TR) device when a pulse is being transmitted.

The indicator sweep voltage (fig. 6-1) is initiated at the same time that the transmitter is triggered. By delaying the timing trigger pulse fed to the indicator sweep circuit, it is possible to initiate the indicator sweep after a pulse is transmitted. (It is also possible to initiate the indicator sweep before a pulse is transmitted.)

Note in figure 6-1 that the positive indicator-intensity gate pulse (applied to the cathode-ray tube control grid) occurs during the indicator sweep time. As a result, the cathode-ray tube trace occurs only during the sweep time and is eliminated during the flyback (retrace) time. The negative range marker gate pulse also occurs during the indicator sweep time. This negative gate pulse is applied to a range marker generator, which produces a series of range marks.

The range marks are equally spaced and last only for the duration of the range marker gate pulse. When the range marks are combined (mixed) with the receiver output signal, the resulting video signal applied to the indicator may appear as shown in figure 6-1.

BASIC REQUIREMENTS

The basic timing circuit should meet three basic requirements:

1. It must be free running (astable)—since the timer is the heart of the radar, it must establish the zero time reference and the pulse-repetition frequency (PRF).
2. It must be stable in frequency (PRF)—the PRF, or its reciprocal, the PRT, must not change between pulses for accurate ranging.
3. The frequency must be variable in steps for the radar to operate on different ranges.

There are three basic circuits that can meet the above three requirements. These circuits are (A) the sine-wave oscillator, (B) the master-trigger multivibrator, and (C) the single-swing blocking oscillator.

Figure 6-2 shows the block diagrams and waveforms of these three timers used in externally synchronized radar systems. Note that in each case equally spaced timing trigger pulses are produced. The repetition rate of each series of timing trigger pulses is determined by the operating frequency of the associated master oscillator.

Sine-Wave Timer

In the sine-wave timer (fig. 6-2, view A), a sine-wave oscillator is used for the basic timing device (master oscillator). The oscillator may be a Wien bridge oscillator or a phase-shift oscillator. The oscillator output is applied to both an overdrive amplifier and the radar indicator. The sine waves applied to the overdriven amplifier are converted into square waves. The square waves, in turn, are converted into positive and negative trigger pulses by means of a short-time-constant RC.

By use of a limiter, either the positive or negative trigger pulses from the RC differentiator can be removed, leaving trigger pulses of only one polarity. For example, the limiter shown in figure 6-2, view A, is a negative-lobe limiter. That is, the limiter removes the negative trigger pulses and sends only positive trigger pulses to the radar transmitter.

One disadvantage of sine-wave timers is the large number of pulse-shaping circuits required to produce the necessary timing trigger pulses. Therefore, these timers are not normally found in modern weapons systems radar.

Multivibrator Timer

In a multivibrator timer, the master oscillator generally consists of an astable multivibrator. If the multivibrator is asymmetrical, as in figure 6-2, view B, it generates rectangular waves. If the multivibrator is symmetrical, it generates square waves. In either case, the timing trigger pulses are equally spaced after the limiter removes undesired positive or negative lobes.

The output of the astable multivibrator consists of two rectangular waves. (Remember, there are two transistors in an astable multivibrator. The two collector output voltages are equal in amplitude, but 180 degrees out of phase.) One set of rectangular pulses is applied to the RC

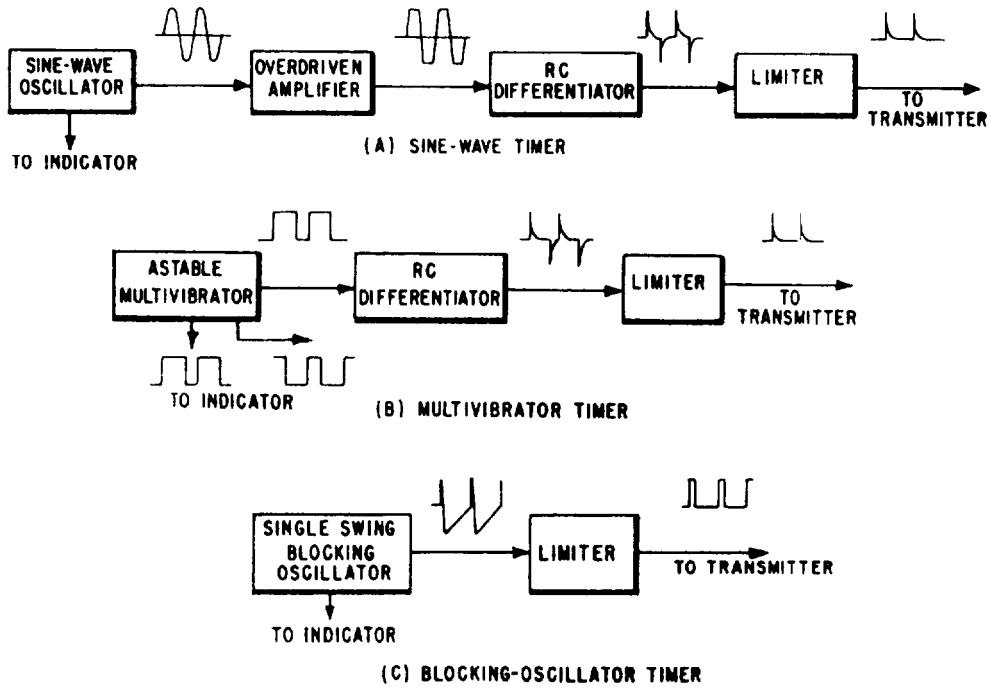


Figure 6-2.-External timer block diagrams.

differentiator and converted into positive and negative trigger pulses.

As in the sine-wave timer, the negative trigger pulses can be removed by means of a negative-lobe limiter. Both sets of rectangular pulses from the astable multivibrator are applied to the indicator for the following purposes:

1. One set of pulses is used to intensify the cathode-ray tube electron beam for the duration of the sweep.
2. The other set of pulses is used to gate the range marker generator. As will be shown later, rectangular pulses can also be used to produce range steps.

Blocking-Oscillator Timer

In the blocking-oscillator timer (fig. 6-2, view C), a free-running, single-swing blocking oscillator is generally used as the master oscillator. The advantage of the single-swing, blocking oscillator is that it generates sharp trigger pulses directly. Timing trigger pulses of only one polarity are obtained by use of a limiter.

Gate pulses for the indicator circuits are produced by applying the output of the blocking oscillator to a one-shot multivibrator or a phantastron. Crystal-controlled oscillators may be used when very stable operation is required at a particular frequency.

RANGE MARKERS

The accuracy of target-range data provided by a radar varies with the use of the radar. For example, a weapons systems radar operating in its search (or map) mode needs to be accurate only within a few percent of its maximum range. However, an intercept type radar operating in its track mode must supply range data that is accurate within a few yards.

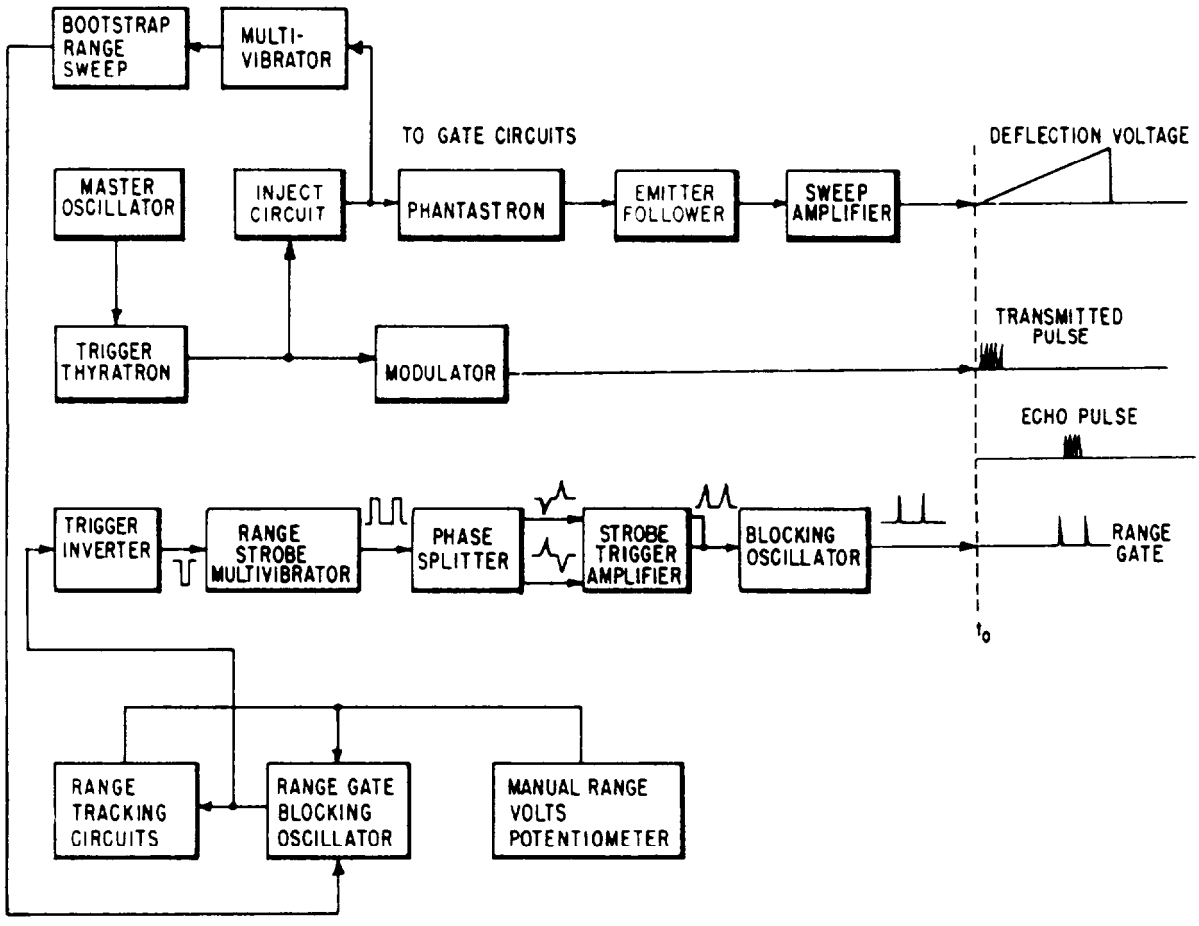


Figure 6-3.-Synchronizer with range gate generator.

In some applications of radar, the indicator sweep is calibrated by a transparent overlay with an engraved scale. This allows the operator to estimate the range of targets. In other applications, electronic range marks are supplied to the indicator. They appear as horizontal lines on a B-scope and as concentric circles on a PPI-scope. The distance between range marks is generally determined by the type of equipment and its mode of operation.

In a weapons systems radar that requires extremely accurate target-range data, a movable range marker may be used. The range marker is obtained from a range marker generator and may be a movable range gate or range step. When a PPI-scope is used, a range circle of adjustable diameter is used to measure range accurately. In some cases, movement of the range marker is done by turning a calibrated control from which you can obtain range readings. In other cases, the range marker may be used as a range gate for automatic range tracking. In this case there may be no direct range readout, or the readout may be a voltmeter calibrated in range and to which range voltage, equivalent to range marker position, is applied.

This discussion describes the operation of three types of range markers (generators): the range gate generator; the range marker generator; and the range step generator. The range gate generator, used in conjunction with a blocking oscillator, generates a movable range gate. The range marker generator and the range step generator, used in conjunction with an astable multivibrator, generate fixed range marks and a movable range step, respectively.

Range Gate Generator

Figure 6-3 shows a simplified block diagram of a typical radar synchronizer that includes a range gate generator. The indicator is a B-scope with the range deflection voltages applied to the vertical plates. Scopes are discussed later in this chapter.

The PRF is controlled by a master oscillator, or multivibrator, whose output is coupled to a thyratron trigger. The output of the thyratron trigger is used to trigger the radar modulator and the B-scope sweep circuits, thus starting the transmitting pulse and the range sweep at the same instant.

The phantastron in the sweep circuits is a timing circuit that supplies a sweep sawtooth to the sweep amplifier. The width of the gate and sawtooth is dependent on the range selected by the operator.

The range gate circuit receives its input pulse from the trigger thyratron and generates a delayed range gate pulse. The delay of this pulse from to is dependent on the position of the target in range when tracking, or on the manual positioning of the range volts potentiometer by the operator when in the search mode. The range gate triggers the range strobe multivibrator, whose output is amplified and sent to the blocking oscillator. This oscillator sharpens the pulses, as shown in figure 6-3, The range gate selects the target to be tracked and, when in track mode, brightens the trace or brackets the target (depending on the system) to indicate which target is being tracked.

Range Marker Generator

Figure 6-4 is a block diagram of a typical range marker generator. This generator consists of a ringing oscillator Q1611-Q1612, an emitter follower Q1613, a countdown multi vibrator Q1616-Q1617, and a pulse-forming amplifier Q1614. Generation of the marks starts at the ringing oscillator, which is excited into operation by incoming trigger pulses. Once in operation, it produces a sinusoidal output, which is synchronized to the trigger pulses. This sinusoidal output is then applied to the emitter follower. This provides interstage buffering by isolating the ringing oscillator from the countdown multivibrator. The output coupling circuit of the emitter follower shifts the average output level to zero (ground) and clips the negative going portions of the signal. This allows only the positive half of each sine wave to reach the countdown multivibrator. The countdown multivibrator receives a high-frequency positive trigger corresponding to a fixed interval. This input drives the countdown multivibrator to develop a negative pulse train. The period of the pulse train is controlled by the range marks select switch. This negative output is applied to the pulse-forming

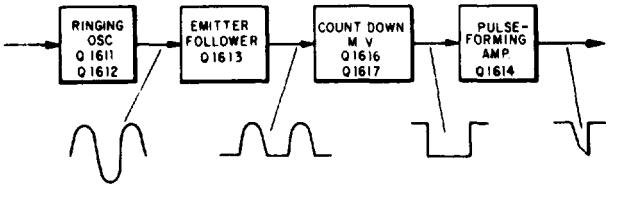


Figure 6-4.-Range marker generator.

amplifier, where it is reshaped and passed on to a marker mixer.

The output of a range marker generator can be applied directly to one of the deflection plates on an A-scope. In this case, range marker pulses appear simultaneously with the radar echo signals, and permit estimation of target range. In B-scope and PPI-scope applications, the output of the range marker generator is applied to a video mixer. In this case, radar echo signals are combined with marker signals before being applied to the grid of the CRT.

Range Step Generator

Figure 6-5 shows the schematic diagram and waveforms of a typical range step generator. The range step generator consists of a sawtooth voltage generator Q1, a negative clipper CR1, and a limiting amplifier. Diode CR1 is frequently referred to as a pickoff diode. The position of the range step along the indicator time base is controlled by potentiometer R3. When the range step coincides with the leading edge of a target echo pulse, the target range can be read directly from a calibrated range dial associated with R3.

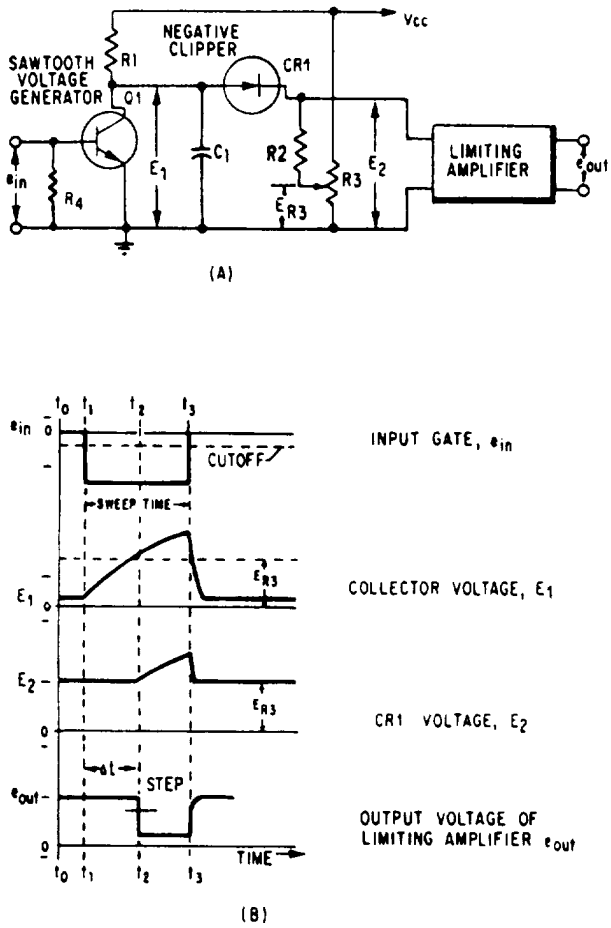


Figure 6-5.-Range step generator with pickoff diode. (A) Schematic diagram; (B) waveforms.

Between times t_0 and t_1 (fig. 6-5, view B), the base of transistor Q1 is at ground potential (zero volts). As a result, Q1 conducts and the Q1 collection voltage (E_1) equals Q1 collector-supply voltage (Vcc) minus the voltage drop across the load resistor R1. The horizontal dashed line across the E_1 waveform indicates the E_{R3} voltage (fig. 6-5, view A) at the adjustable tap of potentiometer R3. Since E_1 is less than E_{R3} between times to and t the anode of the negative clipper CR1 is less positive than the CR1 cathode, and CR1 does not conduct. Hence the CR1 cathode voltage (E_2) equals E_{R3}, the voltage at the R3 tap.

Between times t_1 and t_3, the base of transistor Q1 is driven below cutoff. As a result, Q1 ceases to draw collector current. When no collector current flows in Q1, the capacitor C 1 changes through the Q1 load resistor R 1, and the collector voltage of Q1 (E_1) rises exponentially toward the Q1 collector-supply voltage (Vcc). At time t_2, E_1 exceeds E_{R3}, and diode CR1 conducts. If the CR1 anode resistance is small, the CR1 cathode voltage (E_2) practically equals E_1 between times t_2 and t_3.

Following time t_3 the base of Q1 returns to ground potential, and Q1 again conducts. As a result, capacitor C1 discharges through Q1, and the Q1 collector voltage decays exponentially toward its initial value. As soon as E_1 becomes less than E_{R3}, CR1 no longer conducts, and the CR1 cathode voltage again equals E_{R3}.

When the E_2 waveform is amplified and limited by the limiting amplifier, the amplifier output-voltage (e_{out}) waveform appears, as shown in figure 6-5, view B. Note that a nearly vertical edge (step) appears in the e_{out} waveform the instant CR1 begins to conduct (time t_2).

By varying the setting of the R3 tap, you can vary the instant at which CR1 conducts. You can therefore control the position of the range step by adjusting the setting of R3. If a linear relationship is to be established between the delay of the step (t) and the voltage at the R3 tap (E_{R3}), the Q1 sawtooth collector voltage must be linear.

The e_{out} waveform is applied to the vertical-deflection plates of a cathode-ray tube. Only the portion of the e_{out} waveform that occurs between times t_1 and t_3 is displayed on the CRT screen.

Remember, the indicator trace is blanked out during the flyback (retrace) time.

TRANSMITTERS

Learning Objective: *Recall the purpose of radar transmitters to include modulators ands witching devices.*

The purpose of a radar transmitter is to develop high-power, high-frequency pulses of RF energy to be radiated into space by the antenna system. Our discussion will be limited to transmitting devices that are used in fire control radar. Among these are magnetrons, klystrons, traveling wave tubes, and gallium arsenide (Gunn) oscillators. The construction, operating characteristics, and limitations of these transmitters are covered in Navy Electricity and Electronics Training Series (NEETS), Module 11, *Microwave Principles.*

Basically, a transmitter is an RF oscillator, which is turned on and off by a signal received from a modulator. The oscillator is not normally controlled directly by the signal from the timer. Instead, the timer triggers the modulator, which, in turn, switches the transmitter on and off.

RADAR MODULATORS

The radar modulator controls the radar pulsewidth by use of a rectangular dc pulse (modulator pulse) of the required duration and amplitude. The peak power of the transmitted (RF) pulse depends on the amplitude of the modulator pulse.

Figure 6-6 shows the waveforms of the trigger pulse applied by the timer to the modulator, the modulator pulse applied to the radar transmitter, and the transmitted RF pulse.

Note the following:

1. The modulator pulse is applied to the transmitter the instant the modulator receives the trigger pulse from the timer.

2. The modulator pulse is flat on top.

3. The modulator pulse has very steep leading and trailing edges.

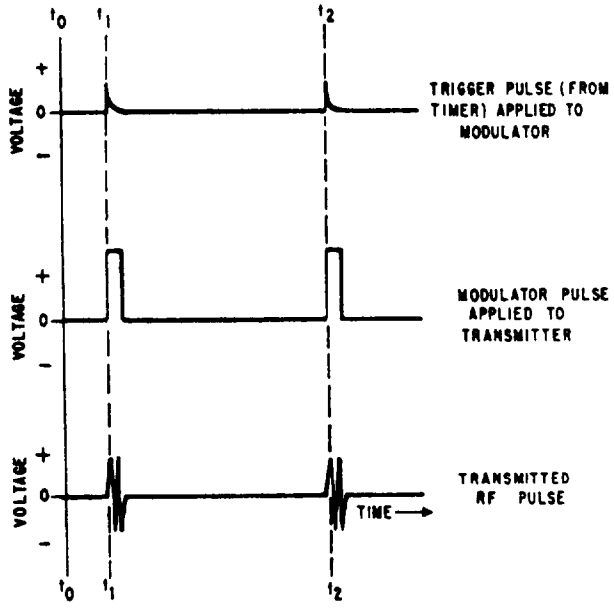


Figure 6-6.-Radar pulse waveforms.

Remember, for accurate determination of target range, the timing circuit must be triggered the instant the leading edge of the transmitted RF pulse leaves the transmitter. Thus, the trigger pulse that controls the operation of the modulator also synchronizes the CRT sweep circuits and target range.

There are two types of modulators—the line-pulsing modulator and the drive-hard-tube modulator. The line-pulsed modulator has replaced the drive-hard-tube modulator for most radar uses. The line-pulsing modulator stores energy and forms pulses in the same circuit element. This element is usually the pulse-forming network. The drive-hard-tube modulator forms the pulse in the driver. The pulse is then amplified and applied to the modulator. The reasons for the replacement are that the drive-hard-tube modulator had lower efficiency, its circuits were more complex, higher power supply voltage was required, and it was more sensitive to voltage changes.

The line-pulsed modulator is easier to maintain because of its less complex circuitry and, for a given amount of power output, it is more compact and light. Because it is the modulator used most in aviation radar, it is the one we will discuss.

Figure 6-7 shows the components of a basic radar modulator. The components of the radar modulator are as follows:

1. A power supply.

2. A storage element (a circuit element or network for storing energy).

3. A charging impedance (to control the charge time of the storage element and to prevent short-circuiting of the power supply during the modulator pulse).

4. A modulator switch (to discharge the energy stored by the storage element through the transmitter oscillator during the modulator pulse).

Figure 6-7, view A, shows the modulator switch open and the storage element charging. With the modulator switch open, the transmitter produces no power output, but the storage element stores a large amount of energy. Figure 6-7, view B, shows the modulator switch closed and the storage element discharging through the transmitter. The energy stored by the storage element is released in the form of a high-power, dc modulator pulse. The transmitter converts the dc modulator pulse to an RF pulse, which is radiated into space by the radar antenna. Thus, the modulator switch is closed for the duration of a transmitted RF pulse, but is open between pulses.

Many different kinds of components are used in radar modulators. The power supply generally produces a high-voltage output, either alternating or direct current. The charging impedance may be a resistor or an inductor. The storage element is generally a capacitor, an artificial transmission line, or a pulse-forming network. The modulator switch is usually an electron tube.

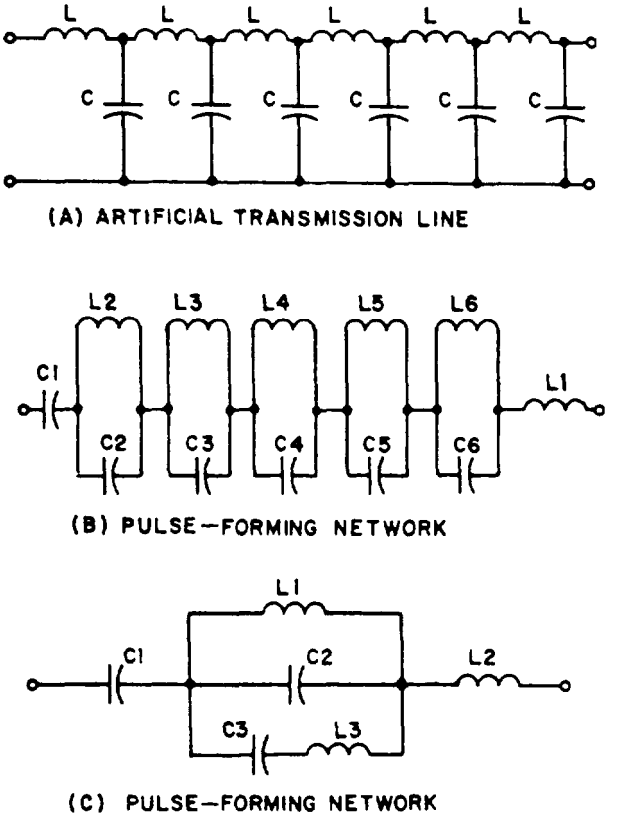


Figure 6-8.-Types of modulator storage elements.

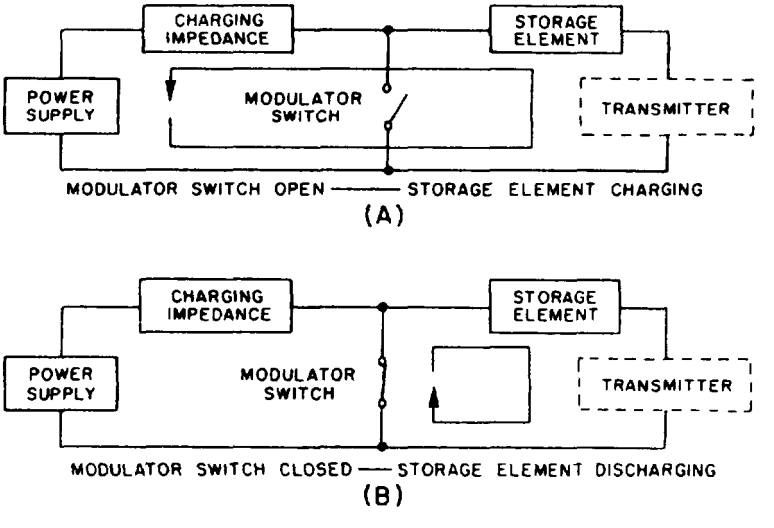


Figure 6-7.-Basic radar modulator block diagram. (A) Modulator switch open; (B) modulator switch closed.

Modulator Storage Elements

Capacitor storage elements are used only in modulators that have a dc power supply and an electron-tube modulator switch. The capacitor storage element is charged to a high voltage by the dc power supply, and releases only a small part of its stored energy to the transmitter. The electron-tube modulator switch controls the charge and discharge of the capacitor storage element.

The artificial transmission line storage element (fig. 6-8, view A) consists of identical capacitors (C) and inductors (L), arranged to simulate sections of a transmission line. The purposes of the artificial transmission line are to store energy when the modulator switch is open (between transmitted RF pulses) and to discharge and form a rectangular dc pulse (modulator pulse) of the required duration when the modulator switch is closed.

The duration of the modulator pulse depends on the values of inductance and capacitance in each LC section of the artificial transmission line (fig. 6-8, view A) and the number of LC sections used. Other arrangements of capacitors and inductors (pulse-forming networks) are very similar in operation to artificial transmission lines.

Capacitor

Figure 6-9, view A, is the schematic diagram of a modulator that uses a single capacitor (C1) as its storage element. The charge and discharge of C1 is controlled by transistor Q1, which is a switching transistor normally held below cutoff by a negative dc bias applied to its base.

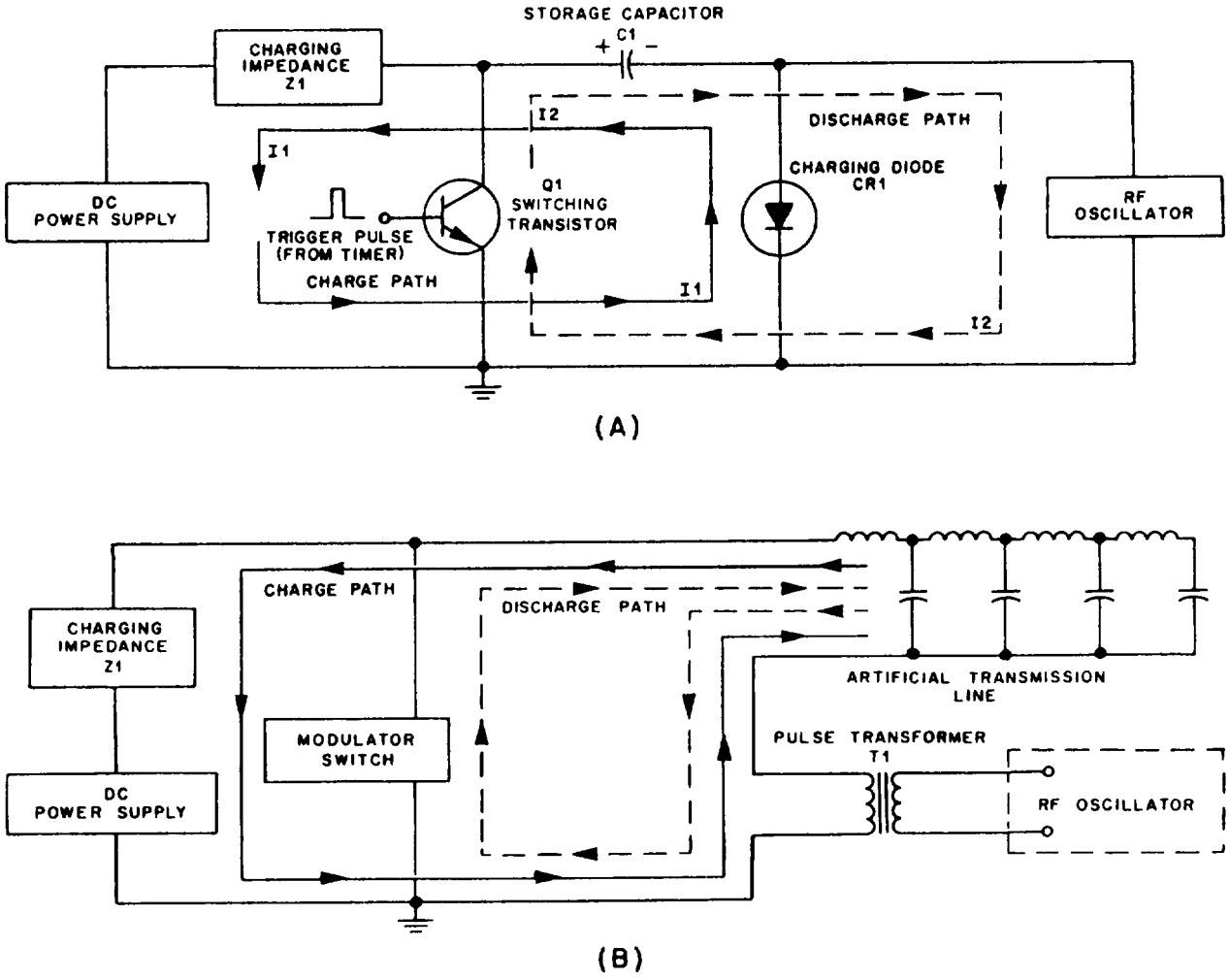


Figure 6-9.-Applications of basic modulator elements. (A) Single capacitor; (B) artificial transmission line.

When a positive trigger pulse (from the radar timer) is applied to the base of Q1, Q1 conducts for the duration of the trigger pulse. When Q1 is cut off, storage capacitor C1 charges through the series circuit, consisting of the dc power supply, charging impedance Z1, and charging diode CR1. The low voltage across CR1 effectively prevents the RF oscillator from operating.

When a positive trigger pulse is applied to the base of Q1, Q1 suddenly conducts, and C1 discharges. The discharge path (fig. 6-9, view A) is a series circuit consisting of switching transistor Q1 and the RF oscillator. Note that discharge current I2 is opposite in direction to charge current I1. Since charging diodes can conduct in only one direction (from cathode to anode), CR2 remains cut off during the modulator pulse. Thus, discharge current I2 flows through the RF oscillator, and an RF pulse is generated by the oscillator.

During the modulator pulse, storage capacitor C1 discharges, and the C1 voltage decreases. Since the C1 voltage (modulator pulse) is applied to the RF oscillator, the frequency of the oscillator changes if there is any significant change in the C1 voltage. To keep the C1 voltage practically constant, C1 must have a large capacitance. Thus, only a small fraction of the charge is removed from C1 during the modulator pulse, and the C1 voltage remains practically constant.

The switching transistor starts and stops the modulator pulse. Thus, the width (duration) of the modulator pulse depends on the width of the trigger pulse applied to the base of Q1. The pulse repetition rate depends on the rate at which trigger pulses are applied to the base of Q1. To obtain a modulator pulse with nearly vertical sides (a dc rectangular pulse), the trigger pulse must also have nearly vertical sides. Modulators that use capacitor storage elements and switching transistors as switches have the following advantages:

1. The pulse repetition rates can be varied over relatively wide limits.
2. The pulsewidth can also be varied over relatively wide limits.

Artificial Transmission Line

Figure 6-9, view B, shows a radar modulator that uses an artificial transmission line as its storage element. A switch (modulator switch) controls the pulse repetition rate. When the modulator switch is open (between modulator pulses), the transmission line charges.

The charge path includes the primary of pulse transformer T1, the dc power supply, and charging impedance Z1. When the modulator switch is closed, the transmission line discharges through the series circuit, consisting of the modulator switch and the primary of pulse transformer T1.

The artificial transmission line is effectively an open circuit at its output end. Thus, when the voltage wave reaches the output end of the line, it is reflected. As the reflected wave propagates from the output end toward the input end of the line, it completely discharges each section of the line. When the reflected wave reaches the input end of the line, the line is completely discharged, and the modulator pulse ceases abruptly. If the oscillator and pulse transformer circuit impedance is properly matched to the line impedance, then the voltage pulse that appears across the T1 primary is one-half the voltage to which the line was charged initially.

The width of the pulse generated by an artificial transmission line depends on the time required for a voltage wave to travel from the input end to the output end of the line and back. Thus, the pulsewidth depends on the velocity of propagation along the line (determined by the inductance and capacitance of each section of the line) and the number of line sections (the length of the line).

Pulse-Forming Networks

A pulse-forming network is similar to an artificial transmission line because it stores energy between pulses and produces an almost rectangular pulse. The pulse-forming network (fig. 6-8, view B) consists of inductors and capacitors arranged so they approximate the behavior of an artificial transmission line.

Each capacitor in the artificial transmission line (fig. 6-8, view A) must carry the high voltage required for the modulator pulse. Since each capacitor must be insulated for this high voltage, an artificial transmission line consisting of many sections is bulky and cumbersome.

The pulse-forming network (fig. 6-8, view B) can carry high voltage, but it does not require bulky insulation on all of its capacitors. Only series capacitor C1 must be insulated for high voltage. Since the other capacitors are in parallel with the corresponding inductors, the modulator pulse voltage divides nearly equally among them.

Thus, except for C1, the elements of the pulse-forming net work are relatively small.

Pulse-forming networks are often insulated by immersing each circuit element in oil. The network is usually enclosed in a metal box on which the pulse length, characteristic impedance, and safe operating voltage of the network are marked. If one element in such a network fails, the entire network must be replaced.

SWITCHING DEVICES

The voltage stored in a storage-element capacitor, artificial transmission line, or pulse-forming network must be discharged through a switching device. The switching device conducts for the duration of the modulator pulse, and is open-circuited between pulses. Thus, the modulator switch must perform the following functions:

1. Close suddenly and reach full conduction in a fraction of a microsecond.
2. Conduct large currents (tens or hundreds of amperes) and withstand large voltages (thousands of volts).
3. Cease conducting (become an open circuit) with the same speed that it starts to conduct.
4. Consume only a fraction of the power that passes through it.

These requirements are met best by the thyratron tube (gas-filled).

The thyratron, normally held below cutoff by a negative grid voltage, conducts when a positive trigger pulse is applied to its grid. Once fired, the thyratron continues to conduct as long as the storage element (artificial transmission line or pulse-forming network) is discharging.

During discharge of the storage element, the gas in the thyratron is highly ionized. While the storage element discharges, the plate-to-cathode resistance of the thyratron is practically zero. When the storage element is completely discharged, current ceases to flow through the thyratron and the gases become deionized. Thus, the negative grid bias regains control, and the thyratron is cut off (the modulator switch opens).

Most radar modulators use a high-voltage, dc power supply. Typical dc power supplies for radar modulators use a half-wave rectifier, a full-wave rectifier, or a bridge rectifier.

The modulator charging impedance (fig. 6-10, view A) prevents the dc power supply from becoming short-circuited when the modulator

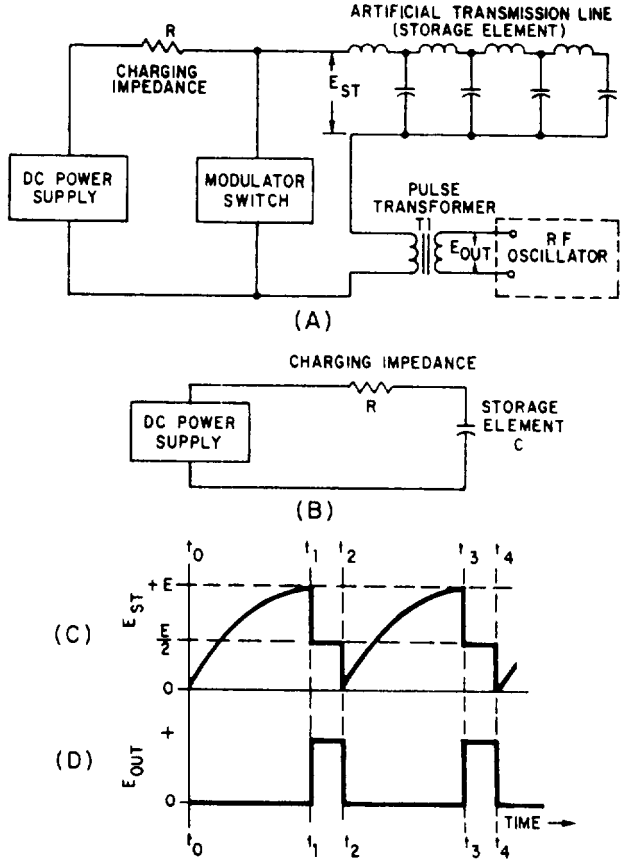


Figure 6-10.-Radar modulator with resistance charging. (A) Schematic diagram; (B) equivalent charge circuit; (C) storage element voltage (E_{st}) waveform; (D) T_1 secondary voltage (E_{out}) waveform.

switch closes. When the modulator switch is open, the charging impedance also controls the rate at which the storage element charges. When the charging impedance is large, the storage element charges slowly. When the charging impedance is small, the storage element charges rapidly.

Many different kinds of charging impedance and charging circuits are used in radar modulators. The type of charging impedance and charging circuit used depends on the following:

1. The type of power supply (ac or dc)
2. The type of storage element
3. The modulator pulse voltage required
4. The pulse repetition rate
5. The frequency of the available ac supply voltage

Most radar modulators charge very slowly compared with the rate at which they discharge.

(The interval between modulator pulses is much longer than the pulsewidth.) Because the charging current is relatively small and changes very slowly, inductances in a modulator storage element have negligible effect on charging. Thus, all modulator storage elements can be represented as a capacitor during their charging interval, as in figure 6-10, view B.

Resistance Charging

In figure 6-10, view B (the equivalent charging circuit of the radar modulator), note that a capacitor (C) represents the storage element (artificial transmission line) of the modulator, A resistor (R) represents the charging impedance. When the modulator switch is open, the storage element charges, along a typical RC charge curve, to a maximum voltage—E (time interval $t_1 - t_1$, fig. 6-10, view C).

When the modulator switch is closed (time t_1), the storage-element voltage (E_{st}) decreases to E/2. (Remember, if an artificial transmission line or pulse-forming network is charged to a maximum volt age, E, and a matching impedance load is suddenly connected across the line, the line voltage decreases instantly to E/2). Voltage E_{st} remains at E/2 for the duration of the modulator pulse (time interval $t_1 - t_2$). At the end of the modulator pulse (time t_2), voltage E_{st} suddenly decreases to zero. Shortly afterwards, the modulator switch opens, and a new charging cycle begins.

With the storage element charging (time interval $t_0 - t_1$), the change in current through T1 pulse-transformer primary (fig. 6-10, view A) is too slow to produce an output voltage. Thus, the T1 secondary voltage E_{out} (fig. 6-10, view D) is zero. When the modulator switch closes (time t_1), the rapid decrease in storage-element voltage (E_{st}, fig, 6-10, view C) appears across the T1 primary and induces a high voltage in the T1 secondary.

During the modulator pulse (time interval $t_1 - t_2$), voltage E_{out} remains constant. At the end of the pulse (time t_2), T1 secondary voltage decreases suddenly to zero. Thus, pulse transformer T1 converts the rapidly changing storage-element voltage to a steep, high-voltage pulse.

Resonance Charging

If the charging resistor is replaced by an induct or, the charging circuit becomes series resonant. When a dc voltage is applied to a series-resonant circuit (fig. 6-11, view A), capacitor C begins charging through inductance L. When E_c approaches the applied dc voltage E, the magnetic lines of force due to current flow through L begin collapsing and sustain the charging current. In this way capacitor voltage E_c rises to its maximum value, which is twice the applied dc voltage, E. Due to circuit losses in actual practice, this voltage is approximately 1.9E. The capacitor voltage then oscillates at the resonant frequency of the LC circuit. These oscillations gradually decay until E_c becomes constant, and equals E, the applied dc voltage. This is called dc resonance charging.

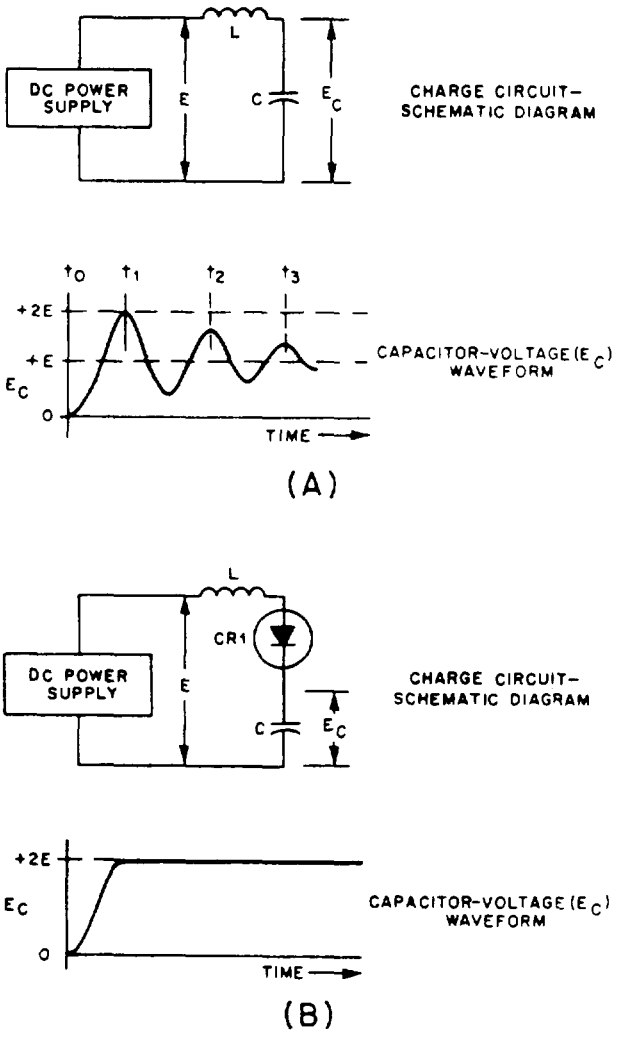


Figure 6-11.-(A) Dc resonance charging; (B) resonance charging with a diode.

Dc resonance charging is used only when the pulse repetition period corresponds to the resonant frequency of the LC circuit. For example, with dc resonance charging, the modulator switch closes at the instant the capacitor voltage reaches its maximum value (time t_1) (fig. 6-11, view A). The advantage of dc

resonance charging is that it permits the storage element to be charged to a voltage twice the dc power-supply voltage. Its disadvantage is that the pulse repetition rate is fixed by the resonant frequency of the LC charging circuit.

Addition of a diode to the resonant charging circuit permits the storage element to charge to a dc voltage twice the applied dc voltage. As a result, E_c increases to its maximum value (2E), and then remains constant. Look at figure 6-11, view B. Notice the schematic diagram and the capacitor voltage (E_c) waveform of a resonant charging circuit that uses a diode. Note that diode CR1 is connected in series with the charging impedance (inductor L) and the storage element (represented by capacitor C). Since CR1 can conduct in only one direction, capacitor C is charged through the dc power supply. The storage element (C) can be discharged at any time after E_c reaches its maximum value. Thus, the pulse repetition rate can be varied over a wide range.

RECEIVERS

Learning Objective: *Recall the purpose of radar receivers to include mixers, oscillators, preamplifiers, amplifiers, and automatic frequency control.*

Because the received RF echo pulses are very small, the radar receiver must have high-gain and low-noise capabilities. Because of the noise produced by RF amplifier stages at microwave frequencies, radar receivers are modified slightly. Instead of RF amplifier stages, the typical radar receiver uses a waveguide balanced mixer (microwave mixer) and an IF preamplifier to produce the gain normally achieved by RF amplifier stages. This is done with much less inherent noise.

MICROWAVE MIXER

The typical radar receiver microwave mixer is a waveguide balanced mixer. (See figure 6-12.)

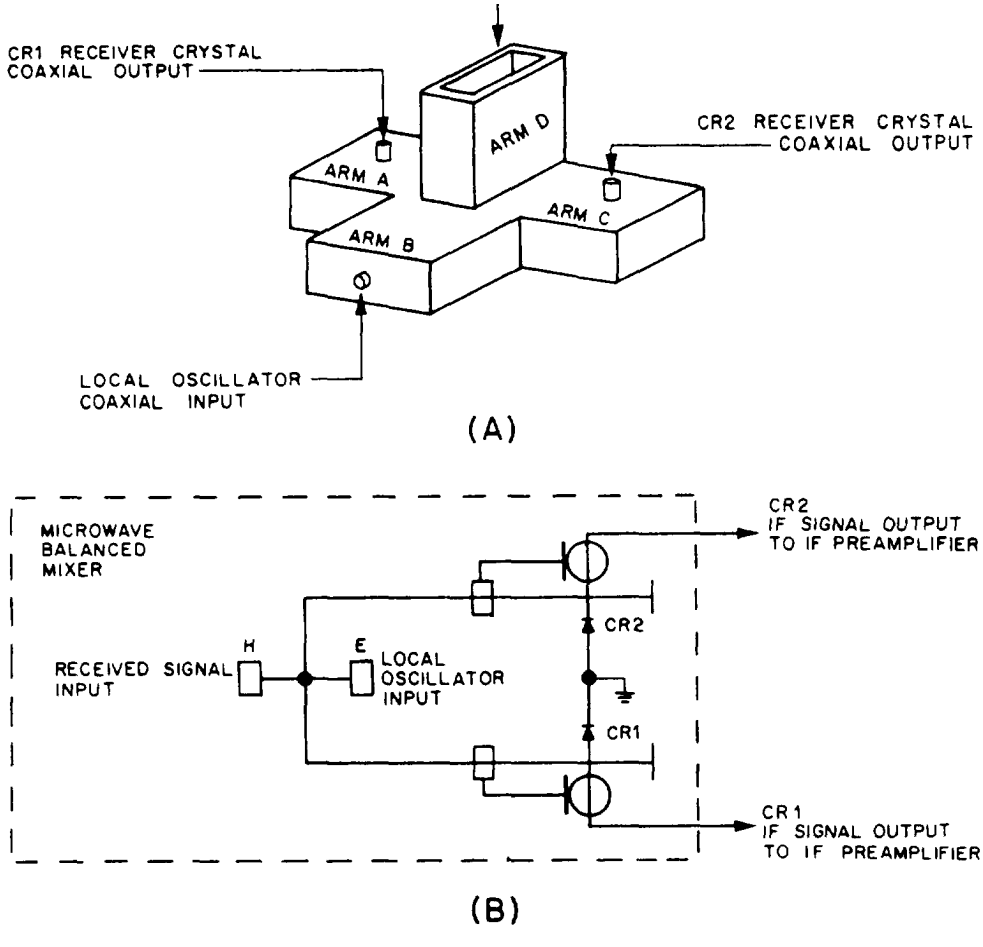


Figure 6-12.-(A) Waveguide balanced mixer; (B) waveguide balance mixer schematic diagram.

This section of waveguide forms a hybrid junction (also referred to as "hybrid T" or "magic T"). It is a waveguide arrangement with four branches. The branches are constructed so that energy (signals) entering one of the four branches is coupled to only two of the three remaining branches. In figure 6-12, view A, the four branches are labeled "ARM A," "ARM B," "ARM C," and "ARM D." The receiver crystals (CR1 and CR2) are inserted directly into the waveguide, and coaxial probes are used to couple the output signals. The crystals are located one-quarter wavelength from their respective short-circuited waveguide ends. This is the point of maximum voltage along a tuned line. The crystals are also connected to an impedance network located in the IF preamplifier, This network can be adjusted for optimum coupling and best noise figure.

The local oscillator signal is injected into Arm B (fig. 6-12, view A) by a coaxial probe. The signal is distributed as shown in figure 6-13. Notice that the local oscillator signal is in phase across the crystals. The received signal is injected into Arm D (fig. 6-12, view A) by waveguide connection from the antenna. The signal is distributed as shown in figure 6-13, view B. Note the signal is out of phase across the crystals. The resulting fields are illustrated in figure 6-13, view C.

Because there is a difference in phase between the received signals applied across the two crystals, and because the local oscillator signal is in phase across both crystals, there will be a condition when both signals applied to CR1 will be in phase and the signals applied to CR2 will be out of phase. This results in an IF frequency (difference between local oscillator and received signal frequencies) signal of one polarity across CR1 and of the opposite polarity across CR2, When these two signals are applied to the input circuit of an IF preamplifier, they will add. Outputs of the same polarity will cancel each other. This action helps to eliminate inherent local oscillator noise. The IF preamplifier will be discussed later in this chapter.

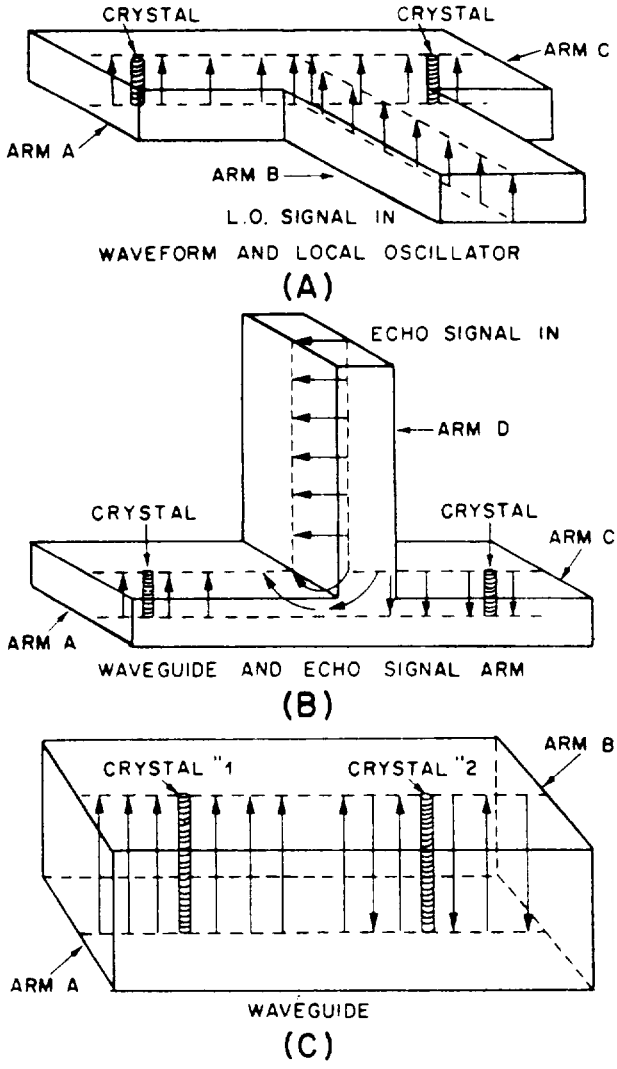


Figure 6-13.-Balanced mixer fields.

LOCAL OSCILLATOR

For years, the local oscillator used in practically all microwave radar systems was the reflex

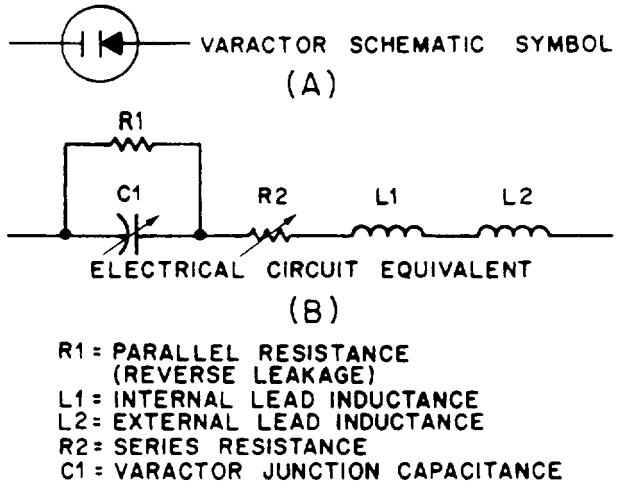


Figure 6-14.-Equivalent circuit for a varactor diode at microwave frequencies.

klystron. With the advent of solid-state devices and particularly the varactor, it became possible to design more efficient oscillator circuits. Most modern radars today use solid-state, variable-frequency, voltage-controlled, varactor oscillators.

The varactor diode is a semiconductor device that is employed as a variable reactance circuit element. The variable reactance is provided by the P-N junction capacitance, which varies as a function of the voltage applied to it. The varactor operates principally between a very small positive bias and the reverse breakdown voltage. Under these conditions, the varactor shown in figure 6-14, view A, can be represented electrically by the equivalent circuit shown in figure 6-14, view B. The applied voltage can vary the junction capacitance sufficiently to provide a useful capacitance change. This variance enables the varactor diode to be used for tuning oscillator tank circuits over a wide frequency range.

Figure 6-15 is a diagram of a typical varactor, voltage-controlled, variable-frequency oscillator circuit used in present-day radars. The input sawtooth voltage (developed by an automatic frequency control (AFC) circuit, discussed later in this chapter) is applied across R7 to varactor diode CR1. CR1 is essentially a voltage-sensitive variable capacitor in series with a semiconductor diode. The diode portion of CR1 is effectively at RF ground because it is connected to the –12V bias line and bypassed to ground by capacitor C5. The incoming sawtooth voltage changes the capacity of CR1 P-N junction, which is in parallel with coil L4, forming a resonant tank circuit in the collector circuit of oscillator Q1. Coil L4 is effectively connected from the collector to the base of Q1, due to bypass capacitor C4. The feedback necessary to sustain oscillations is provided by adjustable capacitor C9. The RF effects of bypass capacitors C4 and C5 effectively place the base of Q1, L4, and CR1 to ground or a common tie. This causes Q1, CR1, and L4 to form a transistor Colpitts oscillator circuit.

The sine wave output signal of Q1 (whose frequency is dependent upon the capacitance of CR1) is applied to the base of Q2. Q2 is a buffer amplifier, whose output is coupled by capacitor C1 to the microwave balanced mixer local oscillator input probe.

IF PREAMPLIFIER

The typical radar IF preamplifier is a low-noise, high-gain amplifier, which is tuned to the receiver's IF frequency (normally in the range

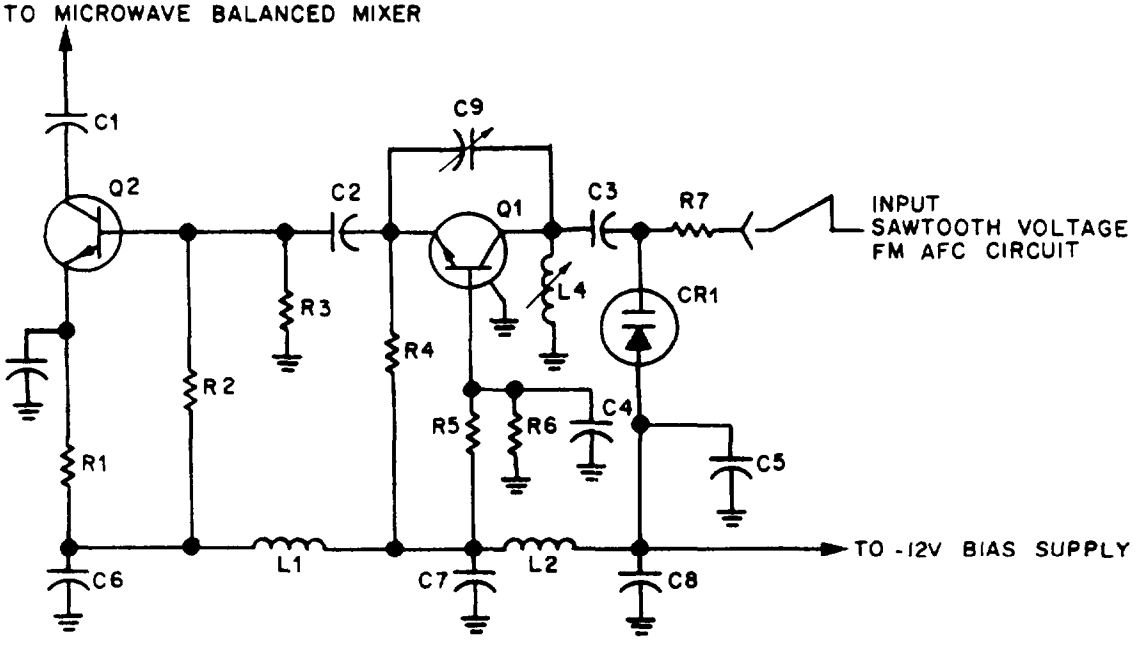


Figure 6-15.-Typical radar varactor, voltage-controlled oscillator (VCO) circuit.

of 60 MHz). Figure 6-16 is a functional signal flow diagram of a typical IF preamplifier. The diagram shows the preamplifier connected to a microwave balanced mixer. Figure 6-17 is a simplified schematic of the IF preamplifier with only those components labeled that appear in figure 6-16. Refer to these figures during the following discussion.

The input from the balanced mixer (received signal at the IF frequency) is coupled across C1 and C2, through L1 to the base of Q1. R1 and R2 adjust the bias of the balanced mixer crystals at the best noise figure. L1 and C4 form an impedance matching network, which is also adjusted for the best noise figure. Q1 and Q2 are high gain, cascode amplifiers. The output of Q2 is taken off the collector and coupled across C3 to the bases of Q3 and Q4. Q3 and Q4 are parallel amplifiers that provide the necessary signal power without gain compression. L2 tunes the center frequency of the amplifier and T1 adjusts the amplifier's bandwidth and gain figure. The output signal is coupled across T1 and is the input to the receiver's IF amplifier stages. The rest of the unlabeled components establish biasing, etc., for the amplifier. The STC (sensitivity time control) input controls the gain of the preamplifier to prevent saturation of the display indicator by large nearby ground clutter targets. This input is a negative ramp voltage from an STC circuit. It consists of a monostable multivibrator, a charging RC network, and a driver amplifier. The resistance of the RC network is controlled by an adjustable pot located on the radar's control panel. The operator adjusts the pot for best picture. By adjusting the RC time, both the ramp duration and amplitude can be set. This lowers the gain of the preamplifier for a period of the receive time. This period is usually from 0 to 20 miles on the indicator. This voltage is applied to the emitters of Q1, Q3, and Q4 and controls the gain of these stages by controlling the emitter-base bias. The more STC adjusted in by the operator, the less negative voltage is applied to the emitters during STC time, thus decreasing the amplification factor.

IF AMPLIFIERS

At this point you might want to review Navy Electricity and Electronics Training Series (NEETS), Module 18, *Radar Principles,* as an aid

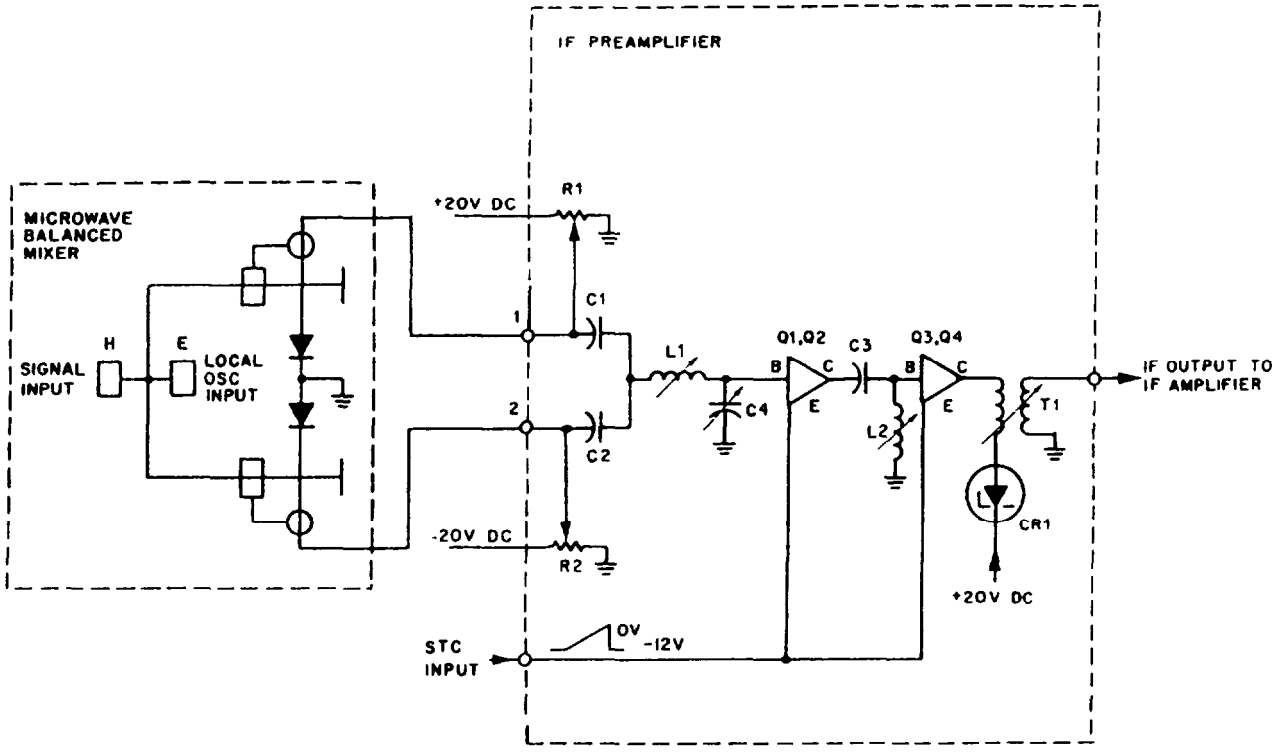


Figure 6.16.-Typical IF preamplifier functional signal flow diagram.

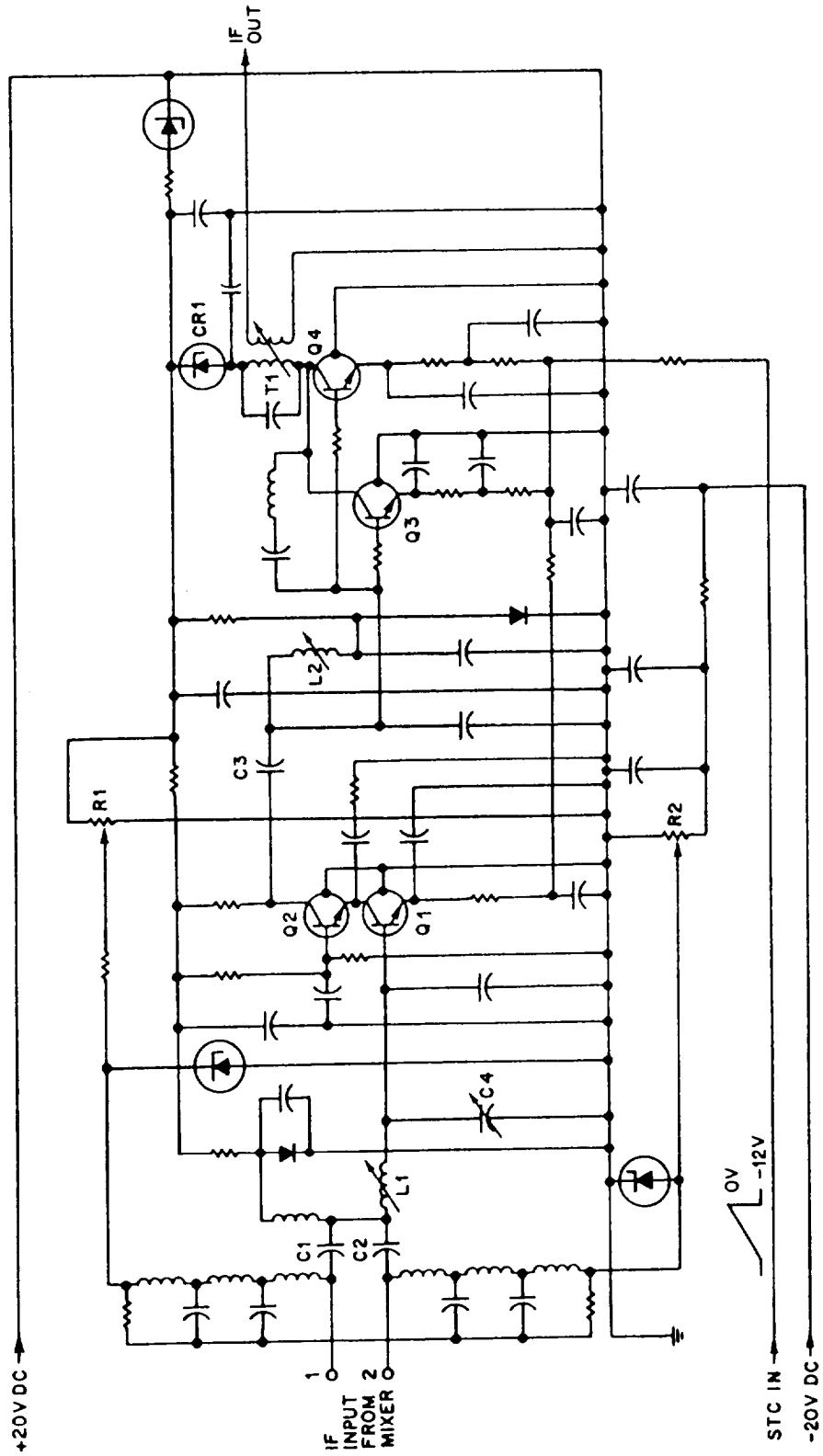


Figure 6-17.—Typical IF preamplifier simplified schematic.

to your understanding of IF amplifier operation. Radar receiver IF amplifiers require high gain to amplify the input signal to the level required to operate the detector. Many different circuit arrangements are used to achieve this required gain. One circuit arrangement, which is increasing in use, is the logarithmic amplifier. This type of amplifier stage produces high gain while maintaining the resonant frequency and bandpass of the tuned coupling circuits fairly constant over the dynamic range of the input signal. The output of a logarithmic amplifier is a logarithmic (as opposed to linear) function of its input signal.

Figure 6-18 is a functional signal flow diagram of a typical logarithmic IF amplifier. Figure 6-19 is a schematic diagram of the first IF amplifier stage shown in figure 6-18. All of the IF amplifier stages (1st through 5th) are identical. Refer to figures 6-18 and 6-19 during the following discussion.

The IF input signal (from the IF preamplifier) is coupled across C1 to the bases of the A and B amplifiers. These amplifiers are in parallel across the output coupling transformers. The A amplifier consists of Q1 and Q2 and the B amplifier consists of Q3 and Q4. Both A and B amplifiers are single-ended differential amplifiers.

The A amplifier (Q1 and Q2) has a constant gain of approximate unity as determined by R3 and R4. Since there is no load resistor for Q1, the collector bias supply filter capacitor (located in the +7 Vdc power supply) effectively grounds the collector for ac signals. The output of Q1 is taken off its emitter and applied through R3 and R4 to the emitter of Q2. The output of Q2 is taken off its collector and applied to T1, the interstage transformer coupler. T1 and C8 form a tuned tank, which is tuned to the IF frequency.

The B amplifier (parallel with the A amplifier) has a high gain for weak signals. It is quickly limited as the input signal strength increases. When Q3 conducts (in the same manner as Q1), the output signal is taken off its emitter at the top of R6. The signal is coupled across C6 to the emitter of Q4. Q4 is so biased that it will cut off before Q3 reaches saturation. The weaker the input signal, the more Q4 will conduct and aid the output of Q2. The output of Q4 is taken off its collector and is applied to T1 along with the

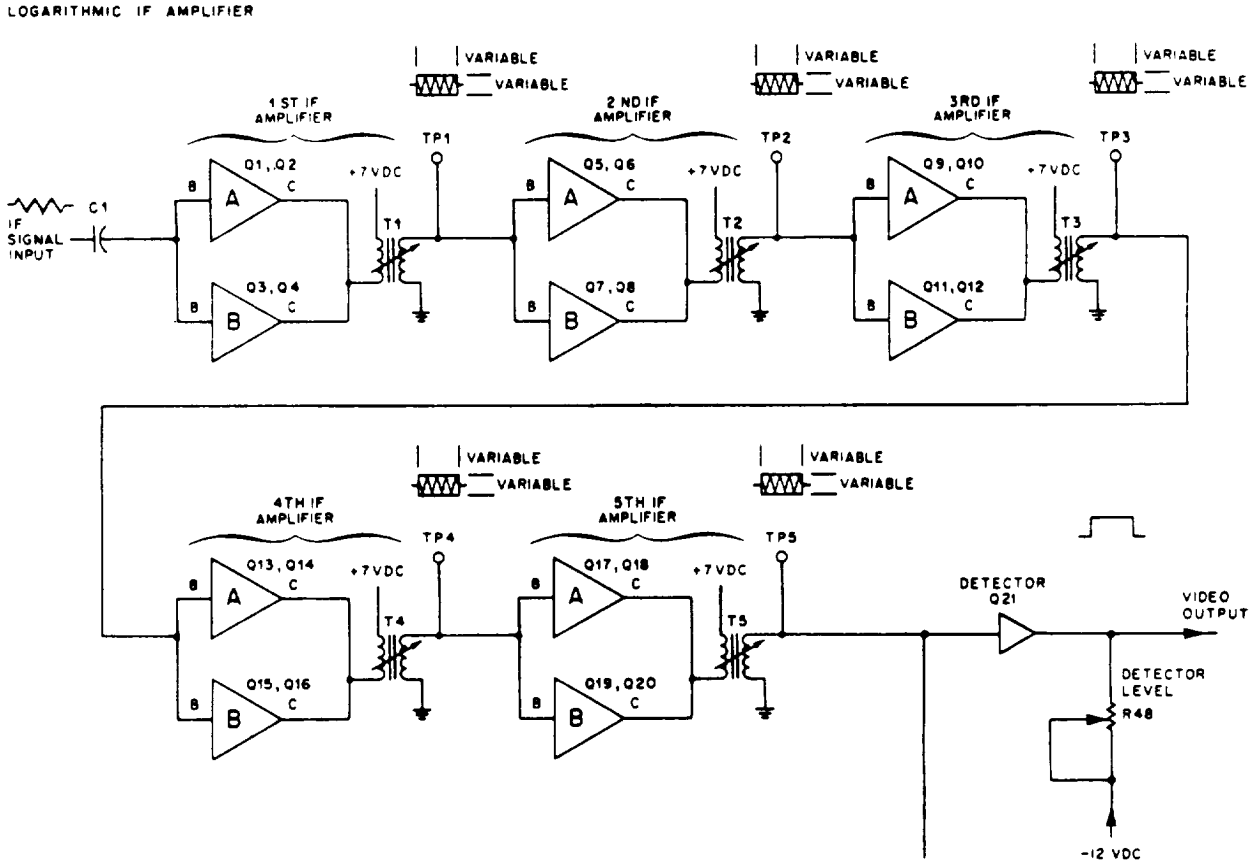


Figure 6-18.-Logarithmic IF amplifier functional signal flow diagram.

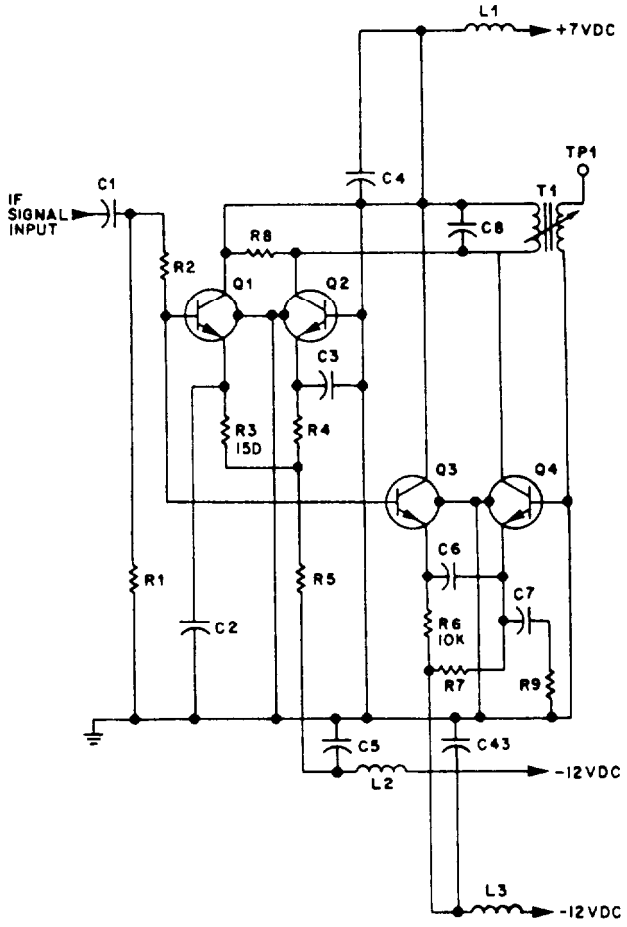


Figure 6-19.-Typical IF logarithmic amplifier stage schematic diagram.

output of Q2. Since Q2 and Q4 are in parallel, feeding a common load, as the input signal increases, the high-gain stage quickly decreases toward unity. This closely approximates a logarithmic response. By cascading amplifiers (lst through 5th), a large dynamic input range is attained. The output is detected after the last (5th) stage and goes to the video amplifier stages.

AUTOMATIC FREQUENCY CONTROL (AFC) CIRCUITS

The purpose of the radar receiver's AFC circuits is to tune the receiver local oscillator to the correct operating frequency. For the receiver to process the received signal, the local oscillator must be tuned to the proper frequency so that the IF frequency output of the receiver's balanced mixer (difference frequency of the local oscillator and received signal) is correct. To ensure this, the local oscillator must be tuned during transmit time (prior to the reception of a return pulse). To accomplish this, the AFC circuits use another microwave balanced mixer similar to the receiver's balanced mixer shown in figure 6-12, view B. The inputs to this mixer are the receiver local oscillator and an attenuated sampled transmitted pulse. The output IF frequency would be the same as the receiver balanced mixer. The received signal is the transmitter pulse reflected off a target. The output of the AFC microwave balanced mixer goes to the receiver's AFC circuits. Figure 6-20 is a

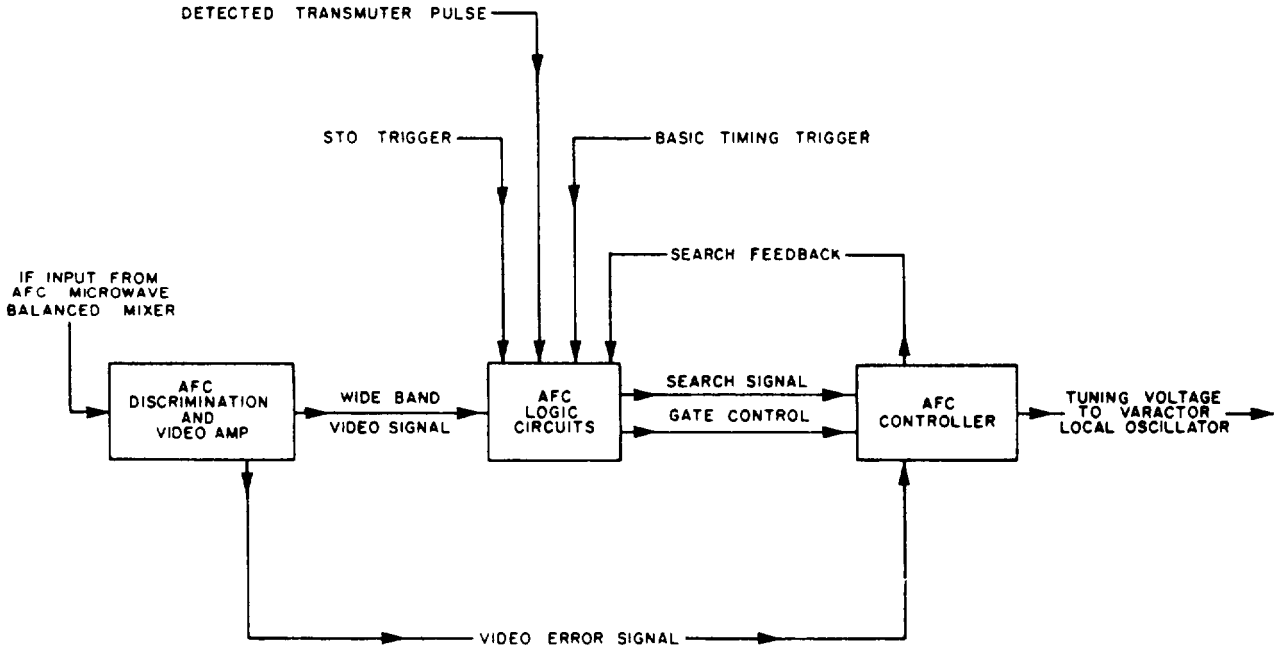


Figure 6-20.-AFC circuit simplified block diagram.

simplified block diagram of a typical AFC circuit. Refer to this figure during the following discussion.

There are three basic steps or modes of operation used by the AFC circuits to tune the local oscillator to the correct frequency. They are the "search," "acquisition," and "loop control" modes.

Search Mode

AFC circuits go into a search mode any time the local oscillator is so far off frequency that the IF frequency produced by the balanced mixer (difference frequency of local oscillator and transmitter frequency) is outside the limits of the receiver's IF tuned circuits and cannot be processed. An IF preamplifier contained in the AFC discriminator is tuned to the receiver's IF frequency. Therefore, if the IF input to the discriminator is beyond the receiver's bandpass limits (usually ±10 MHz), the signal will not be processed by the AFC discriminator. When this occurs, there are no output signals out of the discriminator to either the AFC logic circuits or the AFC controller circuits. Instead, a comparator (fig. 6-21) (AR2), part of the AFC logic circuit, gets an input (search feedback) from an integrator (fig. 6-22) (AR2), which is part of the AFC controller circuit. This input may be either positive or negative, depending on the output state of the integrator. The output of the comparator (fig. 6-21) is coupled through R1, bypasses switch Q22 (which is cut off as there is no wideband video signal input from the discriminator), and goes to CR1 and CR3 (fig. 6-22) of the AFC controller. Depending on the polarity of the comparator output, the output goes through either CR1 and CR2, or CR3 and R1, to the integrator AR2. The gain of the integrator is different for the different polarities of the comparator output. This results in a sawtooth output voltage from the integrator AR2. The sawtooth voltage goes through the summing amplifier AR3 and is also fed back to the comparator AR2, (fig. 6-21). This causes it to change states when a certain voltage level is reached. The output of the summing amplifier AR3 is applied to the sample gate Q2 (N-channel JFET). The gate Q2 is open at this time. (This will be explained when we cover the sample gate control, which is part of the AFC logic circuit.) The sawtooth goes through Q2, across a hold charging circuit consisting of R7, 8, 9, and 10; and C5, 6, 7, and 8, through gate Q3, is amplified by drivers Q4 through Q6, and is applied to the receiver local oscillator varactor. The amplitude

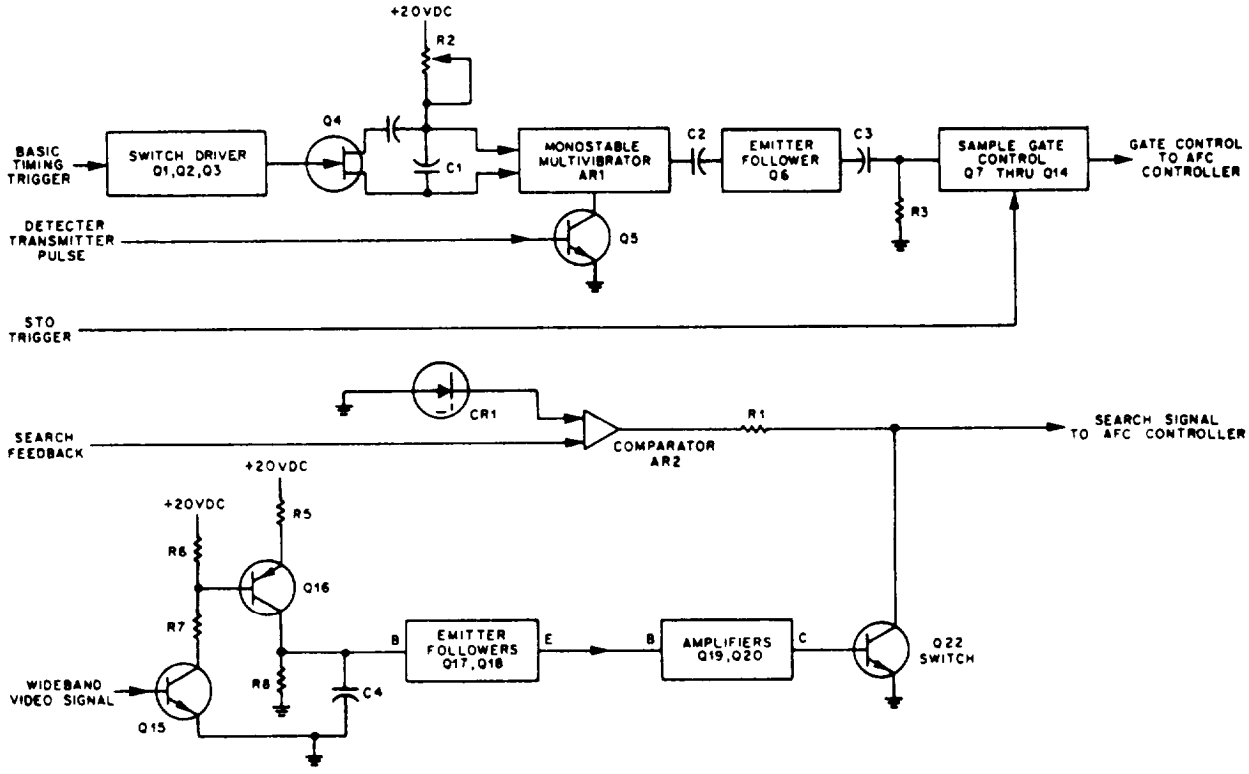


Figure 6-21.-AFC logic circuit functional signal flow diagram.

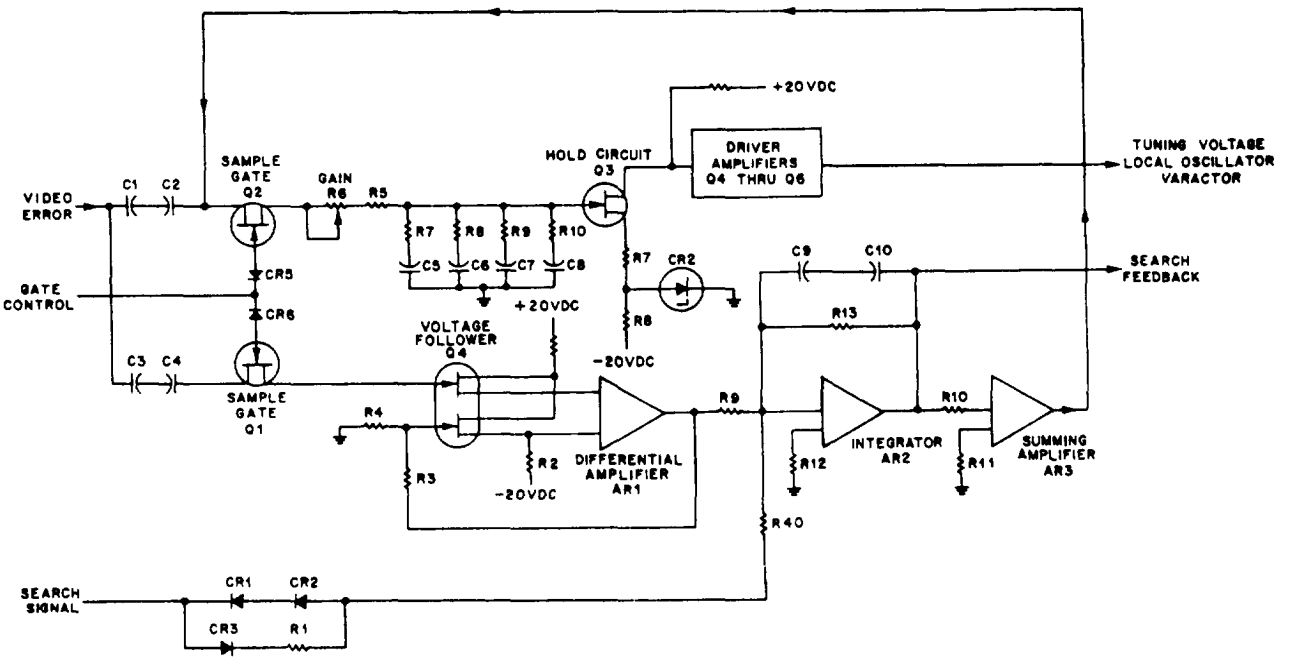


Figure 6-22.-AFC controller functional signal flow diagram.

of the sawtooth will cause the local oscillator to sweep through its entire operating range. It is also large enough to overcome the hold circuit.

When the local oscillator reaches a certain frequency, it is mixed with the sampled transmitted pulse in the microwave mixer. When an IF frequency is produced within the IF bandpass of the receiver (±10 MHz), the IF signal will be processed by the AFC discriminator circuits. At this point, acquisition has occurred.

Acquisition Mode

The AFC mixer's IF frequency output must be within the ±10 MHz bandpass of the AFC discriminator's preamplifier. Once this happens, the input to the discriminator (fig. 6-23) will be

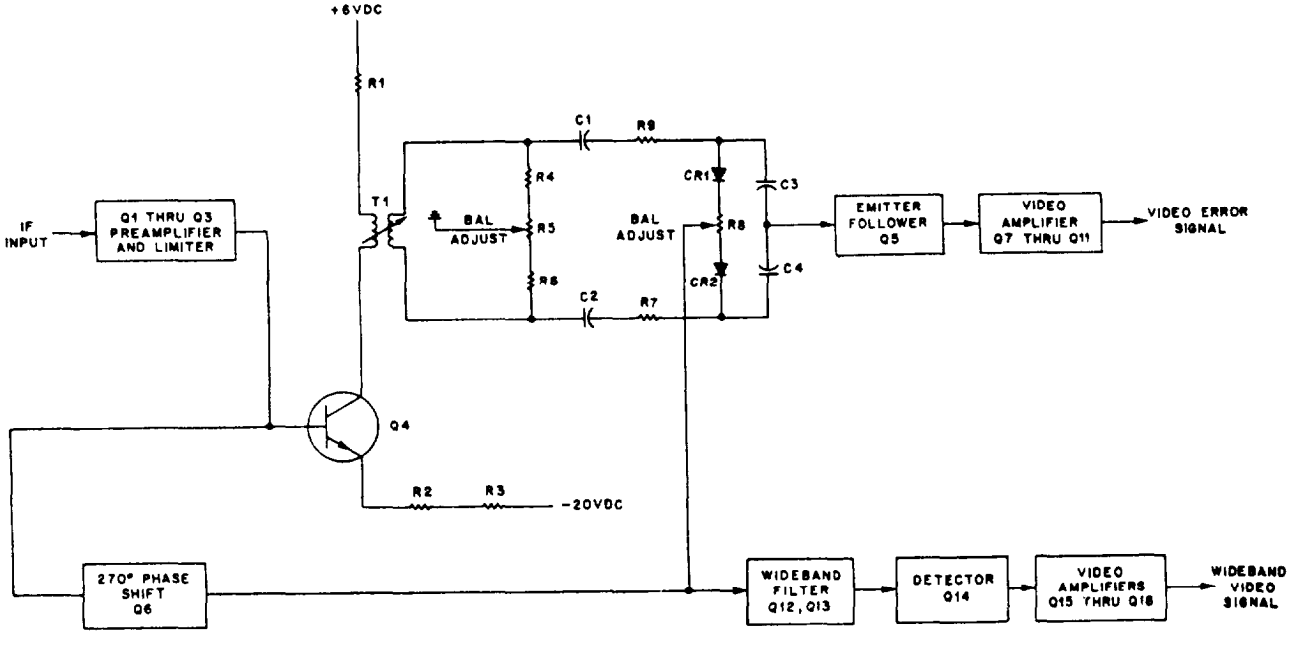


Figure 6-23.-AFC discriminator and video amplifier functional signal flow diagram.

amplified/limited (depending on signal amplitude) by Q1 through Q3. The output of Q3 is fed via Q4 to the discriminator circuit, consisting of T1, R4, R5, R6, C1, C2, R7, R8, R9, CR1, CR2, C3, and C4. The output of Q3 is also fed to Q6. The circuitry of Q6 phase shifts the output (delays the signal) by 270 degrees. The output of Q6 is applied between CR1 and CR2, and is the reference signal for the discriminator. The output of Q6 is also applied to wideband filter Q12 and Q13.

The operation of the discriminator is similar to the Foster-Seeley discriminator explained in Navy Electricity and Electronics Training Series (NEETS), Module 12, *Modulation Principles.* It is basically a phase detector with the output rectified and filtered. The output of the discriminator is a pulse, the width of the transmitter pulse, that has an amplitude and polarity that is determined by the frequency difference between the local oscillator and the sampled transmit pulse. This output is amplified by Q7 through Q11 and is the video error signal, which is applied to C1 and C3 (fig. 6-22) of the AFC controller.

The output of the wideband filter Q 12 and Q13 (fig. 6-23) is detected by Q14, amplified by Q15 through Q18, and is the wideband video signal, which is applied to Q15 (fig. 6-21) of the AFC logic circuit.

The wideband video signal causes Q15 and Q16 to conduct, which charges C4. When the charge on C4 reaches a predetermined value, the signal is coupled by Q17 and Q18 to amplifiers Q19 and Q20, and turns on switch Q22. Q22 clamps the output of the comparator AR2, and it also clamps the feedback circuit (fig. 6-22) of the integrator AR2, which is part of the AFC controller. The integrator will no longer generate the (search) sawtooth voltage to the local oscillator. At this point, the search mode is terminated, acquisition is complete, and tracking (loop-control mode) begins.

Loop-Control Mode

There are actually two paths or loops comprising the loop-control mode. One is commonly referred to as "slow loop," which corrects the oscillator frequency from ±10 MHz to within ±5 MHz of the desired IF center frequency. The second path is referred to as the "fast loop," and it will correct the oscillator frequency from ±5 MHz to the correct frequency.

SLOW-LOOP OPERATION.— The video error signal (fig. 6-22), which was the output of the discriminator (beginning at acquisition), is applied to C1 and C3. C1 is the fast-loop circuit that is ineffective when the local oscillator frequency error exceeds ±5 MHz. (We'll see why shortly.) The video error signal is coupled across C3 and C4 to sample gate Q1. Q1 just as Q2 in search, is open at this time. (We'll explain how when we discuss the sample gate control later in this chapter.) The signal passes through gate Q1 to Q4. Q4 and AR1 operate as a voltage follower circuit. The output of AR1 is integrated by AR2, amplified by AR3, and applied across the opened sample gate Q2 to the hold circuit. Slow-loop operation can be summarized as follows: So long as the oscillator frequency error exceeds ±5 MHz, the slow-loop circuit takes a series of video error signal samples. The slow-loop circuit integrates them until a sufficient voltage level is reached to charge the hold circuit to open gate Q3 and drive the local oscillator frequency to within ±5 MHz of the desired frequency. Once this is accomplished, fast-loop operation begins.

FAST-LOOP OPERATION.— The video error-signal output of the discriminator will now be coupled across C1 and C2 to the sample gate Q2. The sample gate at this time will open periodically as determined by the sample gate control circuits, which are part of the AFC logic circuit. When gate Q2 opens, the video error signal goes to the hold circuit of Q3. The hold circuit, charged by the slow-loop circuits, is at the threshold voltage level to turn on Q3. The error signal amplitude is sufficient to turn on Q3 and send a short voltage burst to the local oscillator varactor to tune the oscillator to the desired frequency. Once the desired frequency is reached, there will no longer be a video error-signal output from the discriminator and no further tuning of the local oscillator. Should the transmitter frequency or local oscillator frequency drift slightly, a video error-signal output from the discriminator would again go through the fast loop circuits and retune the local oscillator. In this manner the AFC circuits maintain the desired

IF frequency output of the receiver's balanced mixer.

Sample Gating Operation

The sample gates, Q1 and Q2 (fig. 6-22), are initially opened by an STO trigger (fig. 6-21) being applied to the sample gate control circuits Q7 through Q14. Prior to receiving the STO trigger, the output of Q14 is a highly negative (approximately –15 Vdc) voltage, which applied across (fig. 6-22) CR5 and CR6 causes the N-channel gate to be closed. The STO trigger is a pulse from the radar synchronizer and occurs a few microseconds prior to the basic timing pulse, which fires the transmitter and the scope sweep circuits, etc. This ensures the sample gate is opened prior to any possible output from the receiver or AFC balanced mixers. (The output of the balanced mixers is the difference frequency of the local oscillator and sampled transmitter pulse.) The gates Q1 and Q2 (fig. 6-22) will close (fig. 6-21) with the detected transmit pulse being applied to Q5. This pulse triggers a monostable multivibrator, AR1. The negative pulse output of AR1 causes the sample gate control Q7 through Q14 to turn off (close) the sample gates Q1 and Q2 (fig. 6-22). The multivibrator will generate different negative output pulses depending on the PRF of the radar. Since a typical radar is normally capable of operating at two different PRFs and pulsewidths for long- and short-range operation, the basic timing pulse (fig. 6-21) is applied to switch drivers Q1 through Q3. The output of the switch drivers opens and closes gate Q4, which changes (determines) the time constant of AR1. Therefore, sample gates Q1 and Q2 (fig. 6-22) will be open for different periods of time, depending on the PRF of the radar. However, the overall time the sample gates are opened is from the STO trigger time to the time of the detected transmitter pulse. The time constant of the monostable multivibrator is such that gates Q1 and Q2 (fig. 6-22) will close just prior to the trailing edge of the transmitted pulse. This is to ensure that the AFC circuits will not inadvertently tune the receiver local oscillator during the receive cycle of the radar operation.

SPECIAL RECEIVER CIRCUITS

Gain control of radar IF amplifiers takes many different forms. The simplest type is manual gain control. More complex forms of gain control are automatic gain control (AGC), instantaneous automatic gain control (IAGC), and sensitivity time control (STC). Gain control is necessary to adjust the radar-receiver sensitivity for signals of widely varying amplitude. Some gain-control circuits are used to overcome unintentional or intentional interference (jamming).

Although it is possible to control IF gain with only one IF amplifier stage, the amount of control is usually insufficient. Because a stage has capacitance between its input and output ends, a signal is coupled to the output end even when the stage is cut off. The maximum variation in gain by the control of a single stage is approximately 20 dB. With two-stage gain control, approximately 40 dB of gain variation can be obtained.

Automatic Gain Control (AGC)

Many radar sets are provided with AGC and manual gain control. Provision is usually made for switching between automatic and manual gain control. In this way, manual gain control can be used, if necessary, to adjust for best reception of a particular signal.

The various types of gain control differ in the following ways:

1. Circuits used
2. Speed of response
3. Type of response

The simplest type of AGC adjusts the IF amplifier bias (and gain) according to the average level of the receiver signal. AGC is not used as frequently as other types of gain control because of the widely varying amplitudes of radar return signals.

With AGC, gain is controlled by the largest received signals. When several radar signals are being received simultaneously, the weakest signal may be of greatest interest. IAGC is used more frequently because it adjusts receiver gain for each signal.

Instantaneous Automatic Gain Control (IAGC)

A typical IAGC circuit is essentially a wide-band dc amplifier that instantaneously controls the gain of the IF amplifier as the radar return signal changes in amplitude. This is accomplished by using an output of the second detector of the receiver as bias for the amplifier. The effect of IAGC is to decrease the amplification of strong signals and allow full amplification of weak signals. The range of IAGC is limited by the number of IF stages in which gain is controlled. This is the reason most modern receivers use sensitivity time control.

SENSITIVITY TIME CONTROL (STC).— In radar receivers, the wide variations in return signal amplitude make adjustment of the radar-receiver gain difficult. You should adjust receiver gain for best visibility of nearby target return signals. Circuits used to adjust amplifier gain with time, during a single pulse repetition period, are called STC circuits.

Figure 6-24 shows a schematic diagram and output-voltage waveform of an STC circuit. The input signal to the STC circuit (fig. 6-24, view A) is a pulse obtained from the radar modulator. When the modulator pulse is applied to the base of emitter follower Q1, a large voltage appears across capacitor C1. At the same time, a negative voltage appears across capacitor C2. The amount of negative bias developed across capacitor C2 is determined by the setting of potentiometer R5. A large voltage across C2 drives Q2 beyond cutoff. Thus, E_{out} (curve D, fig. 6-24, view B) remains constant while the voltage across C2 decays toward zero. When the voltage across C2 becomes equal to the Q2 cutoff voltage, Q2 begins to conduct and E_{out} rises toward zero.

Output voltage E_{out} (fig. 6-24, view A) is applied to the base of an IF amplifier (not shown), and thus places a constant bias on the IF amplifier for a short time after the modulator pulse. During this period, the IF amplifier gain is held constant. When the Q2 output voltage begins to decrease,

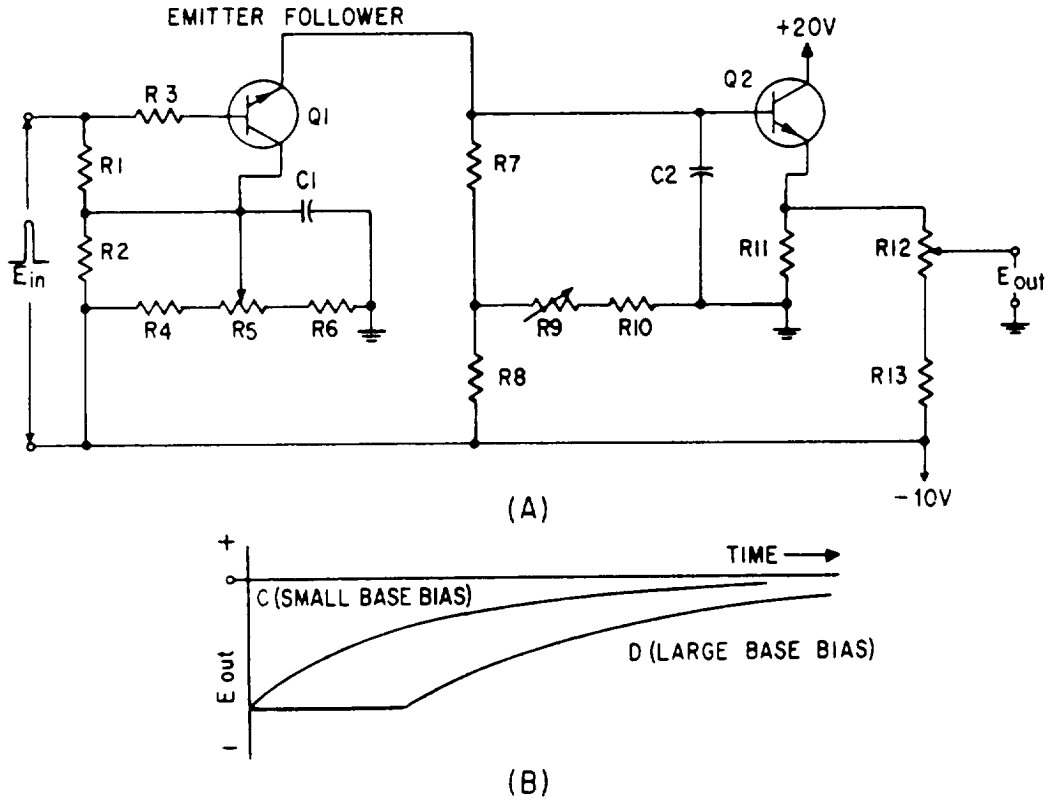


Figure 6-24.-Sensitivity time control circuit. (A) Schematic diagram; (B) waveform of output voltage.

less bias is applied to the IF amplifier, and the receiver sensitivity increases with time. As a result, weak signals from distant targets are amplified more than signals from nearby targets.

If potentiometer R5 is set so that Q1 is not driven beyond cutoff, the bias on Q2 begins to decrease as soon as the modulator pulse ends. In this case, output voltage E_{out} rises toward zero, as shown by curve C (fig. 6-24, view B). As a result, a large negative bias is applied to the IF amplifier at the time of the modulator pulse, thereby decreasing IF amplifier gain.

As soon as the modulator pulse ends, the bias applied to the IF amplifier begins to decrease, and IF amplification begins to increase. Thus, the IF amplifier gain is minimum directly after a modulator pulse. Also, the gain increases at a later time when weak signals from distant targets are expected.

The combination of STC and IAGC circuits results in better overall performance than with either type of gain control alone. STC decreases the amplitude of nearby target return signals, while IAGC decreases the amplitude of larger-than-average return signals. Thus, normal changes of signal amplitudes are adequately compensated by the combination of IAGC and STC.

In some cases very large changes of signal amplitude are encountered. For example, enemy jamming may produce large amplitude interfering signals. The interfering signals may be either continuous or pulsed. The interfering signal amplitudes may be large enough to block the receiver, and thus cover up signal return from enemy aircraft. To overcome jamming, special receiver circuits, called antijamming circuits, are used.

ANTIJAMMING CIRCUITS.— Among the many different circuits used to overcome the effects of jamming, two important ones are gated AGC circuits and short-time-constant circuits. A gated AGC circuit permits signals that occur only in a very short time interval to develop AGC. If large amplitude pulses from a jamming transmitter arrive at the radar receiver at any time other than during the gating period, the AGC does not respond to these jamming pulses.

Without gated AGC, a large received signal from a jamming transmitter would cause the automatic gain control to follow the interfering signal and to decrease the desired signal amplitude to an unusable value. Because gated AGC produces an output signal for only short times, the AGC output voltage must be averaged over several cycles to keep the automatic gain control from becoming unstable.

Although gated AGC does not respond to signals that arrive at times other than during the desired target return signals, AGC can do nothing with interference that occurs during the gating period. Neither can gating the AGC prevent the receiver from overloading due to jamming signal amplitudes far in excess of the desired target return signal of that particular amplitude. As an aid in preventing radar-receiver circuits from overloading during the reception of jamming signals, short-time-constant coupling circuits are used to connect the video-detector output to the video-amplifier input circuit.

A short time constant or a fast time constant (FTC) circuit is a differentiator circuit located at the input of the first video amplifier. When a large block of video is applied to the FTC circuit, only the leading edge will pass due to the short time constant of the differentiator. A small target will produce the same length of signal on the indicator as a large target, because only the leading edge is displayed. The FTC circuit has no effect on receiver gain. Although it does not eliminate jamming signals, it greatly reduces them.

Video Amplifiers

Video amplifiers are used to amplify the output signal from the video detector to a level high enough to be used by the radar presentation system. Because radar receivers are frequently far removed from the presentation circuits, some video amplification is provided in the radar receivers. Video amplifiers may also be located in the radar presentation circuits (scope).

Since the radar video signal may have frequency components up to several MHz, coaxial cables are used to connect the video output circuit of the receiver to the video input circuit of the presentation system. When these coaxial

cables are long, special video-amplifier circuits are generally used in the radar receiver.

Among the video-amplifier problems that must be met in radar circuits are the following:

1. Limitation of low-frequency response. (This limitation occurs when cathode-bypass capacitors are used.)

2. Limitation of low-frequency response by screen-grid bypass capacitance (capacitance between the screen grid and ground).

3. Limitation of high-frequency response by input interelectrode capacitance, distributed wiring capacitance, and output interelectrode capacitance.

Some of the special video-amplifier circuits mentioned above, whose purpose is to compensate for these problems, are discussed in Navy Electricity and Electronics Training Series (NEETS), Module 8.

The high-frequency performance of solid-state circuits has been improved greatly since the early devices. These early circuits were generally limited to about 500 KHz. Transistors are capable of operating at frequencies far above the operating range of conventional vacuum tubes. Transistors do, however, have high-frequency limitations. The

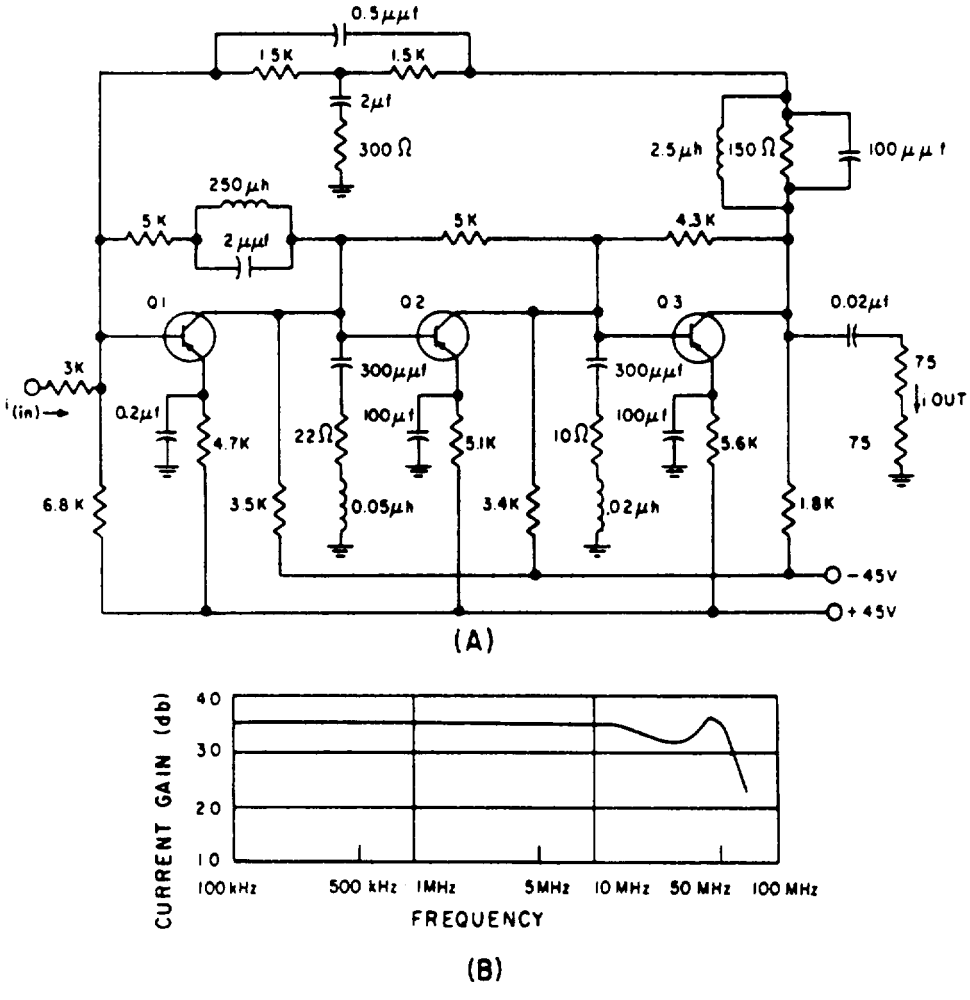


Figure 6-25.-Three-stage, 50 MHz video amplifier. (A) Circuit; (B) frequency characteristic.

design of high-frequency transistor circuits must take into account factors that are not significant at low frequencies.

Basically, transistor high-frequency limitations arise because off transit time effects and the inherent junction capacitance. At high frequencies, these factors become significant and begin to affect the operation of the circuit. Stage gain is lowered and problems involving instability appear as the impedance and gain of the transistor become complex quantities.

When transistors having an upper frequency limit only slightly higher than, or equal to, the high frequency end of the desired video band are used, the attenuation and phase shift due to the transistor must be compensated for in the amplifier. This may be done by the use of compensating networks and/or the use of negative feedback.

Basically, the high-frequency compensation of video amplifiers consists of attenuating the normal midrange gain of the amplifier to within a few dB of the maximum gain obtainable at the highest frequency of interest, so that the bandwidth is extended to this high frequency. This is how negative feedback increases the bandwidth of an amplifier. It may be used around one or more stages, and it results in increased stability as well as bandwidth.

Two terminal high-frequency compensation circuits using two or more compensating elements are commonly used in vacuum tube circuits. These circuits can also be used for the high-frequency compensation of transistor amplifiers, although the design relationships are not quite as straightforward. This method of high-frequency compensation consists essentially of using RLC peaking circuits to maintain a nearly constant amplification factor over the required band.

The three-stage, 50-MHz wide-band amplifier shown in figure 6-25, view A, uses negative feedback frequency compensation in conjunction with an RLC compensating network placed directly in the feedback path. This arrangement provides 34 dB of negative feedback from 50 kHz to 5 MHz.

The current amplification factor is virtually flat at 34 dB from dc to 10 MHz, and within 3 dB to 50 MHz. The feedback RLC network maintains the feedback loss flat to 10 MHz. The interstage RLC networks compensate the gain characteristic between 5 and 100 MHz, and the feedback around the first stage provides compensation in the vicinity of 7 MHz. The emitter bypass capacitors control the frequency response from dc up to approximately 50 kHz. The frequency characteristic of the amplifier is shown in figure 6-25, view B.

Video signals are usually coupled to the presentation circuits through relatively long lengths of low-impedance, large-capacitance coaxial cables. Thus, video output stages generally have a low output impedance.

These coaxial cables may have a capacitance of 20 $\mu\mu$f per foot. Thus, a 5-foot cable would have a capacitance of 100 $\mu\mu$f. To prevent attenuation of high-frequency signals by the shunt capacitance, coaxial cables must be terminated in their characteristic impedance (usually 100 ohms or less).

For coaxial cables that are very short (less than a quarter-wavelength long), the termination resistance may be higher than the characteristic impedance without affecting high-frequency response. Higher values of terminating resistance result in higher output voltages.

INDICATORS

Learning Objective: *Recognize common types of radar indicators to include A-scope, B-scope, and C-scope, E-scan detections and sweeps, range circle generators, and miscellaneous presentations.*

The various types of radar indicators (A-scope, B-scope, PPI-scope, etc.) and some of the fundamental principles of their operation are discussed in Navy Electricity and Electronics Training Series (NEETS), Module 18. You should review the basic radar principles before continuing with this discussion. In the following paragraphs, details of radar indicator operations that go beyond the basic level are presented.

A-SCOPE

Figure 6-26 is the simplified block diagram and scan presentation of a typical A-scope. Although the A-scope is not used in modern weapons systems radar, it is presented to establish a basic understanding of scopes before discussing the more advanced type.

In the operation of the A-scope, an initial trigger pulse from the timer is applied to both the radar transmitter and the one-shot (monostable) multivibrator in the A-scope. The one-shot multivibrator generates the following:

1. A negative gate pulse that is fed to the range marker generator.
2. A negative gate pulse that is fed to the range sweep generator.
3. A positive gate pulse that is fed to the control grid of the cathode-ray tube.

The gate pulse to the range marker generator causes a series of equally spaced range marks to be generated. These range marks are added to the

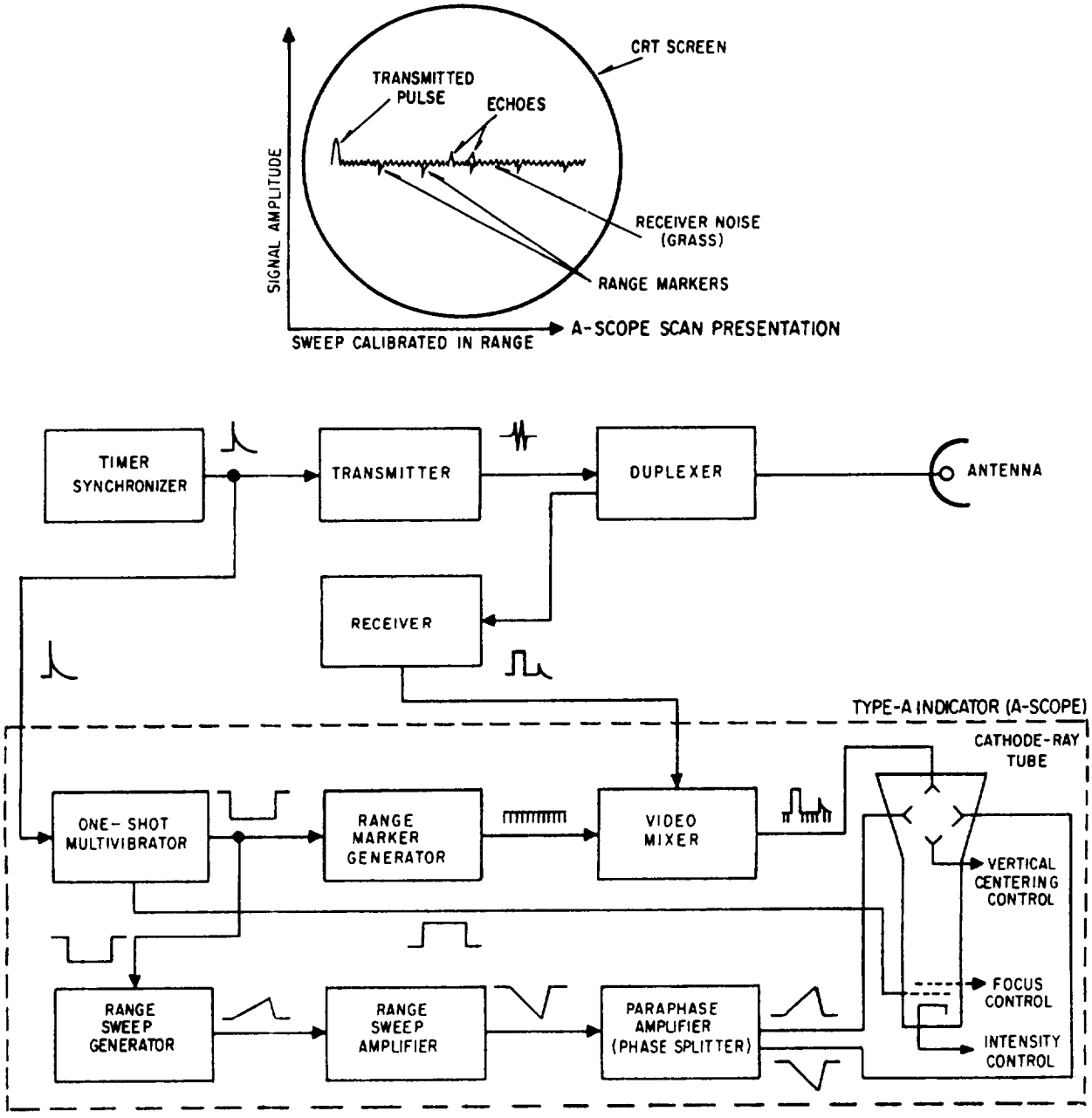


Figure 6-26.-Typica1 A-scope block diagram and scan presentation.

receiver output signal in the video mixer. The output of the video mixer is applied between ground and one vertical-deflection plate of the cathode-ray tube. The other vertical-deflection plate is connected to the vertical-centering control. (In some cases, the receiver output signal is applied to one vertical-deflection plate, and the range marks are applied to the other vertical-deflection plate.)

The negative gate pulse, fed to the range sweep generator, causes a nearly linear sawtooth sweep voltage to be generated. In general, the different timing capacitors in the one-shot multivibrator and in the range sweep generator are connected to a common range switch. In this way, the RC time constants of both circuits are changed simultaneously when the operating range is changed.

When the duration of the negative gate pulse is changed, the duration of the sawtooth sweep voltage is changed, but the amplitude of the sweep voltage is unchanged. Hence, for different operating ranges, the scanning spot travels approximately the same distance across the A-scope screen. However, the speed of the scanning spot increases as the range setting is decreased.

The sawtooth output of the range sweep generator is amplified by the range sweep amplifier, and then applied to the paraphase amplifier (phase splitter). The paraphase amplifier permits the sawtooth sweep voltage to be applied in push-pull to the horizontal-deflection plates of the cathode-ray tube. This reduces defocusing of the electron beam, which usually results when the sweep voltage is applied to only one horizontal-deflection plate.

The positive gate pulse applied to the control grid of the cathode-ray tube intensifies the electron beam during the sweep time. This enables the output of the video mixer to be displayed on the A-scope screen. When the positive gate pulse is removed, blanking results (the electron beam is cut off).

Clamping circuits are frequently used with A-scopes to keep the display properly positioned despite changes in the average (de) value of the sweep or signal voltages. Remember, clampers hold one part of the signal waveform at a constant voltage level. In some A-scopes, expanded sweep circuits are used. These circuits enable a small section of the sweep to be expanded to cover the A-scope screen. Thus, more accurate range measurements can be made.

B-SCAN

Often the situation in which a radar is used calls for simplicity of circuitry and construction; there-fore, B-scan is often used.

In B-scan, three variables are possible. These are range (a function of time), azimuth (a function of antenna rotation), and the intelligence received by the radar or associated equipment.

From the operator's standpoint, the ideal situation would be the presentation of an exact replica of the area scanned. This would involve complicated construction and circuitry. The B-scan represents a compromise between the extremes of simple and complex circuitry.

B-scan involves the simplest circuitry and construction of any two-dimensional presentation, and yet presents information as a reasonably faithful replica of the area scanned by the antenna (fig. 6-27). It works best under conditions

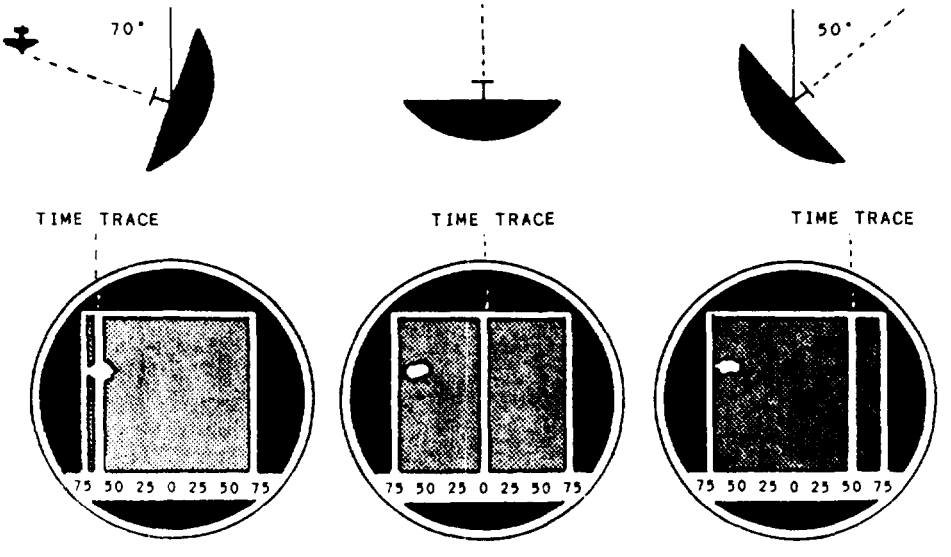


Figure 6-27.-B-scan presentation.

where the antenna scans a sector of less than 180 degrees. It can be employed in a situation where a 360-degree area is scanned.

Range is usually presented vertically by the use of a conventional sweep circuit. (Refer to the B-scope block diagram in figure 6-28.) The scope presentation may be created by either magnetic or electrostatic deflection. Since electrostatic deflection is usually employed, the final amplifiers are operated push-pull to gain the advantage of good sweep linearity.

Azimuth is presented horizontally by the use of a potentiometer that is mechanically connected to the antenna. The output of the potentiometer controls the horizontal push-pull amplifiers, which, in turn, controls the horizontal deflection.

The intelligence is presented on the indicator by intensity-modulating the sweep. The antenna scanning speed is approximately one scan per second, whereas the sweep speed is at the PRF rate; therefore, the intelligence will also have range and bearing.

C-SCOPE

C-scopes (fig. 6-29) are used primarily to present data on the bearing and elevation of targets. C-scopes may sometimes be used in aircraft interception. Like B-scopes, C-scopes provide a rectangular display on their screens. However, in C-scopes, the vertical axis represents elevation and the horizontal axis represents bearing. Thus, in aviation fire control radar, targets may appear on either side of both the horizontal and vertical axes.

To obtain a rectangular display on the screen of a C-scope, both horizontal and vertical-sweep generators are used. Since the sweep frequencies are relatively low, potentiometers (like the azimuth sweep potentiometer of the B-scope) are generally used. These potentiometers are connected to the radar antenna.

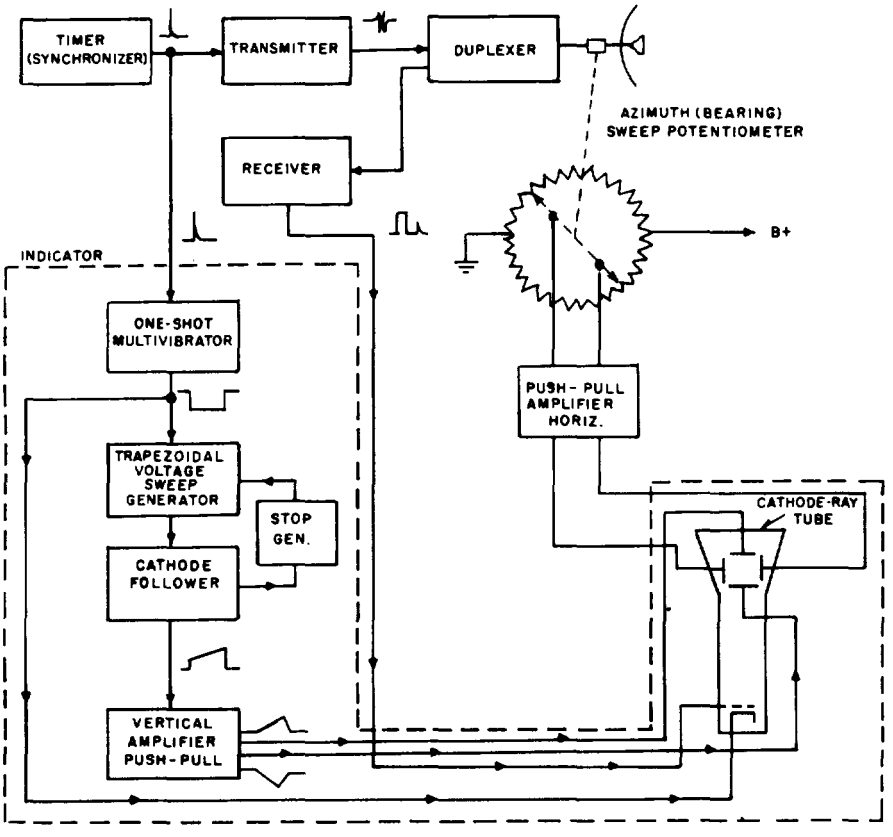


Figure 6-28.-Block diagram of a B-scope system.

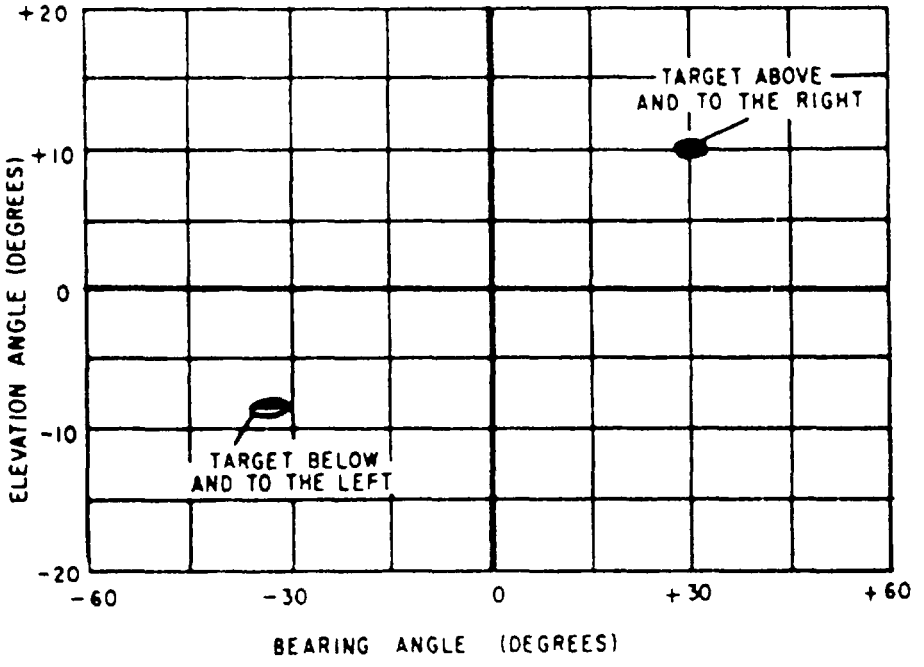


Figure 6-29.-C-scope presentation.

When the antenna turns sideways, the scanning spot on the C-scope screen is deflected horizontally. When the antenna is tilted up or down, the scanning spot is deflected vertically. Echo signals, applied to the control grid (or cathode) of the cathode-ray tube during the sweep period, cause the brightness of portions of the horizontal trace to be increased. The position of a bright spot indicates the elevation and bearing of a target.

Targets at different ranges, but with the same bearing and elevation, appear as a single spot on a C-scope. Targets of this kind cannot be distinguished individually on the C-scope. For this reason, an indicator that presents range data is generally used in conjunction with a C-scope. Once the range of a particular target has been determined, a range gate pulse (rectangular pulse) is applied to the C-scope. This intensifies the electron beam only for the duration of the range gate pulse. Thus, only the desired target echo appears on the C-scope; all other signals are blanked out. In this way, the bearing and elevation of a particular target at a specific range can be determined.

When used in conjunction with modern aviation weapons systems radars, C-scopes will generally be used to display information obtained from an infrared (IR) detector.

PPI-SCOPE

Type-P indicators, also called plan-position indicators (PPI), or PPI-scopes, are used primarily to present data on the range and bearing of targets. Like B- and C-scopes, PPI-scopes generally use cathode-ray tubes with long persistence screens.

The PPI presentation is practically an exact replica of the region scanned by the radar antenna.

Distance along the radial sweep line represents target range. Rotation of the radial sweep line, synchronized with the antenna's rotation, produces a circular display.

When echo signals are applied to the control grid (or cathode) of the PPI CRT during the sweep period, the brightness of portions of the radial sweep line is increased. As in B-scopes, increasing the brightness of some portions of the PPI radial sweep line results

in a maplike picture. Figure 6-30 shows a typical PPI presentation.

Normally, the center of the PPI screen represents the location of the radar. The range and bearing of the target can be determined by means of either an engraved transparent overlay or electronic range circles and an engraved azimuth (bearing) scale. Range circles are usually obtained by adding uniformly spaced pulses to the receiver output signal during the sweep period.

The pulses cause bright spots to appear at equal intervals along the radial sweep line. When

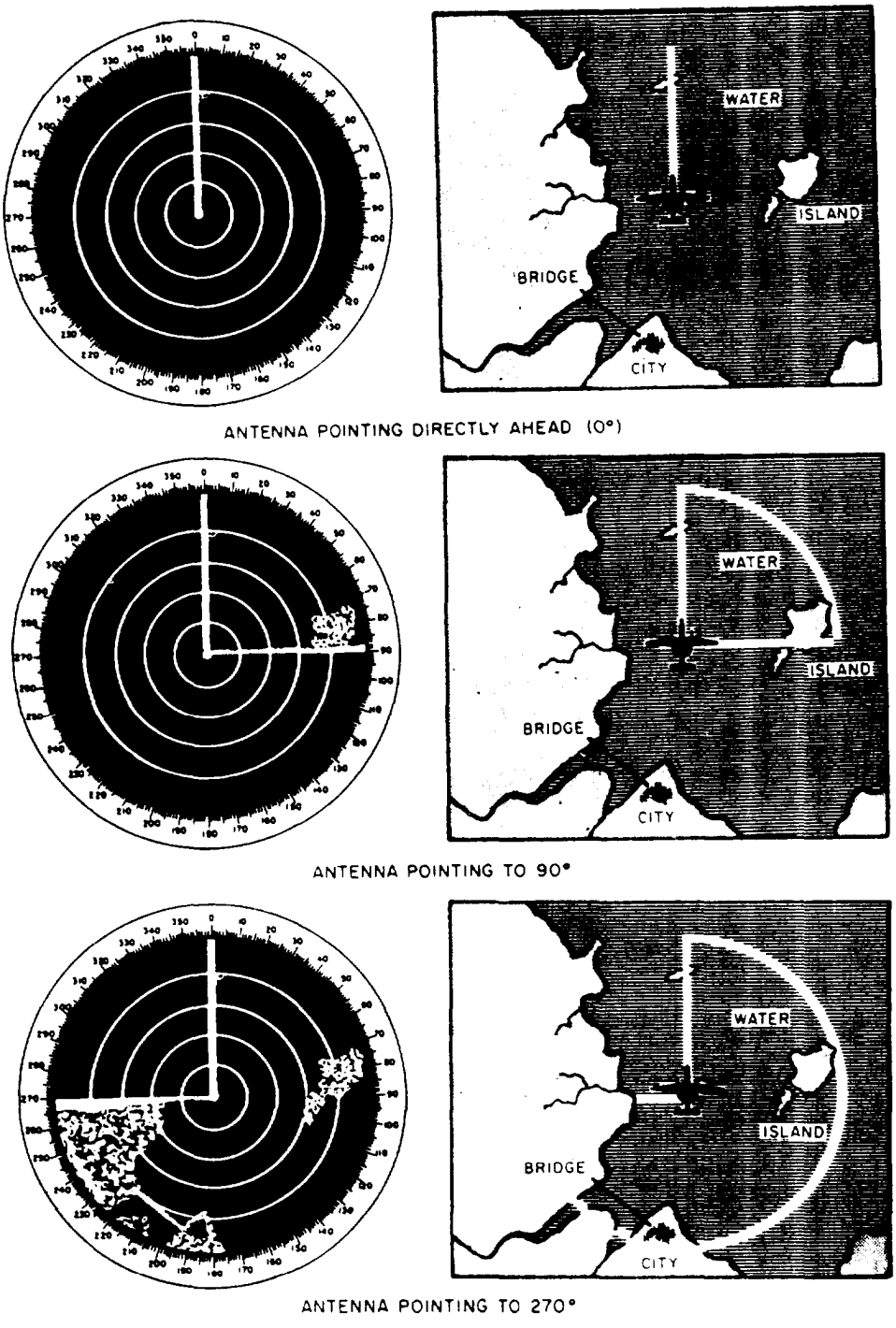


Figure 6-30.-PPI presentation.

the radial sweep line rotates, the spots produce concentric circles. The distance between the center of the PPI screen and a range circle indicates a specific distance.

A movable azimuth index is frequently used to facilitate azimuth (bearing) measurements. The movable azimuth index is scribed on a glass window. The window is rotated by turning a special knob, usually mounted on the PPI chassis. When the window is rotated so that the movable azimuth index coincides with a target echo, the bearing of the target can be determined by noting the position of the index on the azimuth scale.

Figure 6-31 shows a simplified block diagram of a typical PPI-scope. In the case illustrated, the CRT has a fixed deflection yoke. The sawtooth sweep currents required to produce the rotating radial sweep line are obtained from the trapezoidal-voltage sweep generator, a rotary transformer (synchro resolver), and two push-pull amplifiers.

The rotary transformer is a variable-ratio transformer with one primary winding (the rotor) and two secondary windings (the stator). The voltage ratio between the primary winding and each secondary winding changes when the rotor is turned.

The rotor is connected mechanically to the radar antenna. Thus, when the antenna turns, the rotor turns, and the voltage ratio changes. One secondary voltage varies as the sine of the angle of antenna rotation; the other secondary voltage varies as the cosine of the angle of antenna rotation.

The plan-position indicator operates as follows: Trigger pulses from the timer (synchronizer) are fed to both the transmitter and the one-shot monostable multivibrator. The one-shot multivibrator generates negative gate pulses that are applied to the trapezoidal-voltage sweep generator and the cathode of the cathode-ray tube.

The output of the trapezoidal-voltage sweep generator is fed to a power amplifier. The output of the power amplifier is applied to the primary of the rotary transformer. The secondary voltages of the rotary transformer are trapezoidal and have amplitudes that depend on the antenna position.

To apply trapezoidal voltages to the two push-pull amplifiers, a center-tapped resistance network

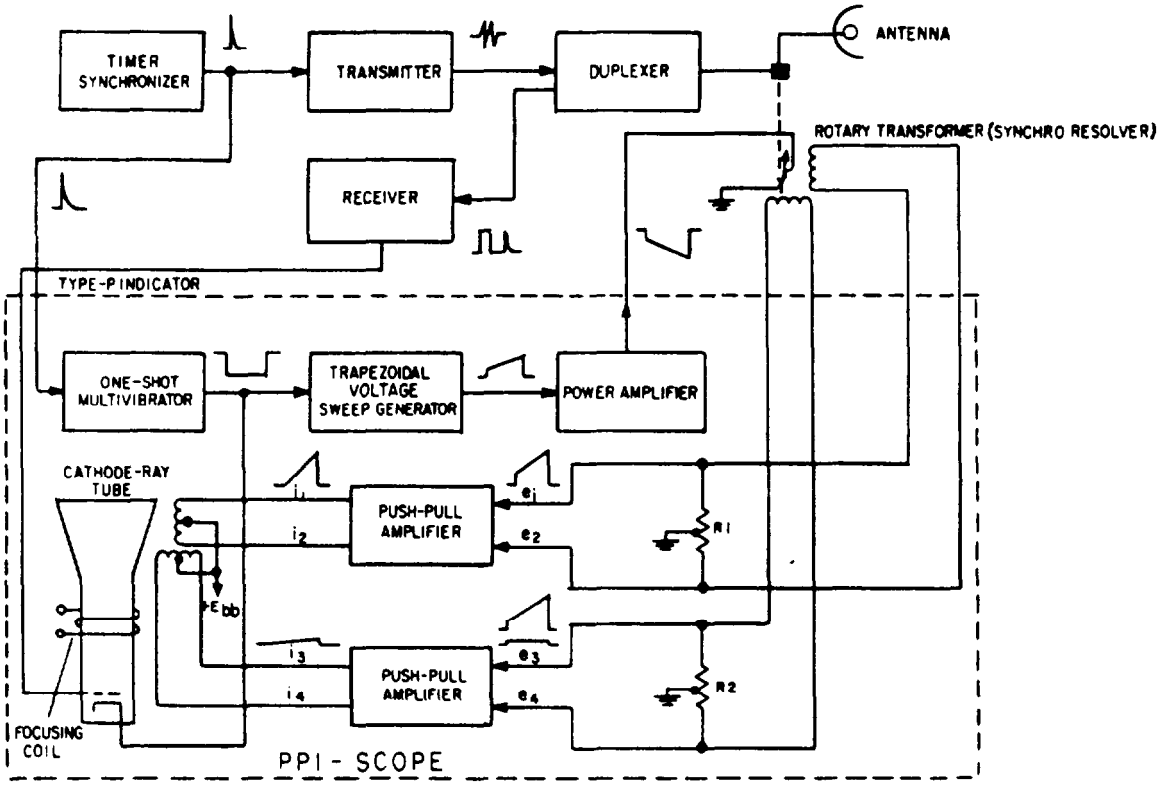


Figure 6-31.-Typical PPI-scope, block diagram.

is connected across each of the two secondary windings of the rotary transformer. Network RI produces two voltages (e_1 and e_2) of equal amplitude and opposite phase, which are applied to one push-pull amplifier. Likewise, network R2 produces two voltages (e_3 and e_4) of equal amplitude and opposite phase, which are applied to the other push-pull amplifier. For simplicity, only the waveforms of e_1 and e_3 are shown in figure 6-31.

Trapezoidal voltages e_1 and e_2 produce sawtooth sweep currents i_1 and i_2, respectively. Trapezoidal voltages e_3 and e_4 produce sawtooth sweep currents i_3 and i_4, respectively. The angular position of the radial sweep line at any instant depends on the relative amplitudes and the phase relationship of the sawtooth sweep currents at that instant.

Sawtooth sweeps current i_1 and i_2 are equal in amplitude, but opposite in phase. Likewise, i_3 and i_4 are equal in amplitude, but opposite in phase. For simplicity, only the waveforms of i_1 and i_3 are shown in figure 6-31. The relative amplitudes and polarities of i_1 and i_3 (also of i_2 and i_4) vary as the rotary transformer is rotated. This causes the radial sweep line to rotate in synchronism with the radar antenna (which is geared to the rotary transformer).

In general, plan-position indicators used in conjunction with rotary transformers also use clamping circuits (clampers). The clamping circuits ensure that the scanning spot always starts from the same point on the PPI screen. The fundamentals of clamping circuits are discussed in Navy Electricity and Electronics Training Series (NEETS), Module 9, *Introduction to Wave Generating and Wave-Shaping Circuits.*

Negative gate pulses are applied to the cathode of the PPI CRT to intensify the electron beam during each sweep period. The electron beam is intensified to the point where the radial sweep line is barely visible. When echo signals are applied to the control grid of the CRT during each sweep period, the brightness of portions of the radial sweep line is increased.

E-SCAN (RHI)

The RHI (range height indicator) presentation (fig. 6-32) is another type of scan for presenting range and height information. The RHI is also known as the E-scan. E-scan is a modification of the B-scan, on which an echo appears as a bright spot with range indicated by the horizontal coordinate and the elevation (height) as the vertical coordinate. This type of scan is used in directing aircraft during ground- and carrier-controlled approaches, and in fire control systems for terrain clearance.

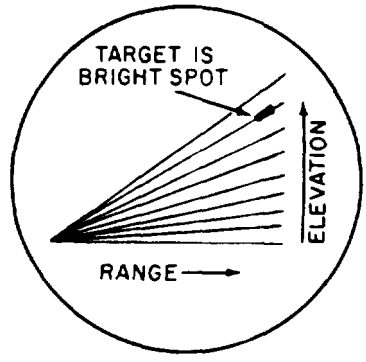


Figure 6-32.-E-scan presentation.

DEFLECTIONS AND SWEEPS

The various types of deflection and sweep circuitry required for a typical radar indicator system are shown in figure 6-33, view A. These circuits, together with the receiver circuits, are contained in a unit called the receiver-indicator. Figure 6-33, view B, shows the waveforms for this indicator system when set for a 30-mile range sweep.

Negative trigger pulses from the timer are applied to the sweep multivibrator. When a negative trigger pulse is applied to it, the sweep multivibrator causes a switching action that initiates the sweep voltage in the PPI sweep generator.

When the sweep voltage reaches a predetermined level, the sweep-limiter circuit conducts and causes a second switching action. As a result, the sweep multivibrator produces positive and negative rectangular pulses, as shown in figure 6-33, view B. The positive rectangular pulses are applied to the PPI sweep generator, which produces a trapezoidal voltage.

The trapezoidal voltage is amplified by the PPI sweep amplifiers, and then applied to the synchro driver (power amplifier). The output of

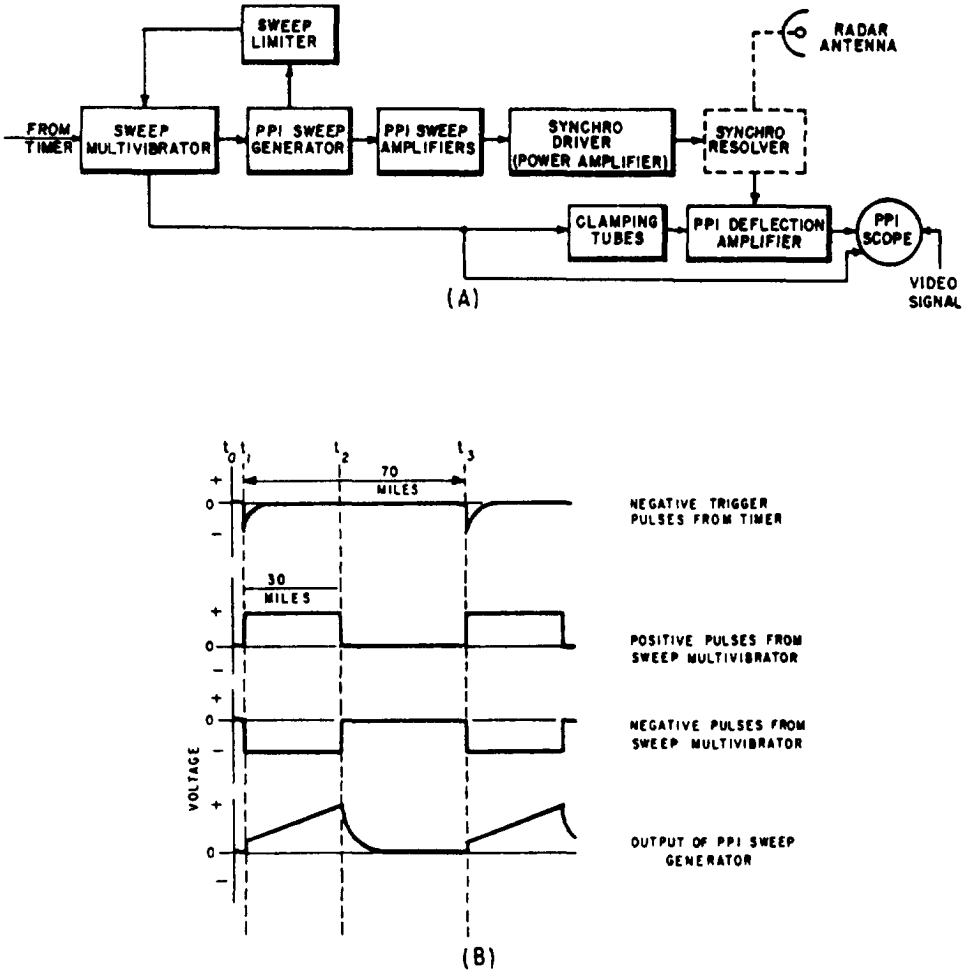


Figure 6-33.-(A) Typical indicator, block diagram; (B) waveforms 30-mile sweep.

the synchro driver is applied to the rotor of a synchro resolver.

The synchro resolver has two stator windings and one rotor winding. The rotor shaft is mechanically coupled to the radar antenna. When the antenna rotates, the synchro-resolver rotor turns and the trapezoidal sweep voltages of varying amplitude are induced in the synchro-resolver stator windings. The amplitude of the induced sweep voltages, at any instant, depends on the angular position of the rotor winding with respect to the stator windings of the synchro resolver.

The induced sweep voltages are amplified by the PPI deflection amplifiers, and then applied to the deflection coils of the PPI-scope. As a result, a radial sweep line is produced on the PPI screen. Because of the mechanical coupling between the radar antenna and the synchro-resolver shaft, the radial sweep rotates in synchronization with the antenna.

The negative rectangular pulses from the sweep multivibrator are applied to the clamping tubes and the cathode-ray tube of the PPI-scope. The clamping tubes ensure that the PPI electron beam always starts its sweep trace from the same point on the PPI screen.

The negative rectangular pulses applied to the CRT intensify the scope electron beams during the sweep time. This produces a faintly visible trace on the scope.

The radial sweep line on the screen of a PPI is obtained by sweeping the electron beam from the center to the edge of the PPI screen. To obtain azimuth (bearing) indication, the radial sweep line is rotated about the center of the screen, like a spoke of a wheel. Two different methods that are used to produce a rotating radial sweep

line are mechanical azimuth sweep and electrical azimuth sweep.

In mechanical azimuth sweep (fig. 6-34, view A), a rotating deflection yoke is used. The yoke is mounted in bearings, and is rotated mechanically around the neck of the CRT, in synchronization with the radar antenna.

When a sawtooth sweep current is applied to the deflection coil through the slip rings, as shown in figure 6-34, view B, a sawtooth magnetic field is produced. The magnetic field causes the electron beam to be deflected, thereby producing a radial sweep line. The amount of beam deflection, and consequently, the length of the radial sweep line, depend on the amplitude of the sawtooth sweep current.

The direction of beam deflection, and consequently, the position of the radial sweep line, are always perpendicular to the magnetic field. Thus, when the deflection yoke rotates, the deflection coil rotates, and the magnetic field also rotates.

Since the electron beam is always deflected in a direction perpendicular to the magnetic field, the radial sweep line also rotates. A radial sweep line is produced each time a sawtooth sweep current is applied to the deflection coil. The radar antenna and the rotating yoke are always rotated in synchronism.

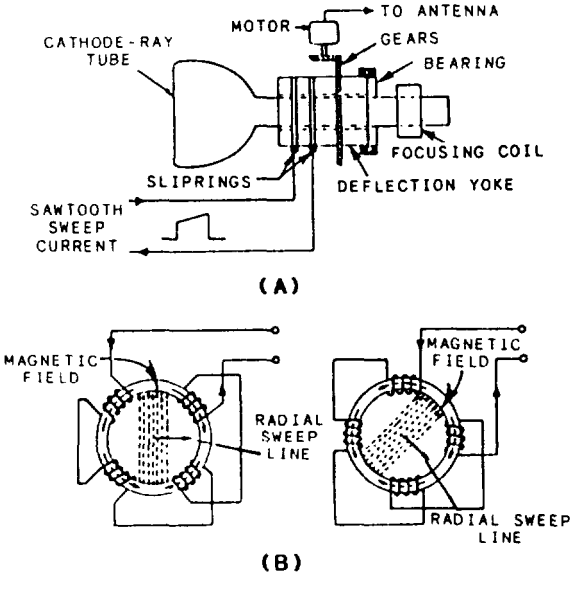


Figure 6-34.-Mechanical azimuth sweep. (A) Rotating deflection yoke; (B) relationship between sweep direction and yoke position.

The second method of producing a rotating radial sweep line is called electrical azimuth sweep. This system uses a fixed deflection yoke with horizontal and vertical deflection coils. Sawtooth sweep currents are applied simultaneously to the two deflection coils. The direction of the radial sweep line obtained depends on the relative amplitudes of the two sweep currents, Unlike mechanical azimuth sweep, electrical azimuth sweep has practically no mechanical difficulties.

MISCELLANEOUS PRESENTATIONS

Many other types of radar indicators are used in addition to A-scopes, B-scopes, C-scopes, and PPI-scopes. It is not unusual to find more than one type of presentation incorporated into one indicator.

Most indicators used in aviation fire control radar use two or more electron guns. One gun is used to develop a B-type presentation, while the other gun or guns develop the various elements of a so-called attack presentation.

These elements may consist of an elevation strobe, artificial horizon, steering information, acquisition circle, and range circle. The range circle element is discussed in the following text.

RANGE CIRCLE GENERATOR

The indicator of a typical aircraft weapons systems radar uses a circle for range presentation. The size of this circle will be directly proportional to the target range. Figure 6-35 is a simplified schematic diagram of a range circle generator. You should already be familiar with the method of producing a circle on a CRT by applying two sine waves, 90 degrees out of phase, to the deflection plates of the CRT. The size of the circle is controlled by the amplitude of the sine-wave inputs.

The range circle generator (fig. 6-35) provides two sine-wave signal outputs 90 degrees out of

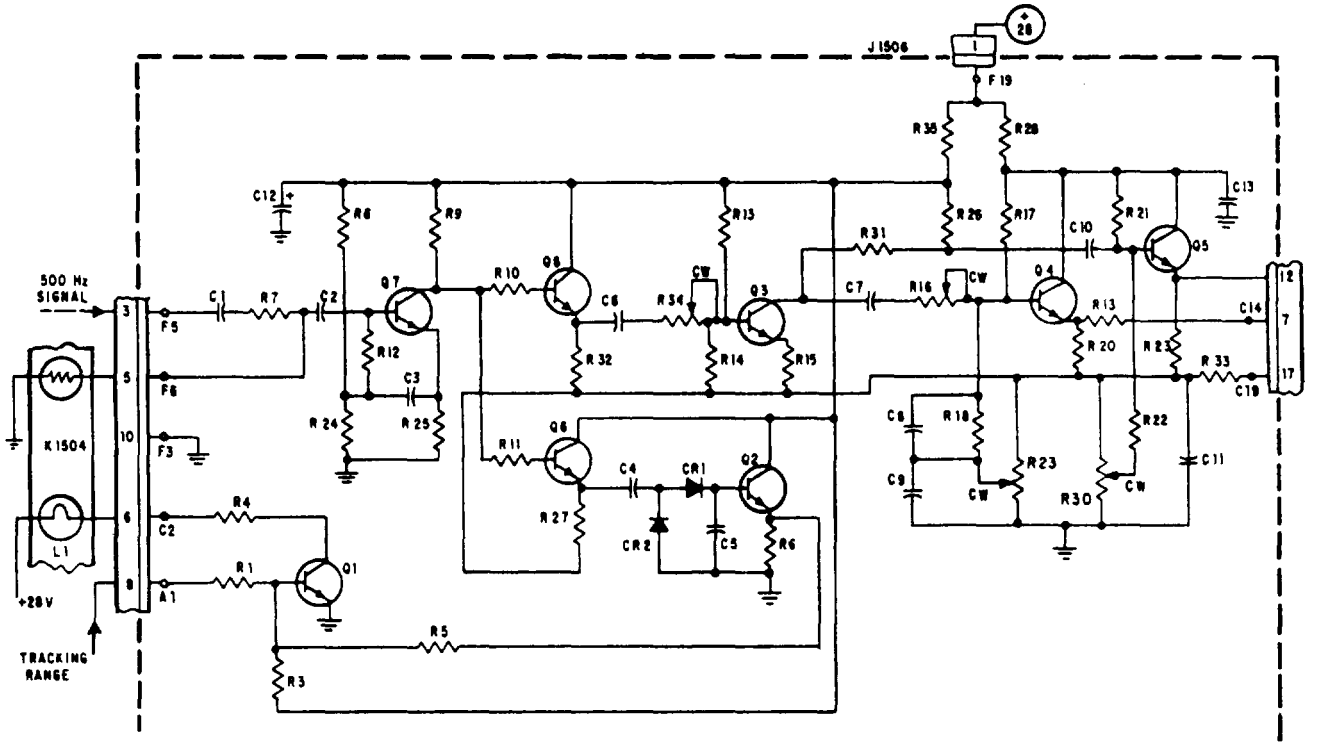


Figure 6-35.-Range circle generator.

phase with each other. The amplitudes of each signal are controlled by a dc range signal from the tracking unit. Inputs to the circle generator are a 500-Hz reference signal and the dc range voltage from the tracking unit. The 500-Hz signal is applied to amplifier transistor Q7 through capacitors C1 and C2 and resistor R7. Resistor R7 is part of a voltage divider consisting of R7 and a resistor contained in raysistor K1504. A raysistor is a series of switching devices, the output consisting of a photoresistor that may be excited by various forms of electrical lights. Resistance of the raysistor resistor varies with the intensity of the raysistor light L1.

Output for transistor Q7 is applied to signal emitter follower Q8 and to feedback emitter follower Q6. The output of Q6 is rectified by the action of diodes CR1 and CR2 to obtain a positive voltage whose amplitude is proportional to the ac signal. This positive voltage is applied through emitter follower Q2 and resistor R5 to transistor Q1, Also applied to transistor Q1 through resistor R1, is the dc range voltage. The resulting effect is a voltage on the base of Q1 that is proportional to the dc range voltage and the dc feedback voltage. The output of Q1 varies the intensity of L1, which, in turn, varies the resistance of the raysistor resistor. Therefore, the resistance of K1504 is proportional to the output of Q1. This directly controls the amplitude of the output of Q7, since it is part of its input voltage divider. Thus, the output amplitude of Q7 is controlled by the dc tracking range input voltage to Q1. The output of Q7 is applied to emitter follower Q8, then on to buffer amplifier Q3 through potentiometer R34, which ultimately adjusts the diameter of the displayed circle.

Two outputs are taken from transistor Q3. One is applied through a lagging phase shift network consisting of potentiometer R16 and capacitor C8 to emitter follower transistor Q4; the other is applied through leading phase shift capacitor C10 to emitter follower transistor Q5. Thus, the outputs of Q4 and Q5 are 500-Hz sine-wave voltages phase-shifted by 90 degrees with their amplitudes controlled by the dc range voltage. Potentiometer R16 provides a fine adjustment of the phase shift. These two outputs,

90 degrees out of phase, are applied to the deflection amplifiers to produce a circular trace on the CRT.

IFF SYSTEM

Learning Objective: *Identify the operating principles and characteristics of an IFF system, to include interrogation modes.*

With the destructive power of modern weapon systems and the speed of modern weapon delivery systems, it is not practical to wait until a detected radar target is identified by visual means before preparing for battle. Therefore, a means of identifying friendly targets from enemy targets at long range is required. It was this need that brought about the advent of IFF. The IFF system permits a friendly craft to identify itself automatically (by means of an IFF transponder) when interrogated (by means of an IFF interrogator) by either a ground station or another craft. The average operating range of IFF is in excess of 300 miles.

Additionally, due to the high density of air traffic, a special operating mode of the IFF enables an aircraft to identify itself (by special codes) to ground stations equipped with an Air Traffic Control Radar Beacon System (ATCRBS). Still another IFF mode of operation enables an aircraft to automatically report its altitude to an ATCRBS ground station.

A typical IFF system is shown in figure 6-36. It consists of an interrogator unit, a coder synchronizer unit, a search radar unit, and a transponder unit. The interrogator, synchronizer, and radar units comprise the challenging station, and the transponder unit is the responder station. It should be noted that the challenging station can be a ground station, a ship, or another aircraft. The responder station is normally an aircraft.

INTERROGATION (CHALLENGE)

The interrogator is a pulse-type transmitter, which is triggered by the coder synchronizer. The coder synchronizer is synchronized to the radar system (IFF challenges are transmitted 1 to 40 μsec after the radar transmitter pulse, depending on the particular system) so that reception of the IFF response and radar echo signals cannot occur simultaneously. The output (challenge signal) of

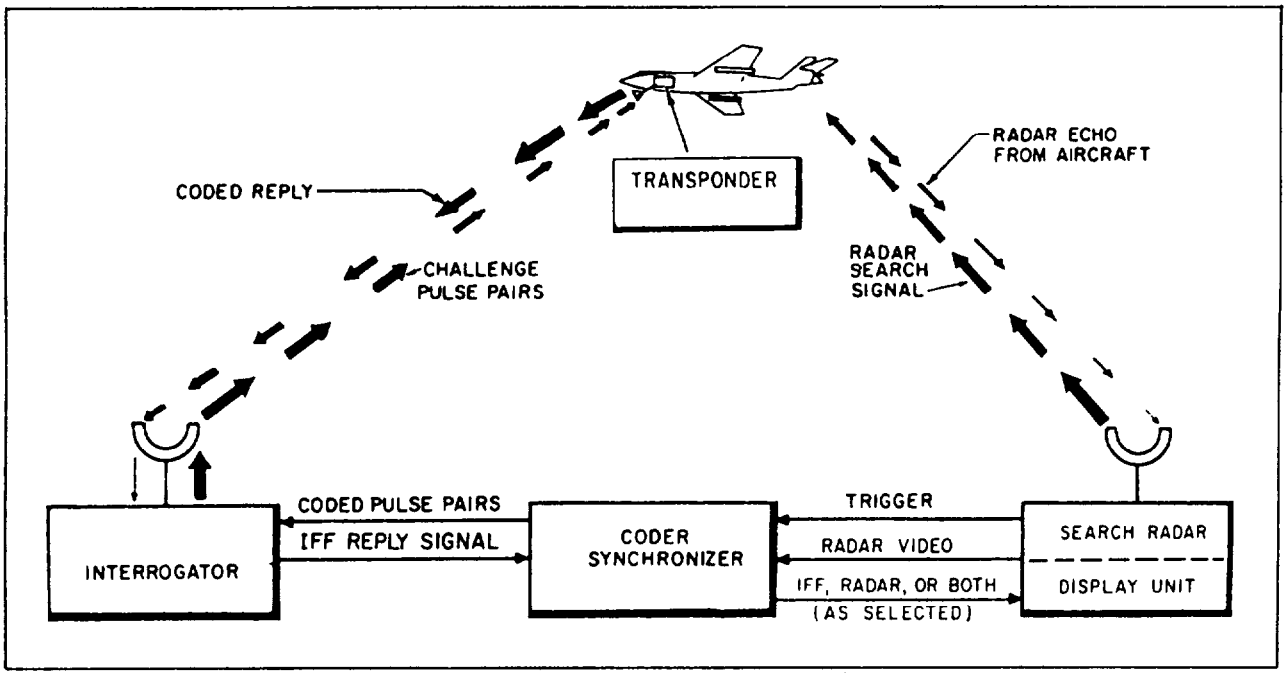


Figure 6-36.-IFF system block diagram.

the interrogator is different for different modes of IFF operation. (These modes will be explained later in this chapter.) As long as the aircraft transponder is operating in the correct mode, it will receive these interrogation signals and will transmit back to the interrogator the proper coded reply pulses. The IFF reply pulses (fig. 6-36) are sent via the coder synchronizer to the radar PPI indicator. These reply pulses will appear as dashed lines just behind the aircraft target on the scope. Figure 6-37 shows how these reply signals appear.

Interrogation Modes and Codes

There are presently five modes of IFF operation used by ATCRBS and naval aircraft. They are designated Mode 1, Mode 2, Mode 3/A, Mode C, and Mode 4. In addition, there is a test mode of operation used only by the aircraft transponder as a self-check of the transponder equipment.

Modes 1 and 2 are for exclusive use by the military as tactical modes for target identification. Mode 3/A is used mainly by military and civilian air traffic control stations. Mode C is used in conjunction with an external pressure altitude digitizer to report the aircraft's altitude to an ATCRBS. Mode 4 is a military encrypted mode, which is controlled by an external computer. The operation of Mode 4 is classified. Only interrogators and transponders using the same encrypted codes can communicate with each other.

Interrogation Pulse Characteristics

The interrogation pulse characteristics for the various modes of IFF operation are shown in

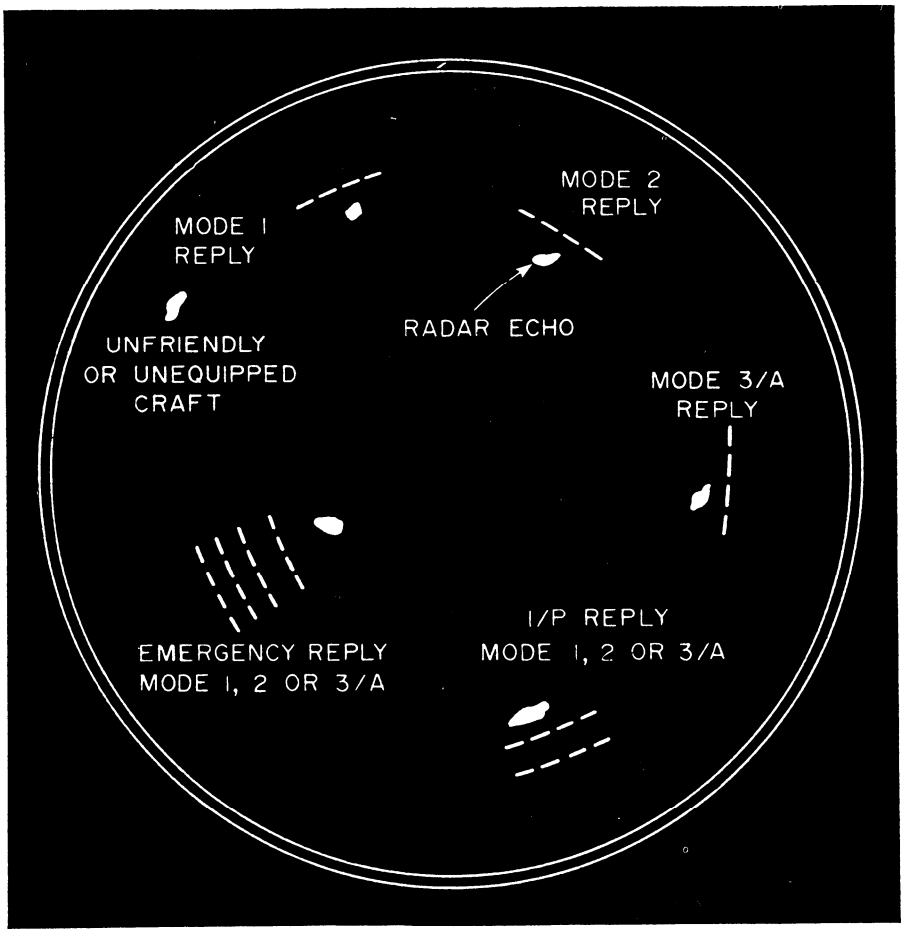


Figure 6-37.—Typical radar PPI composite display showing several IFF responses.

figure 6-38. These pulses are transmitted at a frequency of 1030 MHz, and are recognized by the transponder through pulsewidth and spacing. Modes 1, 2, 3/A, C, and test each use two interrogation and one side-lobe suppression (SLS) pulse 0.8 μsec wide. The pulse spacing is different for each mode of operation. Note (fig. 6-38) the pulse spacing for Modes 1, 2, 3/A, C, and test are 3, 5, 8, and 7 μsec, respectively. The side-lobe suppression pulse occurs 2 μsec after the leading edge of the first interrogation pulse in each case. The SLS pulse is used by the transponder receiver circuits to prevent jamming of the IFF system and to suppress the receiver during operation of TACAN, or other L-Band equipment operating in the vicinity of the transponder.

Mode 4 interrogation pulses consist of four pulses 0.5 μsec wide referenced from the leading edge of the first pulse, in multiples of two. (This is determined by the external equipment used to encrypt Mode 4.) The four pulses may be followed by as many as 32 additional pulses spaced as close as 2 μsec apart. The SLS pulse for Mode 4 is spaced 8 μsec from the leading edge of the first interrogation pulse used.

TRANSPONDER (REPLY)

In the early days of IFF, the transponder of the aircraft being interrogated would receive the interrogation pulses (either 3, 5, or 8 μsec spacing depending on mode of operation), and would automatically respond with reply pulses of the same spacing. But with the increase of air traffic, a more positive means of identification was required. This brought about the advent of the Selective Identification Feature (SIF). This feature has led to the common usage of the designation IFF/SIF system, instead of just the nomenclature IFF system. With the present IFF/SIF system, there are 32 separate reply codes that can be transmitted on Mode 1, and 4,096 separate reply codes that can be transmitted on Mode 2 and on Mode 3/A. Mode C has 1,024 separate reply codes available.

Transponder Control Panel

The operation of the interrogated aircraft's transponder (mode of operation and specific codes, etc.) is controlled by a control panel similar to the one shown in figure 6-39, view A,

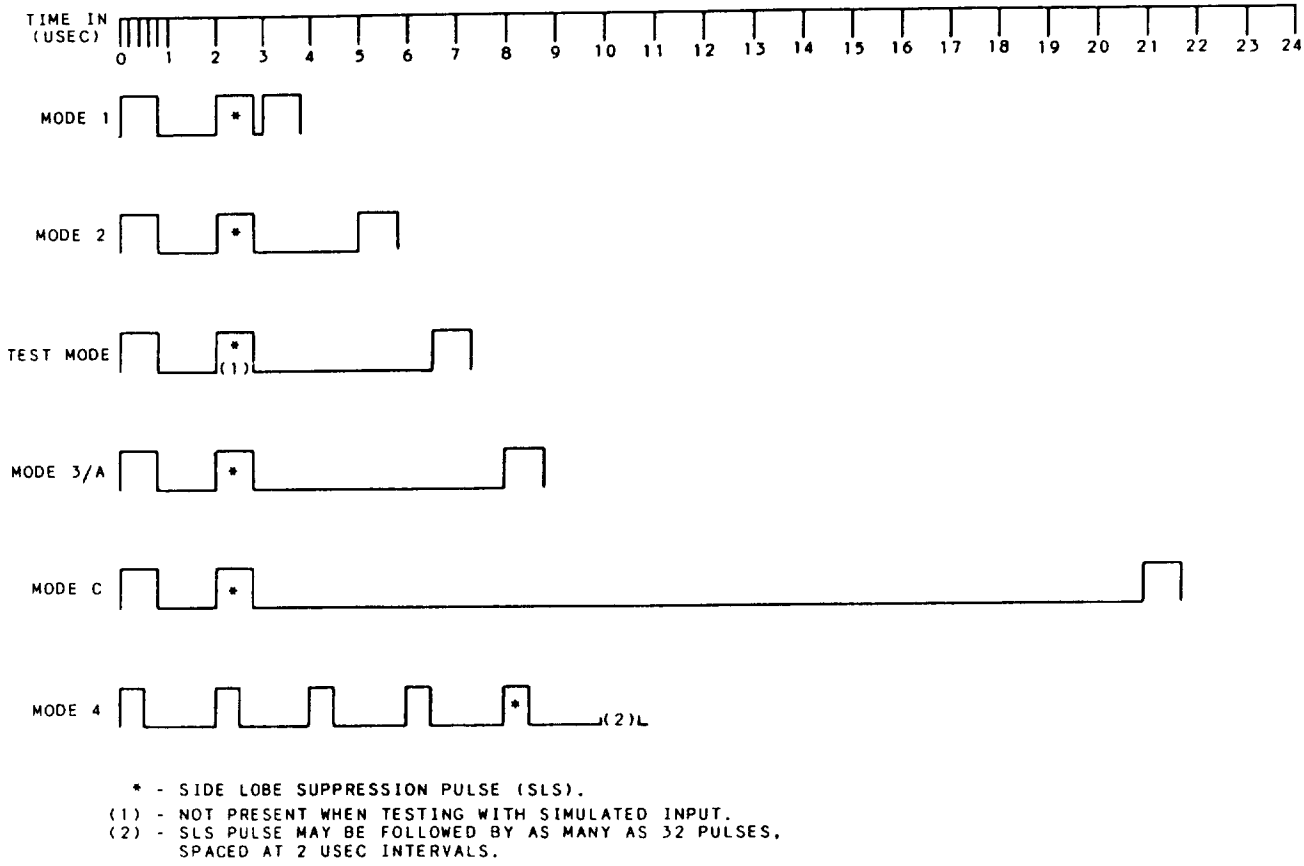


Figure 6-38.-IFF interrogate pulse characteristics.

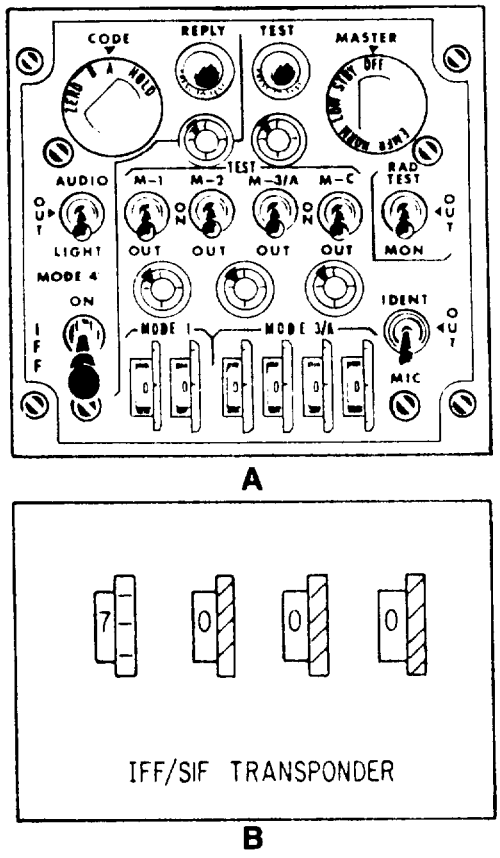


Figure 6-39.-(A) Typical IFF/SIF transponder control box; (B) IFF/SIF transponder front panel Mode 2 control dials.

which is normally located on the pilot's control panel.

MASTER Control Switch

The MASTER control switch is a five-position rotary switch placarded OFF, STBY, LOW, NORM, and EMER. In the STBY position, power is applied to the IFF coder, but interrogations are inhibited. In the LOW position, the IFF coder is operational, but the receiver sensitivity is reduced. In the NORM position, the IFF coder is fully operational at normal receiver sensitivity. In the EMER position, the IFF coder transmits emergency replies to interrogations in Modes 1, 2, or 3/A. The Mode 3/A emergency reply includes code 7700. With EMER selected, Mode 4 is enabled regardless of the position of the Mode 4 switch, and Mode C continues to function normally if selected. To select the EMER position, the control is pulled outward and rotated to the EMER position.

IDENT/OUT/MIC Switch

The IDENT/OUT/MIC switch is a three-position toggle switch. Momentarily actuated (the switch has spring-loaded return) to the IDENT position, the IFF coder adds an identification of position response to Modes 1, 2, and 3/A, for 15 to 30 seconds. In the MIC position, the identification of position function is activated for 15 to 30 seconds whenever the microphone switch is actuated. In the OUT position, the IDENT reply is disabled.

Mode 1-3A Code Selectors

Six code-selector switches are provided for selection of Mode 1 and Mode 3 codes. Mode 1 has two thumbwheel selectors, which allow selection of 32 different codes. Mode 3 has four thumbwheel selectors that provide the capacity of selecting 4,096 codes.

Mode-Selector Test Switches and Light

Four mode-selector/test-selector switches labeled "M-l," "M-2," "M-3/A," and "M-C" have TEST, ON, and OUT positions. In the momentary TEST position with the MASTER switch at NORM, lighting of the TEST light indicates proper operation of the mode selected. The mode switches for the modes not being tested should be OUT when testing on the ground to prevent unnecessary interference with nearby ground stations.

NOTE

The TEST light may flash once as each mode switch is released from TEST, and as the RAD TEST/MON switch is moved. This flash has no significance.

The OUT position for each switch disables its respective mode. The ON position for each switch enables the transponder to reply to interrogations for the mode selected. The M-C switch provides automatic coded altitude reporting in response to a ground station's interrogation for air traffic control identification.

RAD TEST/MON Switch

The RAD TEST/MON switch is a three-position toggle switch spring-loaded to the OUT (center) position. When MON is selected, the TEST light comes on for 3 seconds each time an acceptable response is made to an interrogation in Modes 1, 2, 3/A, and C. When RAD TEST is selected, Mode 3/A or Mode 4 responds to TEST mode interrogations from a ramp test set during ground maintenance testing.

Mode 4 Operation

The Mode 4 controls and indicator light are grouped on the left side of the control panel. The MASTER rotary switch controls transponder operation in Mode 4 as well as in the other modes of operation. Mode 4 will operate normally, when selected, in either the NORM or EMER position, and at reduced receiver sensitivity in the LOW position. The Mode 4 function will be inoperative in either the STBY or OFF position. With the transponder functioning, Mode 4 operation is selected by placing the Mode 4 ON-OUT toggle switch to ON. Placing the switch to OUT disables Mode 4 operation.

The Mode 4 CODE control selects either of the two (A or B) Mode 4 codes. It has two additional positions, HOLD and ZERO. The switch is spring loaded to return from HOLD to the A position. At a designated daily time and/or prior to the day's mission, maintenance or operations personnel mechanically set in the Mode 4 code for the present code period in position A, and the code for the succeeding code period in position B, with a single insertion of the KIK-18/TSEC code changer key. Both code settings will automatically zeroize when power is turned off or lost after the landing gear has been retracted (that is, after initial takeoff). The code settings can be retained by activating the HOLD function. This will normally be done after the aircraft has landed (landing gear must be down and locked), and before power is removed from the transponder. Place the Mode 4 CODE control to HOLD, and release. Allow transponder power to remain on for at least 15 seconds, and then turn it off. The code setting is now mechanically latched and will be retained when aircraft power is turned off.

NOTE

If power is removed from the transponder less than 15 seconds after selecting HOLD, either by turning the transponder off or by turning off aircraft electrical power, the code setting may not be mechanically latched and will zeroize.

Mode 4 settings should be manually zeroized if the aircraft does not make an initial takeoff (that is, gear was never raised), or if the hold function has been engaged and a subsequent takeoff was not carried out.

With aircraft power on and the MASTER rotary switch in any position except OFF, both code settings can be zeroized at any time by placing the CODE switch to the ZERO position. Both code settings will also be zeroized if the HOLD function has not been properly actuated before the MASTER switch is turned to OFF. (Inadvertent selection of OFF is prevented by switch design, which requires that the rotary knob be pulled out before it can be turned to OFF.) When the CODE switch is placed in the A position, the aircraft transponder will respond to Mode 4 interrogations from an interrogator using the same code setting as that set into the aircraft's code A position. In the B position, interrogations from an interrogator using the same code setting as that set into the aircraft's code B position will be answered. The changeover time from code A to code B use is operationally directed.

The Mode 4 AUDIO-LIGHT switch selects the aircraft indication for Mode 4 replies. In the LIGHT position, the Mode 4 REPLY lamp (green) will light when Mode 4 replies are transmitted. In the AUDIO position, an audio signal in the pilot's headset indicates Mode 4 interrogations are being received, and lighting of the Mode 4 REPLY light indicates when replies are transmitted. Mode 4 audio volume can be adjusted by the appropriate aircraft intercom audio volume control. In the OUT position, both light and audio indications are inoperative.

Mode 2 Code Selectors

The code control dials (similar to Mode 3/A) for Mode 2 are located on the front of the IFF/SIF transponder, as illustrated in figure 6-39, view B. The Mode 2 reply code setting is usually established and dialed in by the AT prior to flight. As with Mode 3/A, the four thumbwheels provide for 4,096 available codes. The Mode 2 reply code normally cannot be changed in flight due to the inaccessibility of the transponder unit.

Transponder Normal Reply Modes and Codes

During IFF/SIF interrogation, the interrogating station will notify the pilot (via UHF or VHF communications) to set the transponder to a specific mode (1, 2, 3/A, and/or C or 4) of operation, to undergo interrogation (identification), and in the case of Modes 1 and 3/A, will direct the pilot to set a specific reply code. (Mode C and Mode 4 codes are automatically controlled by external equipments, and Mode 2 is preset.)

Figure 6-40 illustrates the normal reply pulse characteristics of the IFF/SIF transponder. These reply pulses are transmitted at 1090 MHz, between

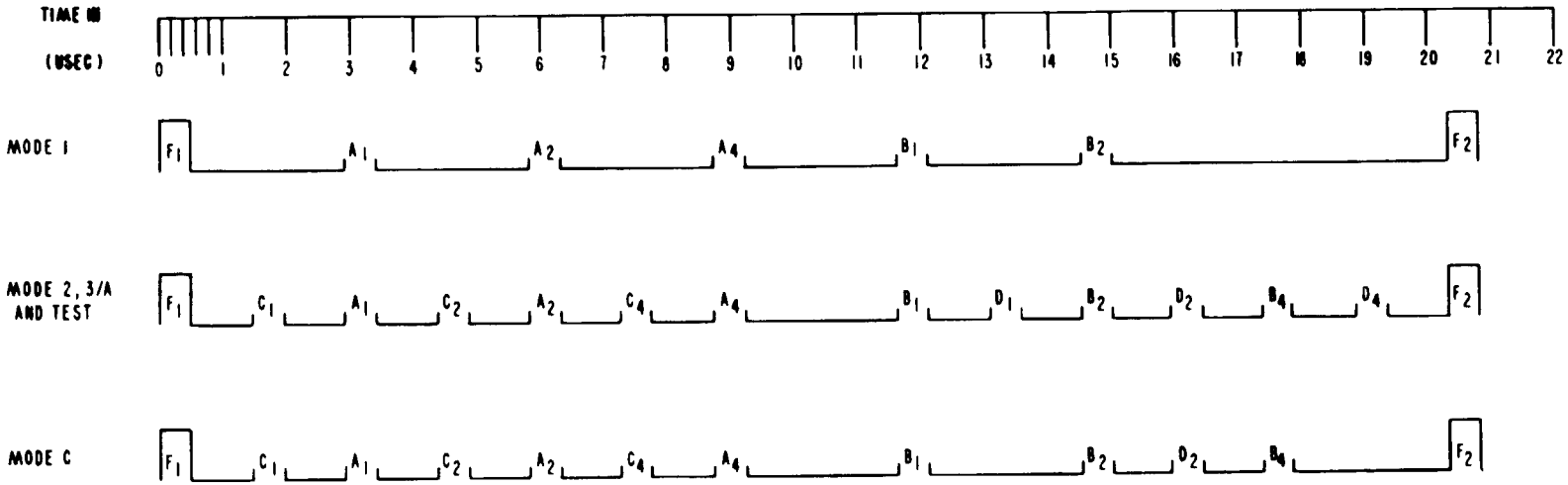


Figure 6-40.—Normal reply pulse characteristics.

two framing pulses labeled "F1"' and "F2." The interrogator transmits on 1030 MHz and receives on 1090 MHz; the transponder receives on 1030 MHz and transmits (replies) on 1090 MHz. Refer to figures 6-39 and 6-40 during the following discussion.

When the transponder processes a Mode 1 challenge, a reply pulse train is transmitted containing two framing pulses 20.3 µsec apart, plus 0 to 5 information pulses (dependent upon the Mode 1 control dial settings). The information pulses are spaced in multiples of 2.9 µsec starting from the initial framing pulse. The position where a sixth pulse (17.4 µsec) would appear is not used. The information pulses are designated by a letter, with a number subscript, to identify a specific reply pulse. The subscript numbers are constructed in binary form (1, 2, 4, and so on). In the case of Mode 1, a maximum of three A pulses are possible (A, A_2, and A_4) and a maximum of two B pulses (B_1 and B_2) are available. These reply codes (pulses) are set by the two control dials (M-1) on the control box. The left dial is used for setting the A reply pulses and the right dial is for the B pulses. In figure 6-40, all five of the information pulses are present in the reply train. This corresponds to the maximum dial setting of Mode 1, which is 73. [The left (A) dial has 8 settings, 0 through 7; and the right (B) dial has 4 settings, 0 through 3; which gives a maximum of 32 possible codes available.] The way to construct a specific reply pulse train is governed by the addition of the particular pulse (such as A) subscript numbers. As an example, if 0 were selected on the left M-1 (A) dial, no A pulses would be present in the reply train. If 1 were selected, only the A_1 pulse would be present in the reply train. If 3 were selected, the A_1 and A_2 (by adding subscript numbers you get 3) would be present, and so on. Keep in mind that any number (0 through 7) selected on the A control dial would read out in binary form in the A pulses subscript numbers. So, for figure 6-40 Mode 1 reply pulse train, with A_1, A_2, and A_4 pulses present, the dial setting is 7 (1 + 2 + 4 = 7).

When the transponder processes either Mode 2, 3/A, or test challenges, a reply pulse train is transmitted containing two framing pulses 20.3 µsec apart, plus 0 to 12 information pulses (dependent upon the respective dial settings for each mode). The information pulses are spaced in multiples of 1.45 µsec apart, starting from the initial framing pulse. The position where the seventh pulse (10.15 µsec) would appear is normally not used. (This position, called the X-pulse position, will be explained later in this chapter.) As in Mode 1, the 12 information pulses are designated by a letter and subscript numbers. Since Modes 2 and 3/A have four code dials (labeled M-3 on the control box and M-2 on the front of the transponder), there are four possible reply pulses that can be set. They are labeled "A," "B," "C," and "D," and correspond to the four control dials from left to right. The four control dials (for Modes 2 and 3/A) have 8 settings each (0 through 7) for a total of 4,096 available codes (8 x 8 x 8 x 8 = 4,096). As in Mode 1, the various pulse subscript numbers are in binary form of a particular pulse dial setting. In figure 6-40, all 12 information pulses are present in the reply train, signifying a dial setting of 7777. (Add A subscript, B subscript numbers, etc., to obtain the dial reading.) If the dial setting for Mode 3/A were 1, 2, 3,4, then the pulses present in the reply train would be A_1, B_2, C_1, C_3, and D_4. Remember, the dial setting will read out in binary form in the reply pulse subscript numbers.

The Mode C function of the transponder (altitude reporting) is often referred to as AIMS. When this function was first introduced, it was believed that AIMS was an acronym for Altitude Information Monitoring System. However, this is not true. AIMS is actually an acronym made up of other acronyms; A . . . . ATCRBS, I . . . . I F F , M . . . . MARK XII Identification System, S . . . . System.

When the transponder processes a Mode C challenge, a reply pulse train is transmitted containing two framing pulses 20.3 µsec apart, plus 10 information pulses (as determined by an external pressure altitude digitizer based on the aircraft's pressure altimeter reading). The information pulses are spaced 1.45 µsec apart from the initial framing pulse. The positions where the 7th (10.15 µsec), 9th (13.05 µsec) and 11th (18.85 µsec) pulses would appear are not used. From the 10 information pulses, a total of 1,024 codes are available (8 x 8 x 8 x 2). Note that (fig, 6-40) in Modes 2, 3/A and C, the pulse positions are not in sequence (A, B, C, etc.) as they are in Mode 1. However, reading the subscript numbers for A pulses will give you the A dial setting, B for the B dial setting, and so on.

As previously stated, Mode 4 reply pulse coding is accomplished by external crypto equipment and is classified information. This information can be found in technical manuals for the KIT-1A/TSEC equipment.

Transponder Special Reply Functions

The special reply pulse characteristics for Modes 1, 2, and 3/A are shown in figure 6-41.

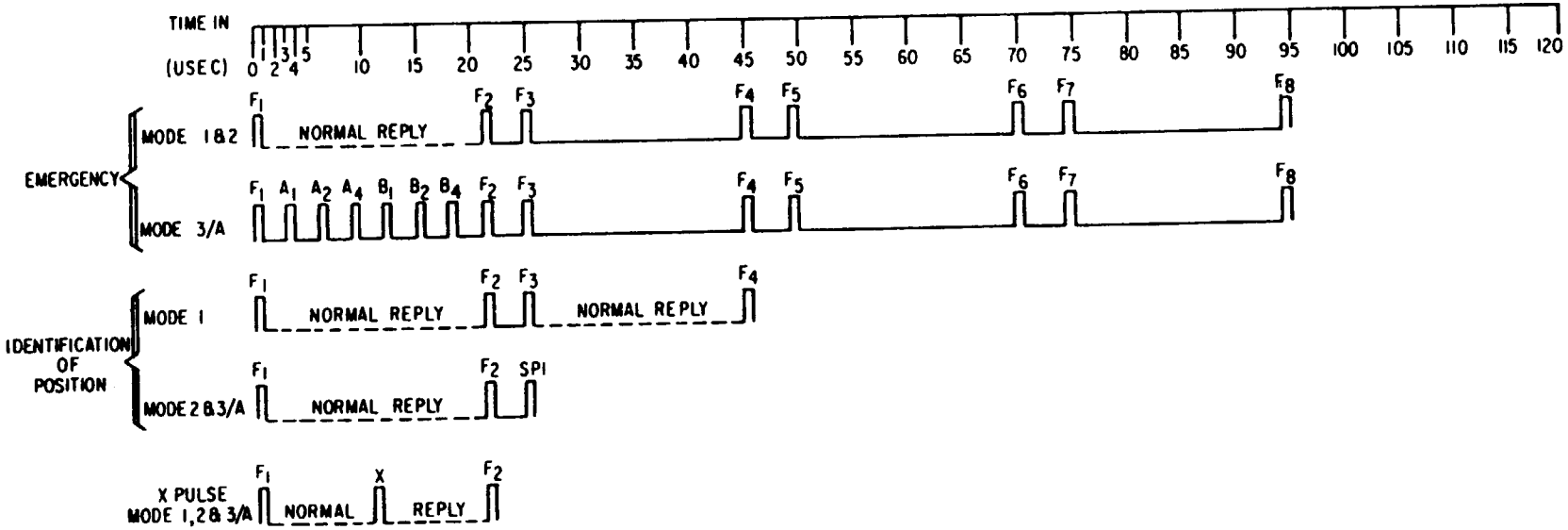


Figure 6-41.—Special reply pulse characteristics.

Modes C and 4 are not affected by the special reply functions.

Identification of Position (I/P)

The I/P function is controlled by the IDENT/ OUT/MIC (fig. 6-39, view A) switch and affects the operation of Modes 1, 2, and 3/A. In Mode 1, the reply pulse train containing the code in use is transmitted twice, instead of once, for each interrogation pulse received. The second reply pulse train is spaced 24.65 μsec from the leading edge of the first framing pulse in the first train. In Modes 2 and 3/A, the reply pulse train containing the code in use is transmitted once, followed by a special position indicator (SPI) pulse for each interrogation pulse received. The SPI pulse is spaced 24.65 μsec from the leading edge of the first framing pulse of the first train. The result of the extra pulse train in Mode 1 and the SPI pulse in Modes 2 and 3 can be seen in figure 6-38. Two rows of dashed lines appear behind the target, instead of just one row as in normal operation. This feature is normally used to distinguish between two aircraft operating in the same mode and SIF code.

Emergency Reply Mode

The emergency mode of operation of the transponder is controlled by the master switch. Also, most aircraft have emergency IFF override switches on either the crew seats or the canopy. These switches energize when a crew member or canopy is jettisoned. These switches bypass the master switch and will automatically turn on the IFF and transmit emergency replies when interrogated.

The emergency function affects the operation of Modes 1, 2, and 3/A. In Modes 1 and 2, the reply pulse train containing the code in use is transmitted once for each interrogation pulse received, followed by three sets of framing pulses and no information pulses. The framing pulses will appear at 24.65 μsec, 44.95 μsec, 49.3 μsec, 69.6 μsec, 73.95 μsec, and 94.25 μsec. For each interrogation pulse received in Mode 3/A, one reply pulse train containing the code 7700 (regardless of the Mode 3 control dial settings) is transmitted, followed by three sets of framing pulses and no information pulses. The framing pulses are spaced the same as Modes 1 and 2. The result of these extra pulses can be seen in figure 6-38. Four rows of dashed lines appear behind the target, instead of just one in normal operation.

X-Pulse

The X-pulse (10. 15 μsec) from the initial framing pulse appears in a normally unused position of Modes 1, 2, and 3/A. It can only be obtained by modifying the control box. When the control box is modified, all replies in Modes 1, 2, and 3/A will include this pulse along with the normally selected information reply pulses, between the two framing pulses. To date, the use of this pulse has been restricted to drones.

As you can see, with the capabilities of the IFF/SIF system to respond with over 8,100 reply codes in Modes 1, 2, and 3/A, plus the encrypted Mode 4, and the ability to report altitude information, the system allows for positive identification at long range, and aids greatly in air traffic control.

CHAPTER 7

OPTIC AND INFRARED SYSTEMS

Chapter Objective: *Upon completion of this chapter, you should have a working knowledge of electrooptical disply units to include the AN/AVQ-7(V) and the AVA-12 head-up displays. In addition you should be able to recall the basic operating principles of infrared systems to include a forward looking infrared (FLIR) system.*

The purpose of an electrooptical display unit is to display attack and flight information directly to the pilot's field of view. The electrooptical sight uses a light source (fig. 7-1) to display information that is useful to the pilot. In figure 7-2, you can see a modern example of an optical sight. A cathode-ray tube (CRT) is used to produce the displayed information. This system is known as a head-up display (HUD). In this chapter, you

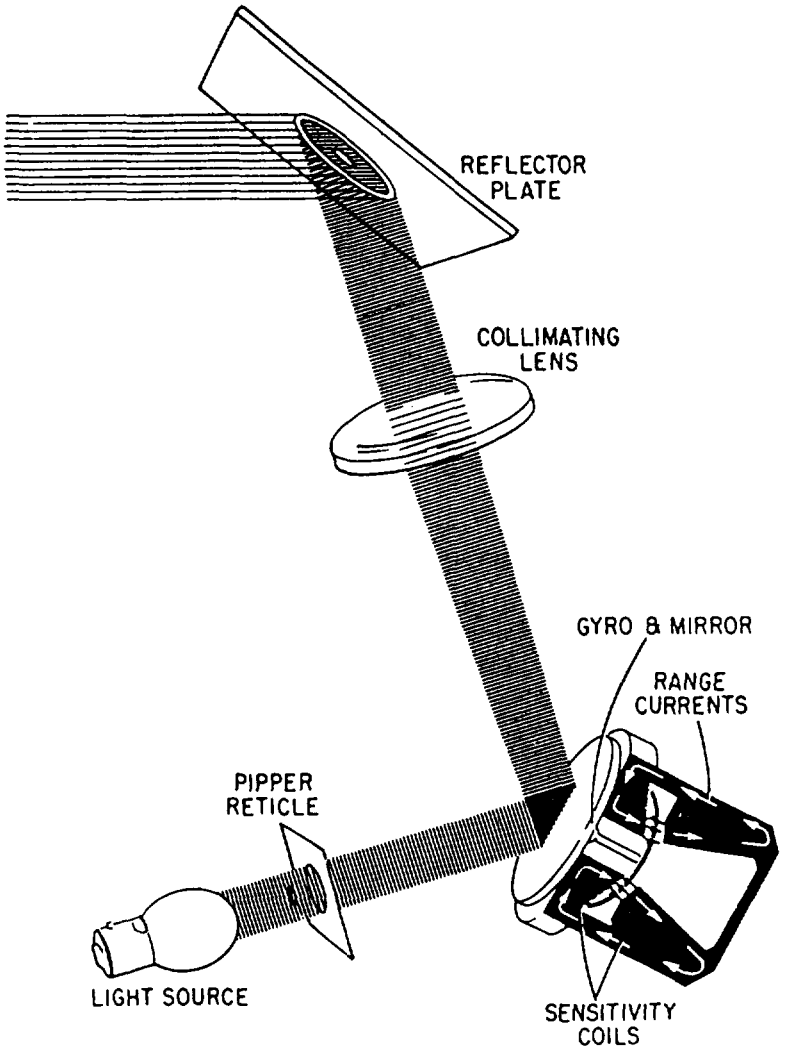


Figure 7-1.-Basic principle of a sight unit.

will see how these systems display attack and flight information.

The systems you will learn about in this chapter are the AN/AVQ-7(V) and the AVA-12. These were chosen for study because they are the systems currently in use. The AN/AVQ-7(V) is used in attack aircraft with a primary air-to-ground mission. The AVA-12 is used in fighter aircraft with a primary air-to-air mission.

AN/AVQ-7(V) HEAD-UP DISPLAY

Learning Objective: *Identify the components and operating principles of the AN/AVQ-7(V) head-up display.*

The AN/AVQ-7(V) receives computed attack and navigation input data from a tactical computer set, aircraft performance data from aircraft flight sensors, and discrete signals from various aircraft systems. Information received from the various systems is displayed on a transparent mirror (combiner) located directly in front of the pilot at eye level. The symbols are focused to infinity and are superimposed over real world objects in line with the aircraft flight path. Certain symbols are positioned on the combiner to correspond with real world object positions relative to the aircraft, even though the real world objects may not be visible.

OPERATION

The HUD set receives numerous input signals from many aircraft systems and sensors (fig. 7-3). The input data is processed by the signal data processor and applied to the HUD as horizontal (X), vertical (Y), and bright-up (Z) signals that provide the symbols that appear on the combiner. All symbols displayed by the HUD are shown in figure 7-4. Self-detected fail signals within the processor or the display unit are transmitted to the tactical computer and caution light panel.

Aircraft ac and dc voltages applied to the power supplies are converted to the required voltage levels and distributed throughout the processor. The precision 15-volt dc power supply output is used for excitation by the angle-of-attack transducer, air data computer, and Doppler radar. Power control of the processor is provided by the display unit. The built-in test equipment (BITE) monitors the power supply voltages and initiates a processor fail signal when an out-of-tolerance condition is detected. The fail signal is not initiated, however, until a self-test command (STC) signal is received from the display unit BITE. There are three types of input signals applied to the processor:. discrete

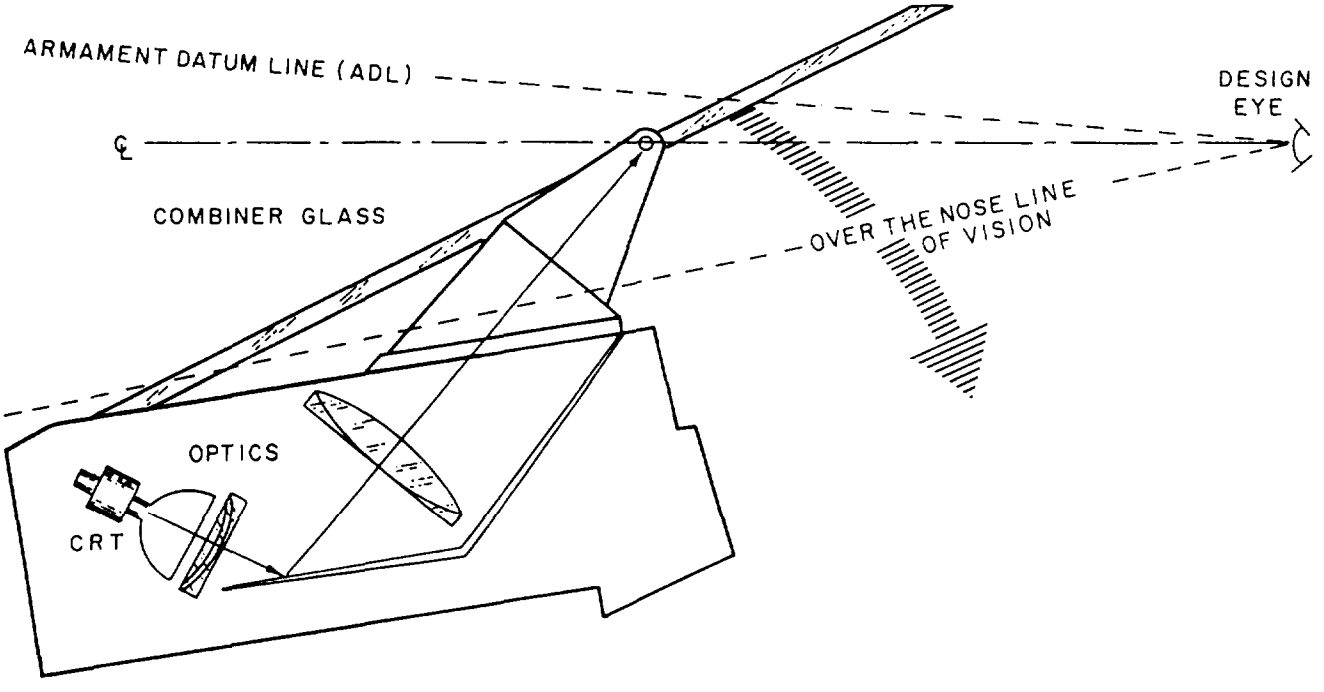


Figure 7-2.-New optical sight with CRT.

signals (switch closure), analog signals (variable), and digital signals (binary quantities). These signals are received by the input receivers within the processor, where each signal is identified. Data from the input receivers is checked for validity, parity, and identity by circuits in the processor. When valid, the data is stored in the data store section of the processor. Analog data from the input receivers is applied to the symbol generator, where the analog data is converted to digital form (ADC), and then transferred to the data store until needed by the processor.

Each time the demand-next instruction (DNI) pulse is applied to the program store, specific instructions (with decodes) are initiated to the processor. The instructions control the processing of all data contained in the data store. Data is transferred from the data store to the processor, where arithmetic functions are performed. The processed data is then returned to the data store. Other instructions transfer processed data (final data) to the symbol generator. The final data controls the generation of X and Y analog voltages used by the display unit. The busy signal prevents the transfer of information to the display unit while it is operating.

The optical module provides power control to the low-voltage power supply and symbol control to the processor. Aircraft power is applied to the optical module for control of the standby reticle and complete HUD operation. The BITE circuits, which sense a go/no-go condition in either the processor or the display unit, provide an output to the tactical computer and

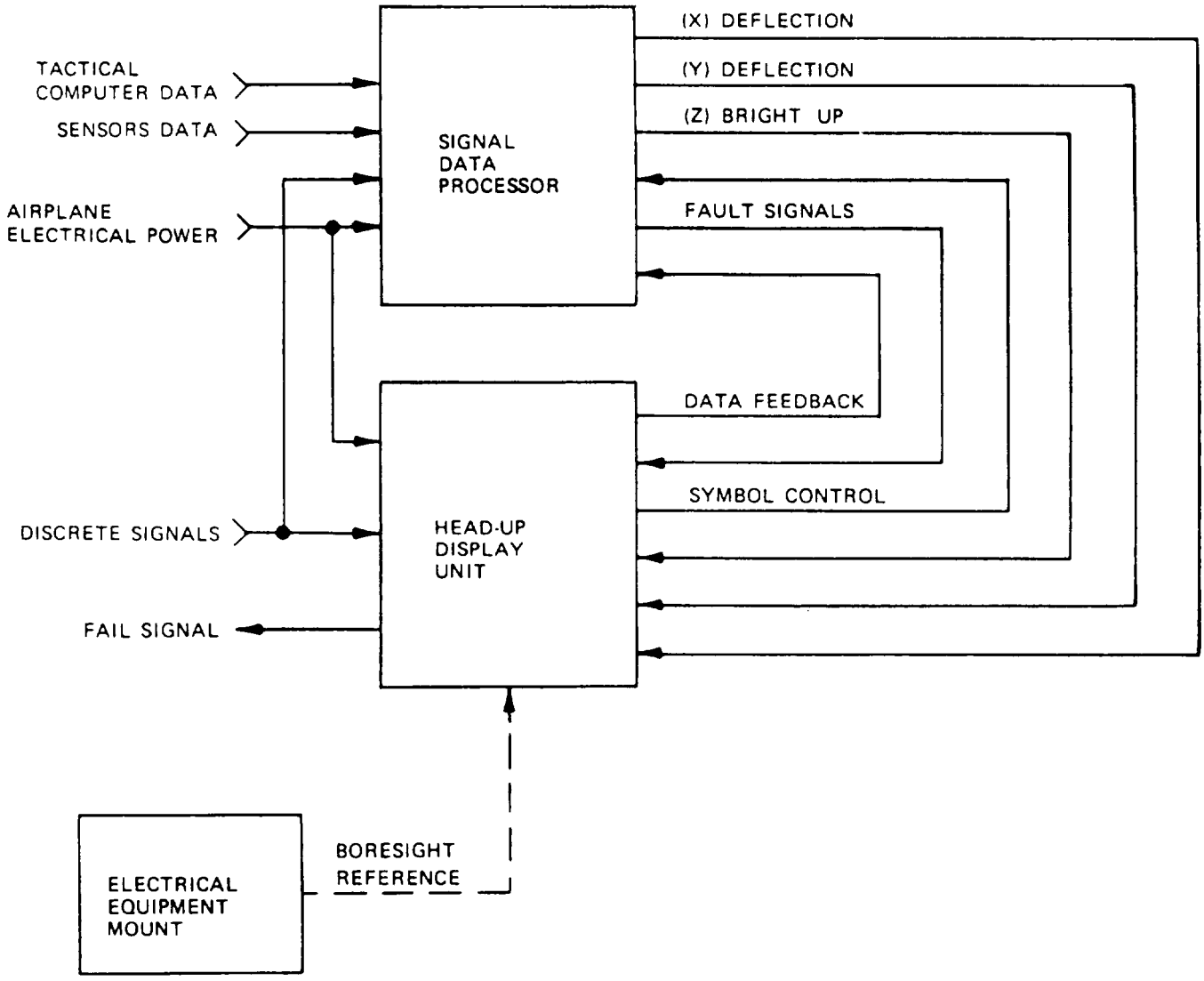


Figure 7-3.Head-up display set block diagram.

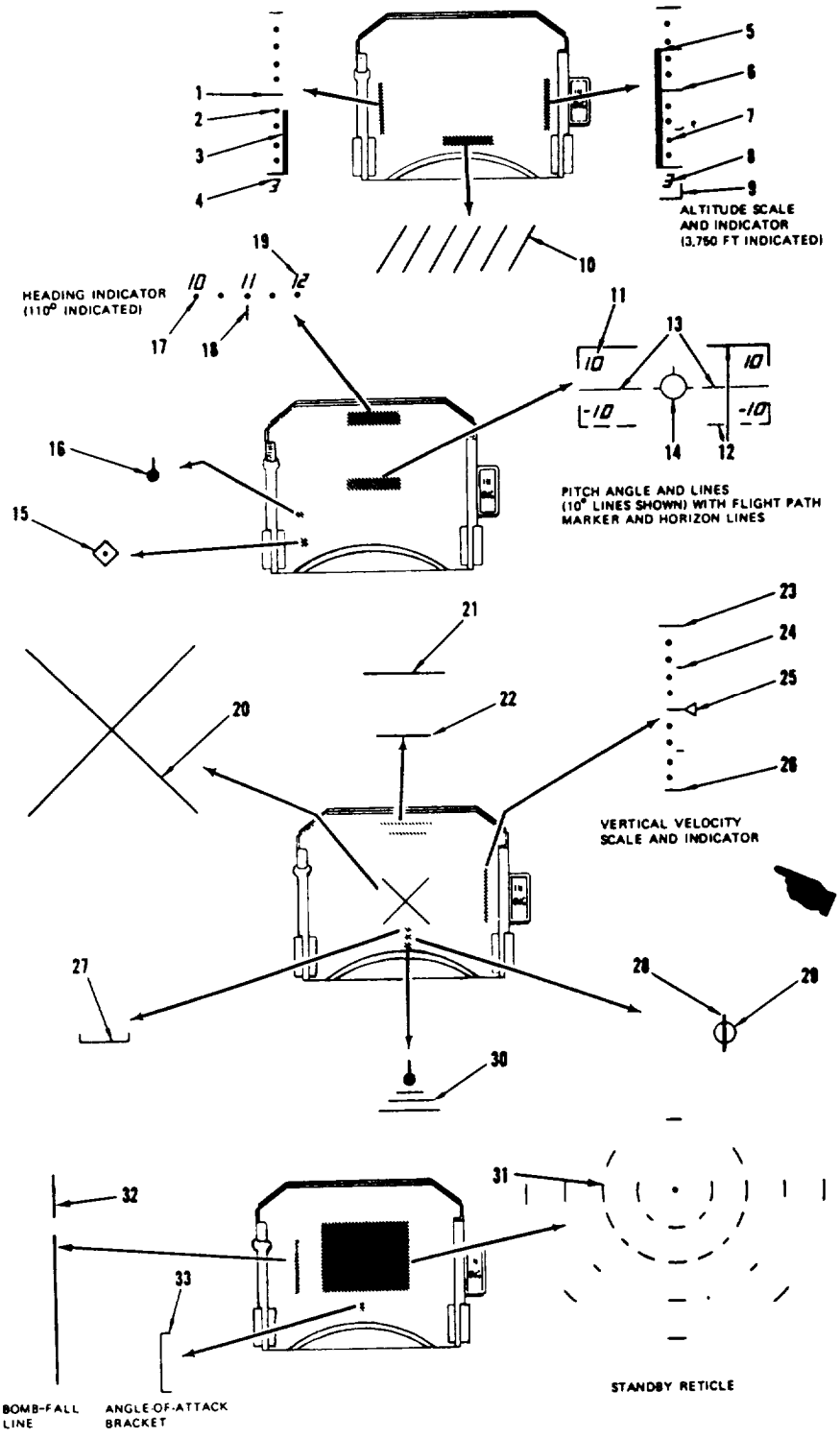


Figure 7-4.-Head-up display set displayed symbols.

INDEX NO.	NOMENCLATURE	FUNCTION
1	Index lines	Divides airspeed indicator into 50-knot increments.
2	Index dots	Divides airspeed indicator into 10-knot increments.
3	Airspeed column	Rises on increasing airspeed through increments of 100 knots.
4	Airspeed numeric	Indicates airplane airspeed in increments of 100 knots.
5	Altitude column	Rises on increasing altitude through increments of 1,000 feet.
6	Index lines	Divides altitude indicator into 500-foot increments.
7	Index dots	Divides altitude indicator into 100-foot increments.
8	Altitude numeric	Indicates airplane altitude in increments of 1,000 feet.
9	Radar altitude advisory symbol	Indicates radar altitude is being displayed (altitude switch in RDR ALT).
10	Warning indicator	Flashes when the master caution light goes on. Goes on steady when fire warning light goes on.
11	Pitch angle numeric	Indicates airplane pitch angle. (Scale is times 1°.)
12	Pitch lines	Airplane pitch axis index. (End tabs point in direction of horizon.)
13	Horizon lines	Represents relative position of horizon to the airplane pitch and roll axes.
14	Flight path marker	Represents the flight path of the airplane.
15	Aiming reticle	Enables pilot to locate and track target in attack mode or obtain fix point reference in navigation mode. (Dot in reticle on airplanes after A-7 Airframe Change No. 196.)
16	Flight director	Indicates the relative position of an optimum flight path.
17	Index dots	Divides the airplane heading scale into 5° increments.
18	Lubber line	Indicates airplane magnetic heading.
19	Heading numeric	Airplane heading scale. (Scale is times 10°.)
20	Pull-up command	Flashes on to indicate an immediate pull-up requirement.
21	Solution No. 2 cue	Symbol moves down the bombfall line to indicate an approaching weapons release point.
22	Solution No. 1 cue	Symbol moves down the bombfall line and indicates an approaching weapons release point.
23	Upper index line	Full scale climb index marker. Indicates 1,000 feet per minute climb.
24	Index mark	Divides the vertical velocity scale indicator into 500-foot increments.
25	Vertical velocity pointer	Moves up and down scale to indicate vertical velocity.
26	Lower index line	Full scale dive index marker. Indicates 1,000 feet per minute dive.
27	Pull-up anticipation cue	Symbol moves up to indicate an approaching pull-up requirement.
28	Lubber line (Airplanes before A-7 Airframe Change No. 196)	Side slip ball reference line.
29	Side slip ball	Indicates airplane lateral acceleration (side slip).
30	Landing director	Indicates the relative position of the landing flight path during a landing approach.
31	Standby reticle	A manually controlled optical sight used when the head-up display set is inoperative.
32	Bomb-fall line	Indicates weapons delivery path relative to the flight path of the airplane.
33	Angle-of-attack bracket	Moves up and down about the flight path marker to indicate the airplane angle of attack.

Figure 7-4.-Head-up display set displayed symbols—Continued.

caution light panel when the HUD is either off or inoperative. If an overheated condition exists, a separate circuit sends a HUD hot signal to the caution panel.

SIGNAL DATA PROCESSOR

Three-phase ac power is applied to six different rectifiers contained in the low-voltage power supply. The ac voltage is rectified by each rectifier, then regulated to a precise value. Each value of dc supply voltage is distributed for circuit operation through the signal data processor (fig. 7-5, a foldout at the end of the chapter).

Input Receivers

Digital data signals (table 7-1) are applied to the input receivers on four channels. Data transfer is in serial form, and all four channels are in operation at the same time. Each channel consists of a signal and a signal return line. Specific input data is applied to each channel as follows:

1. One channel receives data word signal and data word signal return.

2. One channel receives data identity signal and data identity signal return.

3. One channel receives data ready signal and data ready signal return.

4. One channel receives data clock signal and data clock signal return.

Data identity signals are transmitted simultaneously with each data word signal and identify each data word. Data identity signals consist of 20 bits of data, which contain the information shown in table 7-2.

Data ready is indicated by a digital 1 transmitted concurrently with bit 1 of the data word signal and the leading edge of a data clock pulse. The data clock pulse applied to the input receivers is generated within the tactical computer. The clock pulse has a pulse rate of 50 kHz (+10%) with a 50 percent duty cycle. The digital data from the tactical computer is continuously accepted by the input receivers and shifted into the buffer register. Each data word is identified by data identity and clock pulse signals applied to the identity register. At specific intervals in each cycle, accumulated raw data is shifted to the digital computer memory via the adder. Addresses for the raw data are provided by the identity register during the shifting process. There are 12 discrete lines connected to 12 discrete receivers contained in the input receivers. The discrete receivers convert any applied discrete signal to an optimum voltage, which is used to control the content of a 12-bit data word.

During a digital computer cycle, according to instructions from the program store, a discrete word is read into the function decode. The control terms provided by the discrete word are applied to the digital computer as basic instructions, which may or may not affect digital computer operations. Ten analog voltages are divided equally and applied to five X channel amplifiers and five Y channel amplifiers. Each analog

Table 7-1.-Digital Data Signal

DATA BIT	DESIGNATION
1	Control bit (binary one)
2	Validity bit (binary one when valid)
3-5	Not used
6-17	Data bits
18	Sign bit (+ equals binary zero)
19	Spare bit (binary zero)
20	Parity bit (odd)
	Digital one is: 0.0 (±0.5) V dc
	Digital zero is: +5.0 (+1.0) V dc

Table 7-2.-Data Identity Signal

DATA BIT	DESIGNATION
1	Control bit (binary one)
2-11	Not used
12-18	Data identity (LSB least significant bit first)
19	Spare bit (binary zero)
20	Parity bit (odd)
	Digital one is: 0.0 (+0.5) V dc
	Digital zero is: +5.0 (+1.0) V dc

voltage is amplified, then applied to the input selectors. On a gating control command from the symbol generator, one specific analog voltage from each channel is gated to the symbol generator where the analog-to-digital conversion (ADC) is performed.

Digital Computer

Designated letters are used to represent certain phases of digital computer operations. A number following a designated letter indicates a specific step of the phase in progress. The digital computer processes the applied digital data in three phases. After the three phases are completed, the entire cycle is repeated. All steps of each phase are performed sequentially according to a prewired program of instructions contained in the program store. When electrical power is applied to the digital computer, clock pulses are generated within the clock generator of the program store. The clock pulses are distributed by the control logic to the sequence control, input receivers, symbol generator, and processor counter. The processor counter is used to record a specific number of clock pulses so that the DNI pulse is not arbitrarily initiated. However, if a jump signal is received by the function decode, a DNI pulse is initiated. The jump signal maybe repeated until the correct instruction is selected. When the DNI pulse is initiated and is in coincidence with a specific clock pulse in the sequence control, a data request signal is applied to the program of instructions.

At the start of digital computer operations, the first program of instructions selected is data accumulation (phase A) due to resetting of the sequence control when electrical power is applied. All instructions of phase A are sequentially applied (as step A1) to the processor function decode. The function decode determines the addresses of all the instructions received, and applies this information to the memory. The phase A instructions are decoded by the function decode and applied as read-in commands (as step A2) to the adder. All data accumulated in the buffer register is transferred (as step A3) via the adder (as step A4) to the memory, which completes phase A of digital computer operations. The function decode is then ready for the specific clock pulse from the counter that initiates the DNI pulse to the sequence control.

When the next DNI pulse is received by the sequence control, a request for the next sequence of instructions (phase B) is applied to the program of instructions. All instructions contained in the next program are sequentially transferred (as step B1) to the function decode. The phase B instructions are decoded by the function decode, and all addresses contained in the instructions are applied to the memory. The decoded phase B instructions are commands to transfer certain data (as step B2) contained in the memory to the adder. Other phase B instructions command arithmetical functions to be performed (as step B3) by the adder. As soon as solutions are obtained, they are transferred (as step B4) to the memory. Phase B of computer operations is then complete. The sequence control is then ready for the next DNI pulse from the function decode.

The next DNI pulse starts phase C of the digital computer operations. Selection of phase C instructions is initiated to the program of instructions. The next sequence of operations is the application of coded instructions (as step C1) to the function decode. The phase C instructions are decoded by the function decode, and all addresses contained in the instructions are applied to the memory. The decoded phase C instructions are commands to transfer the processed data (as step C2) via the adder (as step C3) to the symbol generator, which completes the three-phase cycle. On the next DNI pulse, the entire process is repeated. This cycle continues until electrical power is removed from the digital computer.

Symbol Generator

The symbol generator operates in three major modes. The three modes of operation are the line, circle, and analog-to-digital conversion (ADC). Each mode is independently and completely performed before repeating or starting a different mode. The mode to be performed is initiated by the function decode in the digital computer. Correct sequencing of each mode operation is provided by the timing pulses from the control logic section of the digital computer. The program in the digital computer determines which mode is initiated to the symbol generator.

LINE MODE.— When a line is to be drawn in the CRT, a line mode signal is initiated to the input/output buffer and control of the symbol generator. This signal starts a counter that controls all operations of the symbol generator. The first operation is to transfer the X1 data from the memory through the adder and input/output buffer and control to the X channel deflection register. The X1 data is the digital equivalent of the analog voltage used to drive the electron beam to the start point (in the horizontal axis) of the line to be drawn.

The second operation is to transfer the Y1 data from the memory to the Y channel deflection register. The Y1 data is the digital equivalent of the analog voltage used

to drive the electron beam to the start point (in the vertical axis) of the line to be drawn. The digital data contained in each deflection register is immediately converted to an equivalent analog voltage by the X and Y channel digital-to-analog converters (DAC). This analog voltage (X and Y deflection voltage) is held by each DAC until gated to the HUD by the input/output buffer and control.

The third operation is to transfer the X2 data from the memory to the parameter register. The X2 data is a digital quantity representing the line slope angle cosine. In the first half of the fourth operation, the X2 data is shifted from the parameter register to the X channel rate register. In the second half of the fourth operation, Y2 data is transferred from the memory to the parameter register. The Y2 data is a digital quantity representing the line slope angle sine. In the first half of the fifth operation, the Y2 data is shifted from the parameter register to the Y channel rate register. In the second half of the fifth operation, the T data is transferred from the memory to the parameter register. The T data is a digital quantity representing the length of the line to be drawn (bright-up pulse width). In the sixth and seventh operations, all data contained in the X and Y channel rate registers are checked for correctness by the BITE circuits. All the data required to draw a specific line (which may be only part of a symbol) is now contained in the symbol generator.

When all the data required to draw a specific line is contained in the symbol generator, the busy signal from the HUD is sampled. If the busy signal is high, the operations stop until the busy signal is low. When the busy signal goes low, the input/output buffer and control sends a start pulse to the parameter register, which causes it to start down counting. The down counting controls the length of time the bright-up pulse is applied to the HUD. At the same instant, a gating pulse is applied to each DAC. This allows an analog voltage (produced by the digital content of each deflection register) to be applied to the deflection circuits in the HUD. Also, at the same instant, the contents of the rate registers are added to or subtracted from the respective residue register, which causes a positive or negative overflow to be generated. The overflow is used to drive the deflection registers up or down. This causes the start point of the line to move in the direction of the line slope angle. After a preset time delay, a bright-up pulse is applied to the HUD. The time delay will compensate for the slower response time of the deflection circuits in the HUD. The bright-up pulse continues until the parameter register has down counted to zero. When the parameter register stops counting, the bright-up pulse is turned off, and the end of the line mode is indicated to the digital computer. The line mode may be repeated as many times as necessary to complete the required symbol.

CIRCLE MODE.— When a circle is to be drawn on the CRT, a circle mode signal is initiated to the input/output buffer and control. This signal starts the operation counter, as in the line mode. The data required to draw a circle, in a specific location, is transferred to the symbol generator in the same manner as in the line mode. However, the X2 data is equal to negative one and the Y2 data is all zeros. The T data set into the parameter register is the digital equivalent of one over the circle radius. Circles are drawn in a counterclockwise direction starting from the top. The overflow from the residue registers control the direction and amount of change in each deflection register. However, in the circle mode, the overflow also controls the circle logic to the opposite channel. The circle logic is such that a positive X channel overflow causes the contents of the parameter register to be added to the contents of the Y channel rate register. If the X channel overflow is negative, then the contents of the parameter register are subtracted from the Y channel rate register. On the other hand, if the overflow from the Y channel residue register is positive, then the contents of the parameter register are subtracted from the X channel rate register. If the Y channel overflow is negative, then the contents of the parameter register are added to the X channel rate register. This cross-coupling causes the two channels to be interdependent on each other. As the rate of change in one channel increases, the rate of change in the other channel decreases, or vice versa. If the circle being drawn is large, the rate of change in the analog deflection voltage output is large. If the circle is small, the rate of change in the analog deflection voltage is small. The rate of change is inversely proportional to the contents of the parameter register, which contains the inverse of the circle radius. Thus, a circle with the desired radius is produced.

The bright-up pulse is controlled by a quadrant counter in the input/output buffer and a control, which turns on the bright-up pulse after the electron beam has traversed halfway around the circle. The bright-up pulse delay compensates for the slow response time of the deflection circuits in the HUD. The bright-up pulse stays on until the quadrant counter has counted four times. A short time after the fourth count, the bright-up pulse is turned off and the symbol generator busy signal is terminated. This completes the circle mode.

ANALOG-TO-DIGITAL CONVERSION (ADC) MODE.— The third and final mode of symbol

generator operation is the analog-to-digital conversion (ADC) mode. In this mode, the operation of both (X and Y) channels is identical. Therefore, only the X channel operation is described. At the start of the ADC mode, the most significant bit in the parameter register is set to the one level. The rate and deflection registers are set to zero and coupled directly to each other. The direct coupling of the two registers is provided so the contents of either register will be duplicated. The third step gates a specific analog voltage from the X channel input selector to one side of the X channel comparator. The fourth step gates the digital-to-analog converter (DAC) voltage from the X channel DAC to the other side of the X channel comparator. The DAC voltage represents the contents of the X channel deflection register (and rate register due to coupling). If a difference in voltage level exists between the two applied voltages, the comparator has an output applied to the deflection register. The output of the comparator represents the sign of the difference and causes the contents of the parameter register to be added to or subtracted from the contents of the rate register.

On the next operation, the one in the parameter register is shifted to the next lower bit location and the voltage comparison phase repeated. This process is continued until the parameter register reaches zero. When the parameter register reaches zero, the contents in the rate register (in digital form) are equal to the input selectors analog voltage applied to the comparator. The next operation transfers the contents of the rate register to the memory by way of the data feedback (DFB) lines. In subsequent operations, the remaining analog voltages are applied to the comparator, and the entire process repeated. The ADC mode continues until all analog voltages are converted to digital data and stored in the memory.

HEAD-UP DISPLAY

Aircraft dc power applied to the control panel in the optical module is used for operational control of the signal data processor, head-up display unit, and standby reticle (fig. 7-6). Aircraft ac power applied to the low-voltage power supply is rectified, then divided into seven different dc levels for distribution throughout the HUD. A transistorized oscillator to the high-voltage power supply receives 24 volts of dc excitation from the low-voltage power supply. The oscillator output is applied to a voltage multiplier and rectifier circuit that increases the input voltage to 15,000 volts. The 15,000 volts is applied to a CRT anode and a voltage divider within the high-voltage power supply. Outputs from the voltage divider are used to control the oscillator frequency and a comparator output. The comparator senses if differences exist between the voltage divider output and a fixed reference voltage. If the high voltage drops below a prescribed level, a fail signal is generated by the comparator and applied to the BITE circuit.

Optical Module

The optical module contains the standby reticle, in-range indicator, control panel autobrilliance sensor, and lens unit. The standby reticle is used only when the HUD is inoperative, and is controlled electrically and mechanically from the control panel. Electrical power used for operation of the standby reticle is obtained from outside the HUD.

Light emitted from the standby reticle is transmitted through the lens unit to the combiner. The in-range indicator indicates when a range discrete signal is received from the forward looking radar set. Symbols are represented by light emitted from a CRT located in the video module. As symbols are drawn on the CRT face, the emitted light is received by the lens unit and applied to the combiner. The desired level of brightness is selected from the control panel by adjustment of the voltage level applied to the cathode bias circuit. The autobrilliance sensor is used to detect ambient light changes. Any change in ambient light proportionally changes the output voltage level of the autobrilliance sensor.

Video Module

The video module contains the bright-up and autobrilliance amplifiers, cathode bias circuit, horizontal and vertical deflection coils, and CRT. Symbols are drawn at the rate of 50 times a second on the CRT. The rapidly moving electron beam is generated by the bright-up pulse from the signal data processor. The bright-up pulse is amplified by the bright-up amplifier, and then applied to the CRT grid. The CRT, which is normally biased into cutoff, is turned on, and an electron beam is emitted. The electron beam strikes the CRT face at a position determined by the amount of current flowing in the horizontal and vertical deflection coils. The amount of light (symbol brightness) emitted from the CRT is controlled by the cathode bias circuit. A control voltage from the control panel is applied to the cathode bias circuit, which serves as a reference voltage for the autobrilliance sensor and the CRT cathode. The voltage output from the autobrilliance sensor is amplified by the autobrilliance amplifier. The output is

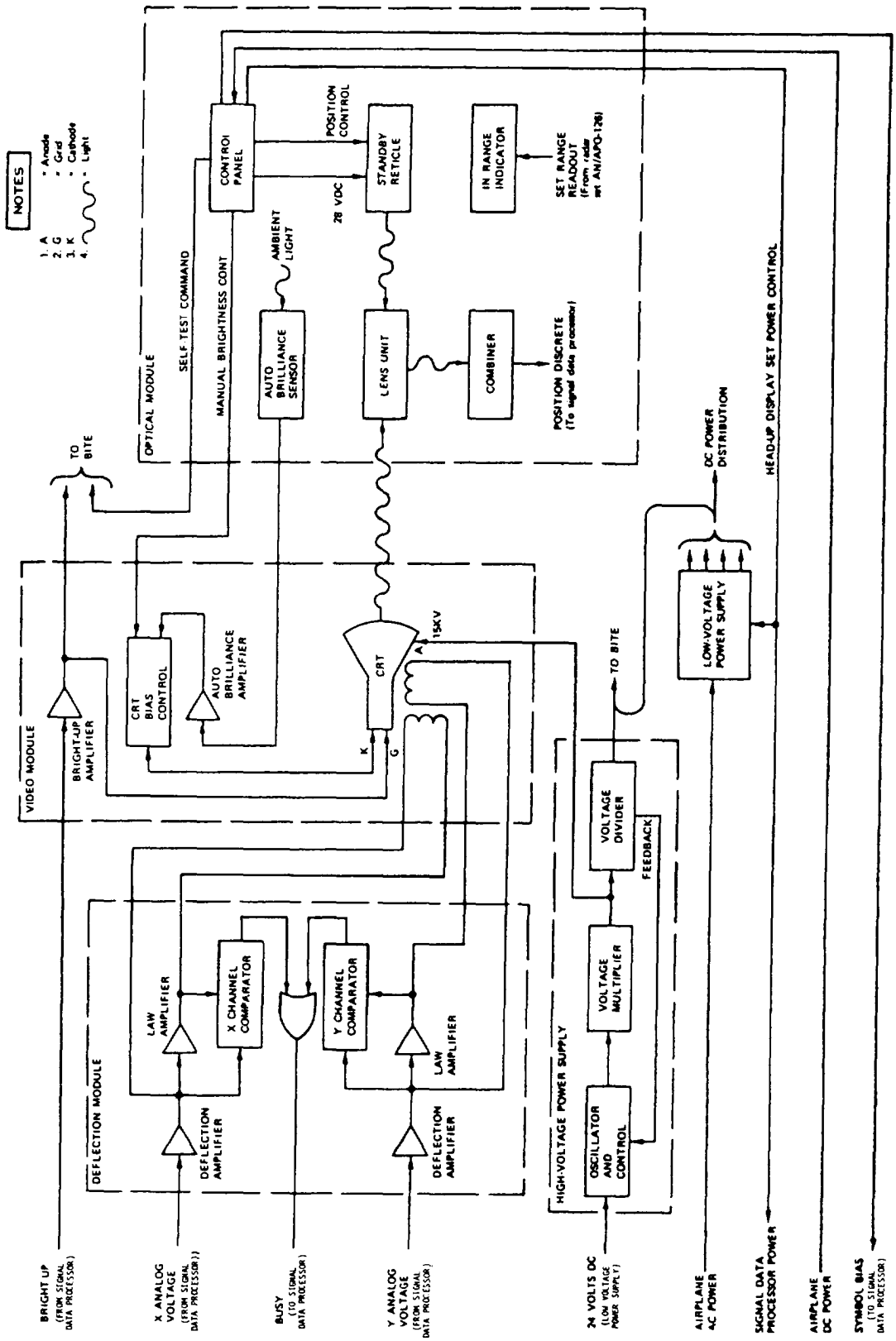


Figure 7-6.–Head-up display unit signal flow diagram.

applied to the cathode bias circuit, which changes the amount of bias (brightness) on the CRT. This, an optimum CRT contrast, is constantly maintained under varying ambient light.

Deflection Module

The deflection module contains the X and Y deflection amplifiers, X and Y law amplifiers, X and Y comparators, and one OR gate. The type and location of each symbol on the combiner is determined by X and Y analog voltages applied to the deflection circuits. As the need for a symbol arises, the busy signal from the HUD is sampled by the symbol generator. The busy signal is generated any time current is flowing in either deflection coil. Comparators, connected across each coil, are used to detect when the deflection coil current is equal to zero. As the current in either deflection coil reaches zero, the output from the comparator, connected across the coil, goes low. When both comparator outputs are low, the OR gate output goes low, and the busy signal is removed from the symbol generator input. At the same instant, the symbol generator applies a precise analog voltage to the X and Y deflection amplifiers. The output from the deflection amplifiers causes current to flow in the deflection coils. The current amplitude is precalculated to drive the CRT electron beam to the start point of the symbol to be drawn. A bright-up pulse is then applied to the bright-up amplifier in the video module. At the same instant, the deflection voltage is modulated at a predetermined rate and amplitude by the symbol generator. The modulation is detected by the deflection amplifiers and applied to each deflection coil. A specific part (or all) of a symbol is then drawn on the CRT face in a precise location. The bright-up pulse continues long enough to draw the prescribed line or circle that makes up the symbol. When the bright-up pulse is removed, the CRT is driven into cutoff until the next bright-up pulse is applied. This process is repeated until all symbols have been displayed on the CRT.

BUILT-IN TEST EQUIPMENT (BITE)

The built-in test equipment (fig. 7-7) contained in the signal data processor (SDP) consists of a clock pulse monitor, data test equipment, raw data test equipment, six low-voltage comparators, signal data processor fail indicator, and thermal overload sensor. The AND and OR gates represent various circuit functions of the built-in test equipment. The thermal overload sensor produces an output when the temperature inside the signal data processor rises above an operational limit. The overheat signal is applied to the HUD BITE circuit for subsequent transfer to the caution light panel. The clock pulse monitor checks the frequency of the clock generated in the digital computer section. When an out-of-tolerance condition is detected, a fail signal is applied to OR gate one (OG-1) from the clock pulse monitor.

The data test equipment provides two types of data checks. One check consists of comparing the data from each memory location (except raw data locations) to known reference data. If an error is detected, a fail signal is applied to OG-1. The other check consists of comparing the data content, set into the symbol generator, with the original data content in the memory. If an error is detected, a fail signal is applied to OG-1. The data clock, data word, and data identity signals from the input receivers are applied to the raw data test equipment, where the validity and identity checks are performed. If an error is detected in the raw data, a fail signal is applied to OG-1 and the input receivers. The fail signal applied to the input receivers is an inhibit signal that prevents the invalid data from being transferred to the digital computer memory.

The six low-voltage comparators are used to compare a specific voltage from the low-voltage power supply with a fixed reference voltage. If the voltage from the low-voltage power supply decreases below the fixed reference voltage (within a given percent), a fail signal will be generated by the comparator and applied to OG-1. The output of OG-1 is at high whenever a fail signal has been detected on any one of its nine input lines. The output line of OG-1 is connected to AND gate one AG-1, AG-2, and the HUD unit BITE circuit. When the inhibit line of AG-1 is high, the output of AG-1 remains low even though a deflection monitor fail signal is present. This prevents the generation of a false deflection monitor fail signal due to a defective signal data processor. The level of the output line of AG-2 is dependent upon the level of its input lines from OG-1 and the HUD unit BITE circuit. When both input lines of AG-2 are high, a fail signal is applied to the SDP fail indicator. The SDP fail indicator is an electromechanical device that must be manually reset once it has been electrically activated.

HEAD-UP DISPLAY UNIT

The built-in test equipment (BITE) contained in the HUD unit consists of a high-voltage comparator, seven low-voltage comparators, a bright-up parity circuit, a HUD fail indicator, and a thermal overload sensor. The

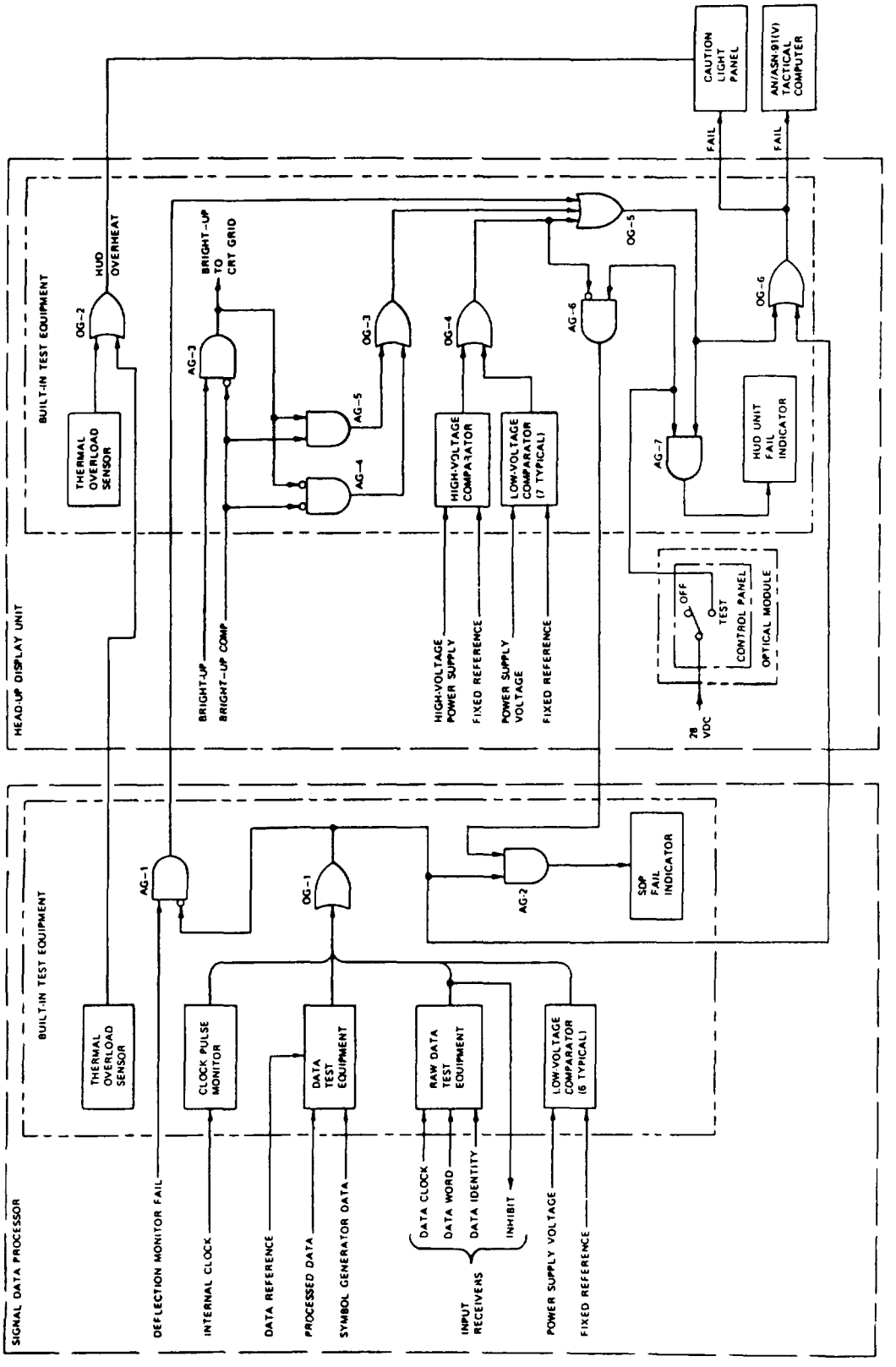


Figure 7-7.–Built-in test equipment signal flow diagram.

AND and OR gates represent various circuit functions of the BITE. The thermal overload sensor provides the same function as the thermal overload sensor contained in the signal data processor. Both sensor output lines are connected to the input of OG-2. If the output of either sensor is high, an overheat signal from OG-2 is applied to the caution light panel. The signal data processor BITE circuit has two other output lines connected to the HUD BITE circuit. One is the deflection monitor fail, which is one of three input lines to OG-5. The other is the signal data processor fail line, which is one of two input lines to OG-6. When either of the input lines of OG-6 is high, a fail signal from OG-6 is applied to the caution light panel and tactical computer. The fail signal indicates that the HUD has failed. When the deflection monitor fail line goes high, the output of OG-5 goes high. The output line of OG-5 is connected to one of the input lines on both AG-7 and OG-6. The bright-up signal input line is connected to one input of AG-3.

The bright-up signal complement line is connected to one input of AG-5, the inhibited input of AG-3, and one of the inhibited inputs of AG-4. When the bright-up input line is high and the bright-up complement line is low, a bright-up pulse is applied to the CRT grid. The bright-up pulse is also applied to the other input of AG-5 and the other inhibited input of AG-4. As long as the bright-up complement line is low and the output line of AG-3 is high (or vice versa), there will be no high level output from AG-4 or AG-5. But, if both the bright-up complement and the AG-3 output lines are low at the same time, the output of AG-4 will go high. Also, if both the bright-up complement and the AG-3 output lines are high at the same time, the output of AG-5 goes high. The outputs of AG4 and AG-5 provide the inputs of OG-3. Thus, if the output of either AG-4 or AG-5 is high, a bright-up parity fail signal is applied to input of OG-5 from OG-3.

The high-voltage comparator is used to compare a known voltage from the high-voltage power supply with a fixed reference voltage. If the known voltage drops below a specified level, the comparator generates a fail signal to OG4. The seven low-voltage comparators are used to compare a specific voltage from the low-voltage power supply with a fixed reference voltage. If the voltage from the low-voltage power supply should decrease below the fixed reference voltage (within a given percent), a fail signal is generated by the comparator and applied to OG-4. The output of OG-4 is high whenever a fail signal is detected on any one of eight input lines. The output of OG-4 is connected to the input of OG-5 and the inhibited input of AG-6. The other input of AG-6 and AG-7 is connected to the self-test switch on the control panel. The output of AG-5 is connected to the signal data processor BITE circuit. The output line of AG-6 is high whenever the self-test command (28 volts dc) is applied to one input of AG-6 and the output line from OG4 is low. However, if the output line from OG4 is high, the output of AG-6 remains low. This prevents a defective HUD power supply from generating a false signal data processor fail signal. The output line of AG-7 is high whenever the self-test command signal is applied to one of its input lines, and the output line of OG-5 is high. The output of AG-7 is the HUD fail signal, which is connected to the HUD unit fail indicator. The HUD unit fail indicator must be manually reset once it has been electrically activated.

AVA-12

Learning Objective: *Recognize the basic operation and modes of operation of the AVA-12 head-up display.*

The basic operation of the AVA-12 head-up display (HUD) is like that of the AN/AVQ-7(V). The symbols of the two HUDs are generated in the same way. The basic differences between the two systems are the types of symbols that are generated, their use, and the type of aircraft on which the systems are used. The AVA-12 is used on tighter aircraft with a basic air-to-air mission.

Look at the block diagram of the AVA-12 in figure 7-8. As you can see, the indicator unit receives the deflection and voltage symbols from the converter. Now, look at the converter in figure 7-9. The converter takes the aircraft system information and converts it into information that is usable by the display unit.

SYMBOLOGY

The symbols used with the various modes of operation on the AVA-12 are listed and described in figure 7-10. These symbols give the pilot important information such as aircraft attitude, heading, altitude, angle of attack and ground track during flight. Attack information including closure rate, range to target, maximum range for weapon launch, minimum range for weapon launch, and boresight reference is available during air-to-air and air-to-ground operations. Ordnance information includes the number of rounds remaining, the type of weapon selected, and the number of weapons ready for launch.

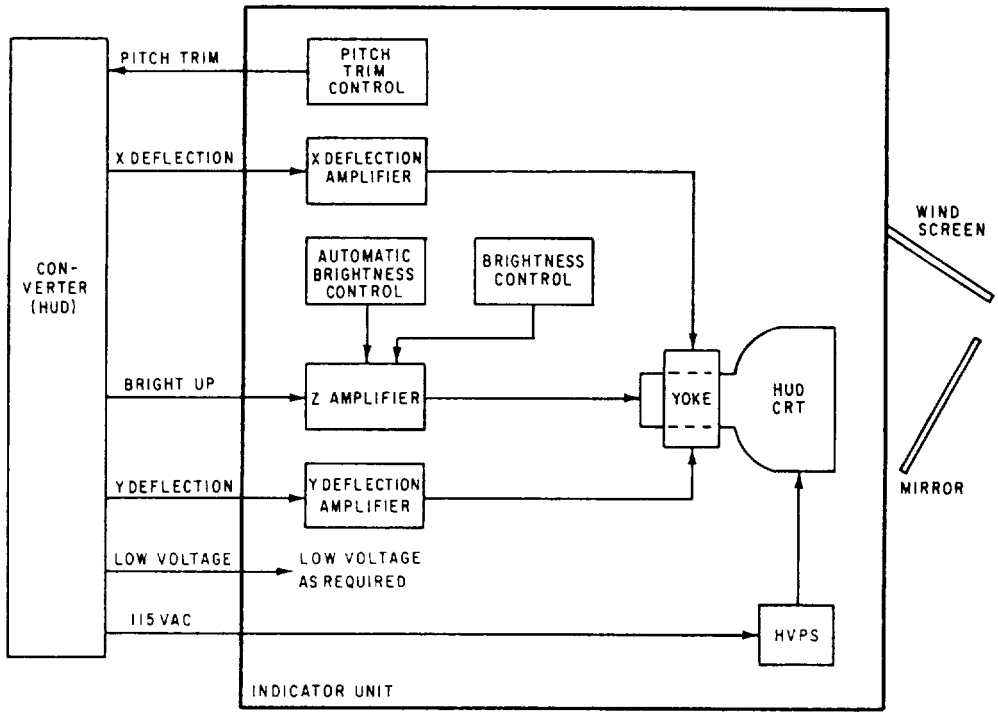


Figure 7-8.-Head-up display indicator block.

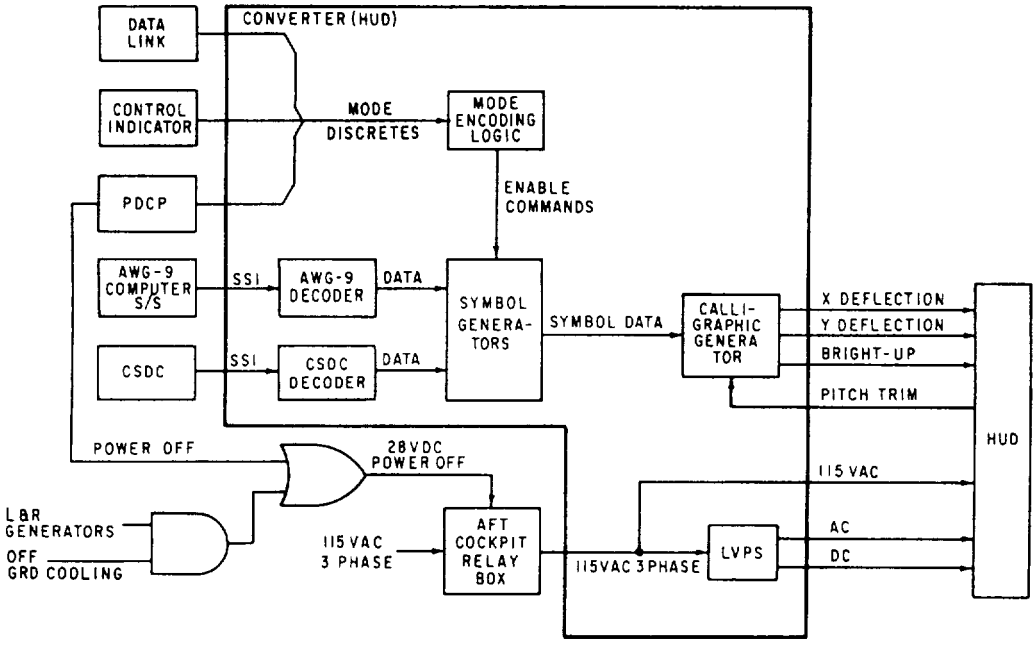


Figure 7-9.-HUD converter block.

SYMBOLOGY

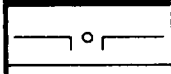
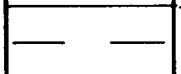
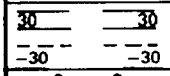
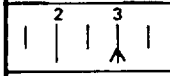
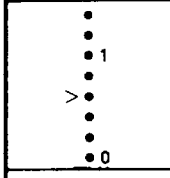
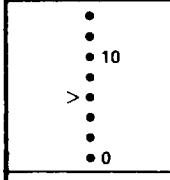
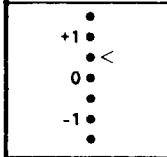
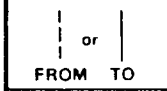
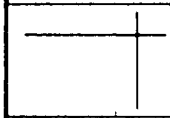
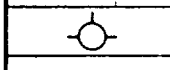
HUD SYMBOL	NAME	FUNCTION
	Aircraft reticle	Depicts own aircraft wings and when lined up with the horizon, the aircraft is in straight and level flight.
	Horizon	This is a demarcation point between ground and sky textures on the VDI. It represents the horizon with respect to the aircraft, and changes orientation with any change in aircraft pitch or roll.
	Pitch lines	Indicates with respect to aircraft reticle pitch attitude. In cruise and A/A mode, HUD pitch lines have a 4:1 compression ratio. Dotted lines indicate a negative pitch. Solid lines indicate positive pitch. Pitch lines above the magnetic heading scale are blanked.
	Magnetic heading (AHRS or WCS via CSDC)	Indicates magnetic heading with respect to index mark.
	Radar Altimter Scale	Indicates altitude derived from radar altimeter. Scale is from 0 to 1400 feet in 200-foot increments and has a movable pointer. This symbol is only available in the takeoff and landing modes.
	Barometric Altimeter Scale	Indicates altitude derived from the pressure altitude via the CADC. Scale is from 0 to 14,000 feet in 2000-foot increments. This scale is available in A/G and TARPS modes. The altitude is referenced to 29.92 inches Hg.
	Vertical Speed Indicator	Indicates rate of altitude change. Scale is from -1500 to +1500 ft/min in 500-foot increments. This scale appears on the left side of the HUD in takeoff and landing modes.
	TACAN Deviation Bar (TACAN via CSDC)	Indicates difference between bearing to TACAN and selected TACAN radial. Deviation is limited to ±5.625° TACAN deviation on the VDI, and ±3° TACAN deviation on the HUD. The symbol never leaves field of view. It will limit at edge nearest selected TACAN radial.
	Breakaway	Appears as a flashing symbol at a 3-cycle per second rate in the center of field-of-view when range-to-go-to minimum or safe pullup point is zero. Symbol is commanded by the WCS computer or by D/L, depending on mode of operation.
	Precision course Vector	This symbol consists of two independent vectors (vertical and horizontal), which form a cross pointer. Elevation glide slope information positions the horizontal vectors, whereas the vertical vector is positioned by azimuth glide slope information. This symbol is also used in D/L bombing modes.
	Velocity Vector	Indicates direction of ground track velocity vector (where the aircraft is going, not where it's pointed).
	Angle of attack error (CADC via CSDC)	Its position in relation to aircraft reticle indicates angle-of-attack error. Symbol is position by true AOA. Small center horizontal bar indicates zero error. When this symbol is in line with the aircraft reticle, the AOA is 15 units (10.31°). If symbol is below aircraft reticle, AOA is too high; above the aircraft reticle indicates AOA is too low. Symbol is displayed in landing mode only.
ORD, G, SW, SP, or PH 0,1,2,3,4,5,6	Armament Ready Legends	ORD – indicates bombs or rockets selected, or bombs and gun selected (A/G GUN switch on ACP set to MIXED) G – indicates gun is selected or gun and bombs are selected (A/G GUN switch on ACP 6 set to OFF). Number under G indicates rounds remaining in hundreds (6, 5, 4, 3, 2, 1, 0). SW, SP or PH – indicates missile type selected (Sidewinder, Sparrow or Phoenix) and the numbers (0 to 6) indicates number of missiles ready for launch.

Figure 7-10(A).-Symbology for operational modes of the AVA-12.

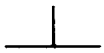
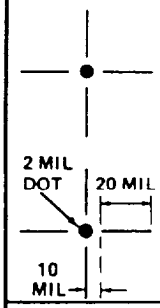
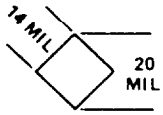
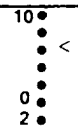
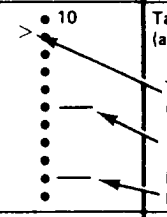
HUD SYMBOL	NAME	FUNCTION
	MASTER ARM switch off	An X symbol through armament ready legend indicates MASTER ARM switch on ACM panel is OFF. Disappears when MASTER ARM switch is set to ON.
	Steering Tee	Provides elevation and azimuth steering in the air-to-air modes when a single target track exists. Type of steering (pursuit, collision, etc.) is dependent on weapon selection and the mode selected on the TID by the RIO. May also provide azimuth steering only on the VDI in TWS. Aircraft steering is accomplished by aligning and maintaining the vertical and horizontal bar of the inverted T with the aircraft reticle center dot. Steering sensitivity on the HUD is 26.5° per inch; on the VDI 25° per inch.
	Boresight Reference	Symbol is a set of crosshairs which is fixed on the HUD and is used to represent the armament data line (ADL) of the aircraft. The reference is located 5.03° above aircraft reticle.
	Moveable Reticle (Impact Point)	Symbol serves as an optical sight for A/A gunnery and for A/G weapons delivery. In A/A gunnery, the symbol is used for lead angle determination. It can be positioned manually using the elevation lead control or automatically by the computer when the pilot selects the gun solution mode. The symbol is so designed that stadiametric ranging techniques can be employed during manual gun firing mode. This symbol in A/G modes indicates instantaneous weapon impact point. This symbol is positioned by the computer in all modes except manual. In the manual mode, it is positioned by the elevation lead control. The computer positions the symbol based on the ballistics of the bomb, wind conditions, and various aircraft parameters. This symbol must overlay the target at the moment of release.
	Target Designator	Indicates radar pointing angle. This symbol is used in all computer A/G modes. It is positioned in this case by the pilot using the TARGET DESIGNATE switch. Once the symbol is over the target, the switch is depressed and the computer knows the slant range to the target. In real-time gunsight mode of air-to-air gun, the target designator is positioned at the 1000-foot bullet solution. In all other weapon modes or with OFF selected, and with a valid sensor angle track on the target, the target designator symbol represents the approximate line of sight to the target. If the target is not within the HUD field of view (FOV) the symbol is positioned at the edge of the FOV in the direction of the target. Limits of designator are 10° in horizontal and vertical.
	Closure Rate	Indicates closing velocity from -200 to +1000 knots between aircraft and target.
	Target Range Scale (azimuth range bar) Target Range Mark Maximum Range Mark Minimum Range Mark	Appears on right side of HUD during A/A modes. Scaling is determined by RIO selecting RANGE pushbuttons on the DDD panel. On the VDI, the range scaling is indicated in the lower left corner of the display. Limits are ±35° in horizontal. The symbol appears on the left side of the VDI during A/A modes, except in STT when it will be located at target azimuth. Indicates range to target. Indicates maximum range for weapon launch. Indicates minimum range for weapon launch.
	Upper Solution Cue	Cue is a measure of instantaneous weapon range. It is constrained to motion on the bomb fall line. Displayed with respect to velocity vector to indicate range-to-go to weapon release. When it crosses the velocity vector symbol the computer commands a weapon release.
	Lower Solution Cue	Cue is a measure of maximum range of weapon calculated from instantaneous aircraft position, constrained to motion on the bomb fall line and is used in conjunction with velocity vector to indicate range-to-go-to in range. In range when cue crosses velocity vector and indicates weapon can reach target if pilot executes a pullup. Cue appears after designate.
	Pullup Cue	Cue is a measure of range at which a 4 g pullup is required to clear weapon fragmentation pattern or ground. Positioned directly below velocity vector. Used in conjunction with velocity vector to indicate range-to-go to minimum safe pullup point, when one crosses the other.

Figure 7-10(B).-Symbology for operational modes of the AVA-12 continued.

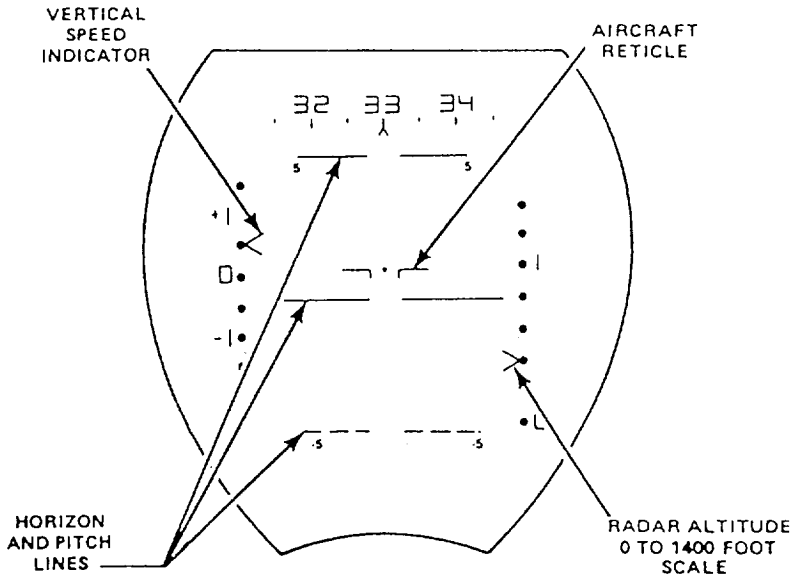


Figure 7-11.-Takeoff mode.

DECLUTTER

The AVA-12 has a declutter feature. This feature is used to remove preselected unwanted symbols from the pilot's field of view and head-up display during certain modes of operation. An "uncluttered" display is especially important during air-to-air combat situations.

DISPLAY MODES

The AVA-12 has five basic modes of operation. These modes are takeoff, cruise, air-to-air (A/A), air-to-ground (A/G), and landing.

Takeoff

The takeoff mode (fig. 7-11) displays vertical speed, altitude, and aircraft attitude.

Cruise

The cruise mode (fig. 7-12) is used primarily for flight from one point to another. The display gives the pilot attitude reference, magnetic heading, and various weapons information.

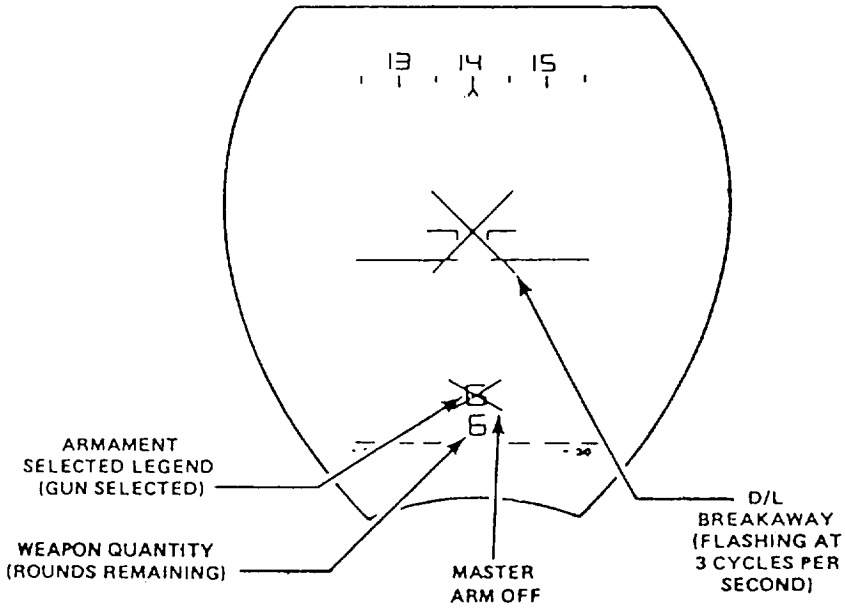


Figure 7-12.-Cruise mode.

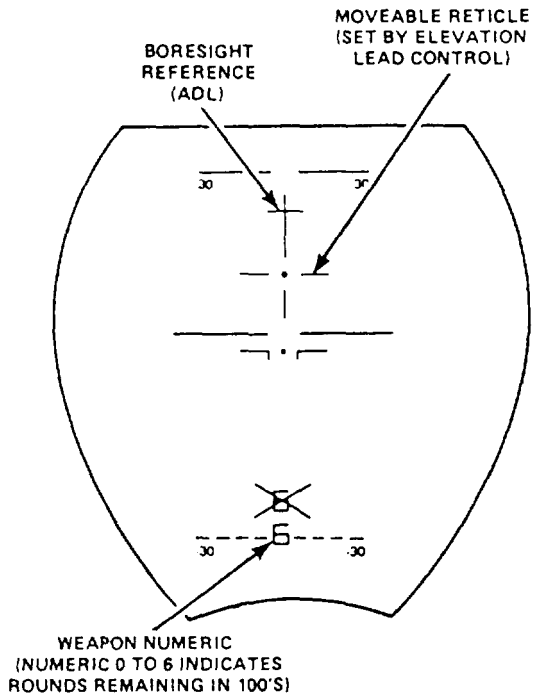


Figure 7-13.-Gun mode.

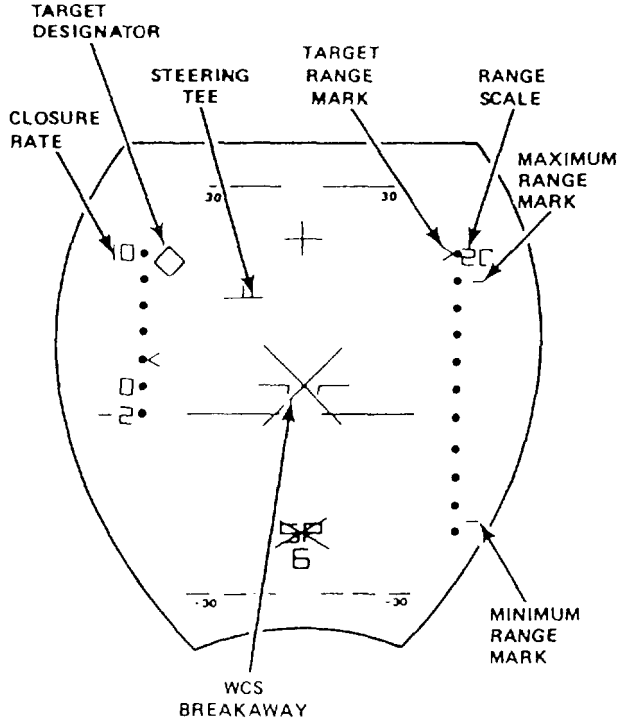


Figure 7-15.-Sparrow normal mode.

Air-to-Air

The air-to-air mode (figs. 7-13 through 7-16) gives the pilot attack data with the various types of air-to-air weapons selected. In addition to attitude reference, the weapon type and number of weapons on board are displayed along with various types of target information.

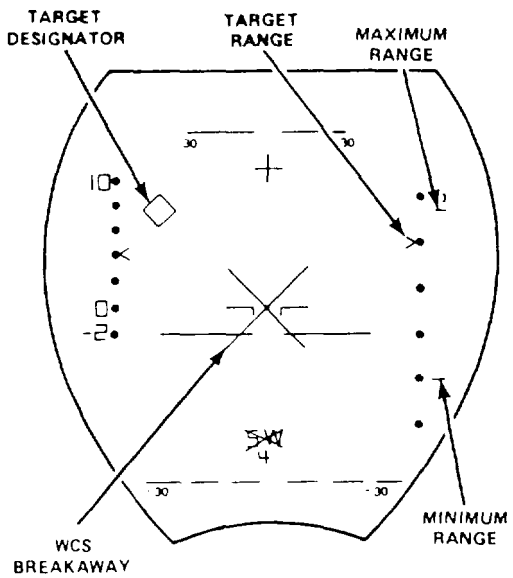


Figure 7-14.-Sidewinder normal mode.

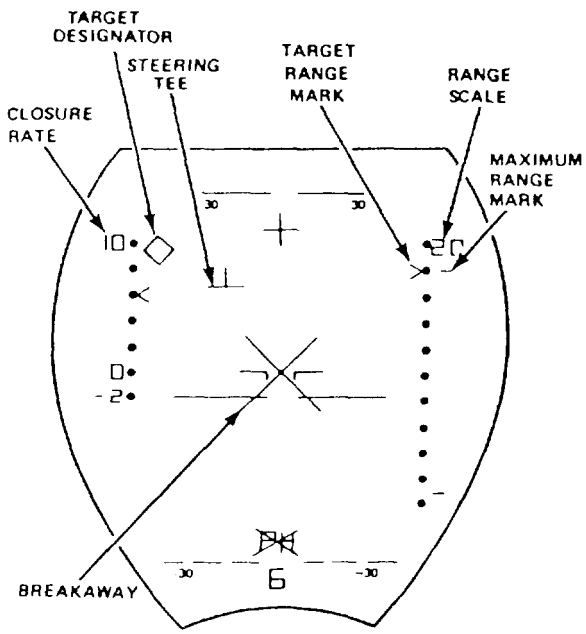


Figure 7-16.-Phoenix normal mode.

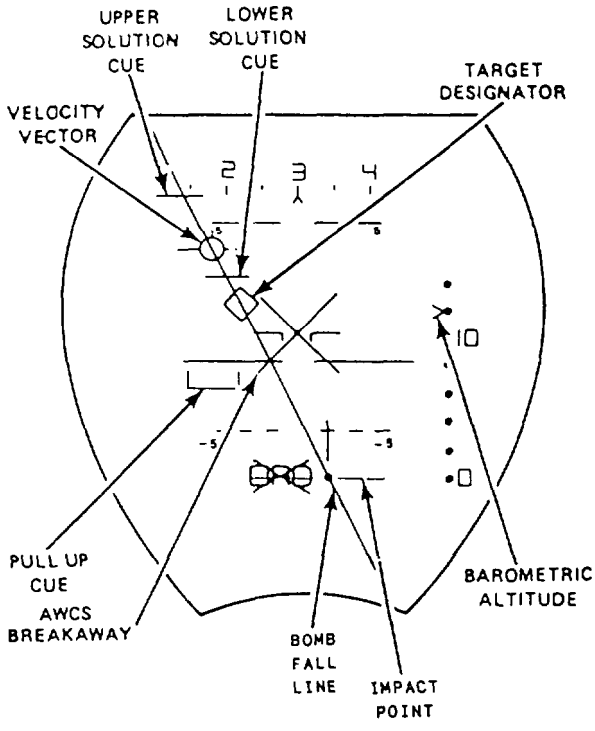


Figure 7-17.-A/G computer target mode.

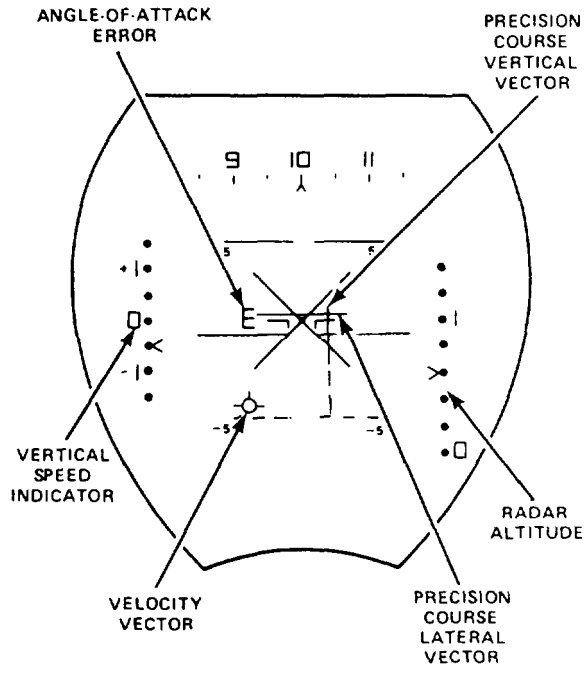


Figure 7-18.-Landing mode.

Air-to-Ground

The air-to-ground mode (fig. 7-17) is used primarily to deliver bombs or other types of air-to-ground weapons. Along with navigation information, it displays bomb ballistic information along with target information.

Landing

The landing mode (fig. 7-18) gives the pilot altitude, vertical descent, angle of attack, and velocity information. This mode is used when making an approach to an airfield or a carrier until the pilot has visual contact with the ground.

THERMAL IMAGING

Learning Objective: *Recognize functions, characteristics, components, and operating principles of thermal or infrared (IR) imaging.*

Before reading this section, a basic knowledge of IR detection principles is essential to understanding thermal imaging and the FLIR system. If needed, you should review IR detection principles in *Aviation Electronics Technician 3,* NAVEDTRA 12329.

Humans can see only a small part of the entire electromagnetic spectrum; however, other parts of the spectrum contain useful information. The infrared spectrum is a small portion of the entire electromagnetic spectrum.

Infrared radiation is also known as thermal or heat radiation. Most materials emit, absorb, and/or reflect radiation in the IR region of the electromagnetic spectrum. For example, an aircraft parked on a sunlit runway absorbs and radiates varying amounts of IR radiation. After sunset, the aircraft continues to radiate the absorbed heat, making detection at night possible. Even if the aircraft is moved, detection of the aircraft is possible because the runway surface, which was directly below the aircraft, will be cooler than the surrounding runway.

Thermal imaging is referenced in terms of temperature instead of reflectivity (radar) or color (visible light). Variations of the temperature in a scene tend to correspond to the details that can be visually detected. It is the function of the IR imaging system to process this information and convert it into information that the system operator can use. Currently, the types of imaging systems generally used are mechanical-scanning, fast-framing devices. They use the frame rate (information update rate) that is similar to television. They are known as forward looking infrared (FLIR) devices.

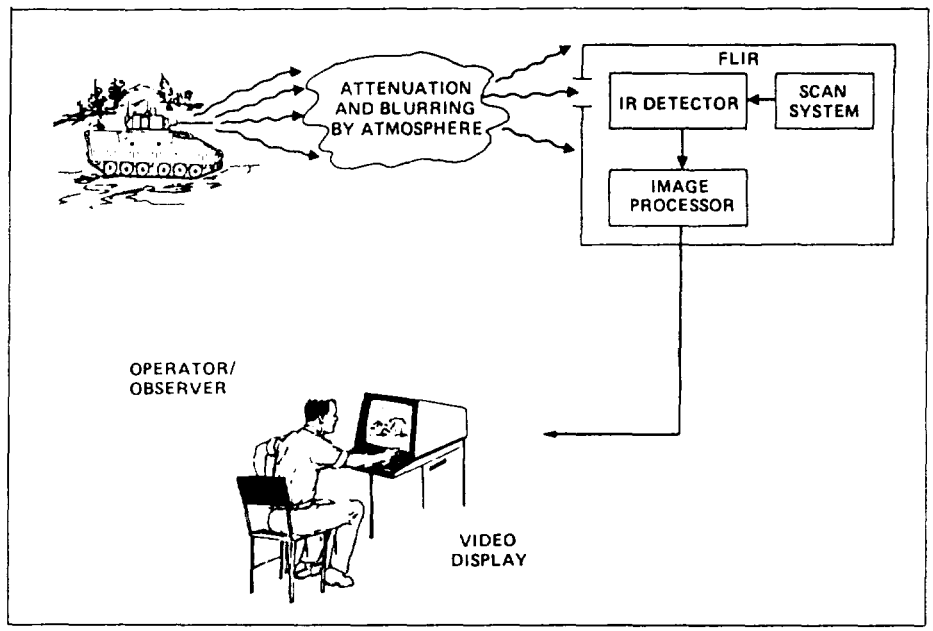


Figure 7-19.-Thermal imaging.

Before a target can be detected, it must exchange energy with its environment, be self-heating, have emissivity differences, and reflect other sources. Look at figure 7-19. Notice that the atmosphere between the target and the FLIR attenuates (weakens) and blurs the target signal. The FLIR operator aims the limited-field-of-view FLIR to search the scene for targets, using a search pattern and clues, such as radar targets or laser designators.

Thermal sensitivity, image sharpness, spectral response, contrast, and magnification are used in the FLIR system to produce a visual image of the thermal scene. The operator uses training, experience, and image interpretation skills to detect and identify targets.

INFRARED RADIATION

The atmosphere is a poor transmitter of infrared radiation because of the absorption properties of CO_2 (carbon dioxide), H_2O (water), and O_3 (ozone). Infrared radiation is broken into four regions by characteristics: near, middle, far, and extreme (fig. 7-20). Only the first

NAME		WAVELENGTH (μM)	FREQUENCY (HZ)
NEAR INFRARED	NIR	0.7 μM – 3 μM	$4.3 \times 10^{15} - 1 \times 10^{14}$
MIDDLE INFRARED	MIR	3 μM – 6 μM	$3.0 \times 10^{14} - 5 \times 10^{13}$
FAR INFRARED	FIR	6 μM – 15 μM	$5.0 \times 10^{13} - 2 \times 10^{13}$
EXTREME INFRARED	XIR	15 μM – 1000 μM	$2.0 \times 10^{13} - 3 \times 10^{11}$

Figure 7-20.-Characteristics of IR radiation.

three are discussed in this chapter. Figure 7-21 shows the transmission spectrum of the atmosphere. You can see that the best transmission is between 3m and 5m and between 8m and 14m. The range between these frequencies is known as a window. Infrared imaging devices are designed to operate in one of the two windows, usually the 8m and 14m.

Infrared Radiation Sources

All matter with a temperature above -273°C (absolute zero) emit IR radiation. The amount of the IR radiation emitted is a function of heat. Theoretically, a perfect emitter is a black body with an emissivity of 1. Realistically, the best emissivity is somewhere around .98. The emissivity of various objects is measured on a scale of 0 to 1.

The total energy emitted by an object at all wavelengths is directly dependent upon its temperature. If the temperature of a body is increased 10 times, the IR radiation emitted by the body is increased 10,000 times. If the energy emitted by a black body and its wavelengths is plotted on a graph, a hill-shaped curve results (fig. 7-22). By looking at this graph, you can see that the energy emitted by short wavelengths is low. As the wavelengths get longer, the amount of energy increases up to a peak amount. After the peak is reached, the energy emitted by the body drops off sharply with a further increase in wavelength.

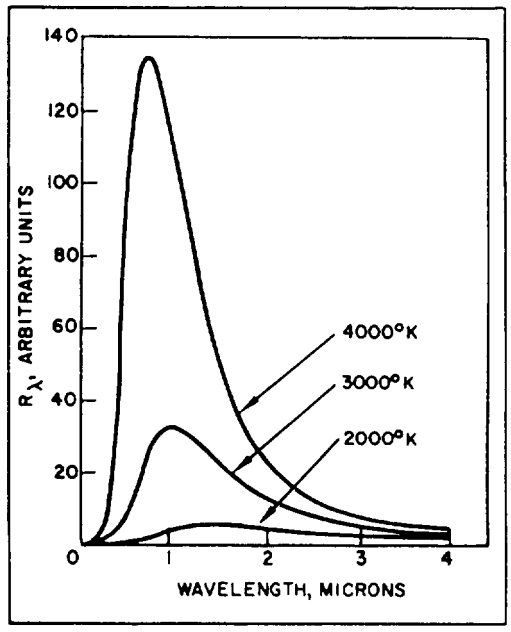


Figure 7-22.-Black body radiation.

Infrared Optics

Many of the materials commonly used in visible light optics cannot be used in IR imaging systems because, at IR frequencies, these materials are opaque. The optical materials used in IR imaging systems should have a majority of the following qualities:

- Be transparent at the wavelengths on which the system is operating
- Be opaque to other wavelengths

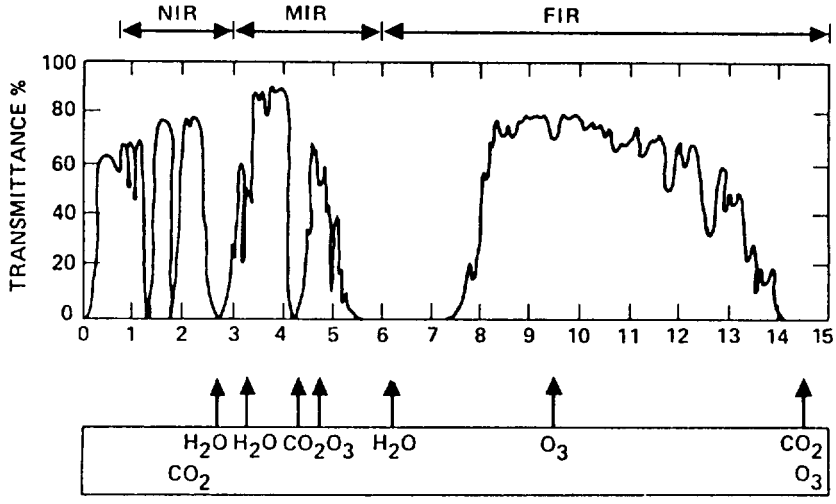


Figure 7-21.-Transmission spectrum of the atmosphere.

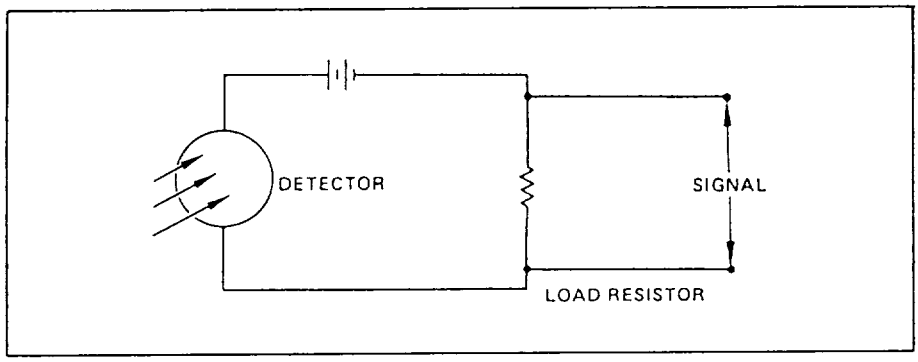


Figure 7-23.-Photoconductive detector circuit.

- Have a zero coefficient of thermal expansion to prevent deformation and stress problems in optical components
- Have high surface hardness to prevent scratching the optical surfaces
- Have high mechanical strength to allow the use of thin lenses (high-ratio diameter to thickness)
- Have low volubility with water to prevent damage to optical components by atmospheric moisture
- Be compatible with antireflection coatings to prevent separation of the coating from the optical component

None of the materials currently used for IR optics have all of these qualities; however, silicon, germanium, zinc selenide, zinc sulfide, and IRTRAN have many of them.

Infrared Detectors

The detector is the most important component of the IR imaging system. There are many types of detectors, each having a distinct set of operating characteristics. Bolometers, Golay cells, mercury-doped germanium, lead sulfide, and phototubes are the most commonly used types of detectors. Detectors can be characterized by their optical configuration or by the energy-matter interaction process. There are two types of optical configurations-elemental and imaging.

ELEMENTAL DETECTORS.– Elemental detectors average the portion of the image of the outside scene falling on the detector into a single signal. To detect the existence of a signal in the field of view, the detector builds up the picture by sequentially scanning the scene. The elemental detector requires time to develop the image because the entire scene must be scanned.

IMAGING DETECTORS.– Imaging detectors yield the image directly. An imaging detector is considered a myriad of point detectors. Each of the detectors responds to a discrete point on the image. Therefore, the imaging detector produces the entire image instantaneously. A good example of an imaging detector is photographic film.

Energy-Matter Interaction

There are two basic types of energy-matter interaction; the thermal effect and the photon effect.

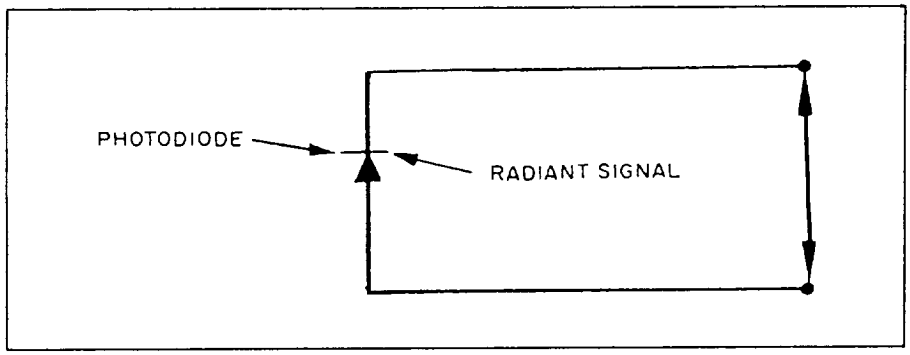


Figure 7-24.-Photoelectric effect.

THERMAL EFFECT.– The thermal effect type of energy-matter interaction involves the absorption of radiant energy in the detector. This results in a temperature increase in the detector element. The radiation is detected by monitoring the temperature increase in the detector. Both the elemental and imaging forms of detectors use the thermal effect.

PHOTON EFFECT.– In the photon effect type of energy-matter interaction, the photons of the radiant energy interact directly with the electrons in the detector material. Usually, detectors using the photon effect use semiconductor material. There are three specific types of photon effect detection.

1. Photoconductivity. Photoconductivity is the most widely used photon effect. (See fig. 7-23.). Radiant energy changes the electrical conductivity of the detector material. An electrical circuit is used to measure the change in the conductivity.

2. Photoelectric (also referred to as photovoltaic). In the photoelectric effect (fig. 7-24), a potential difference across a PN junction is caused by the radiant signal. The photocurrent (current generated by light) is added to the dark current (current that flows with no radiant input). The total current is proportional to the amount of light that falls on the detector.

3. Photoemissive. The photoemissive effect (fig. 7-25) is also known as the external photo effect. The action of the radiation causes the emission of an electron from the surface of the photocathode in the surrounding space. The electron is photoexcited from the Fermi level above the potential barrier at the surface of the metal.

INFRARED IMAGING SYSTEMS

The infrared imaging system has the following components: detectors, a scene dissection system, front end optics, a refrigeration system (if required), and an image processing system.

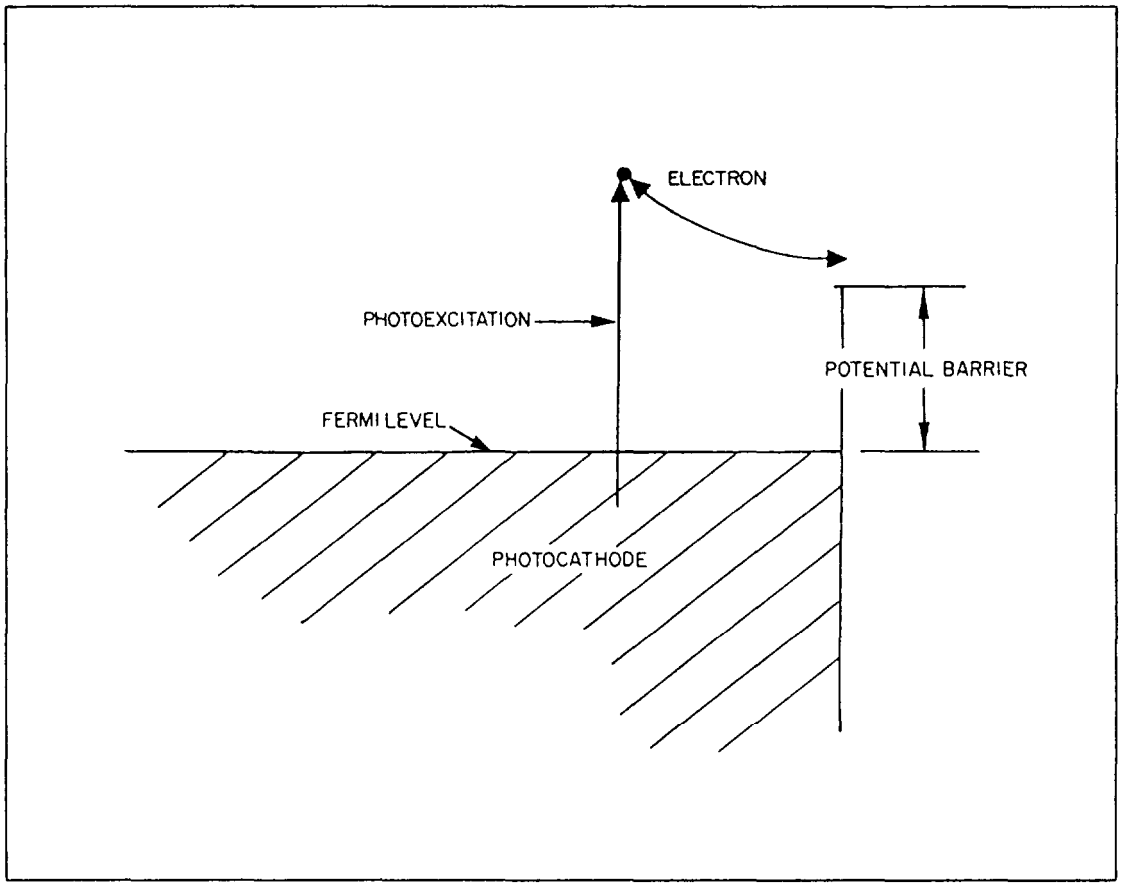


Figure 7-25.-Photoemissive effect.

Detectors

Detectors convert the IR radiation signal into an electrical signal that is processed into information used by the operator. Detectors can be arranged in many different configurations for their use in IR imaging systems.

DETECTOR ARRAY.– Only a small portion of the image scene is taken by a detector (or detectors) to achieve maximum resolution. A large number of detector elements can be grouped together to form an array (fig. 7-26, view A). The elements of this array are packed closely together in a regular pattern. Thus, the image of the scene is spread across the array like a picture or a mosaic; each detector element views a small portion of the total scene. The disadvantage of this type of system is that each detector element requires a supporting electronic circuit to process the information that it provides. Also, each detector element requires a preamplifier to boost the signal to a usable level.

SINGLE DETECTOR.– Another method that is used to provide the operator with information is the single detector (fig. 7-26, view B). Here, there is one detector requiring one set of supporting circuitry. In this type of system, the image is scanned across the detector so that the detector can see the whole image. An optical system is required that can supply the scanning. This type of system is adequate if real-time information is not needed or if the object of interest is stationary or not moving quickly.

Scene Dissection System

The scene dissection system is used to scan the scene image. There are many types of scanning-one associated with each type of detector array. When a

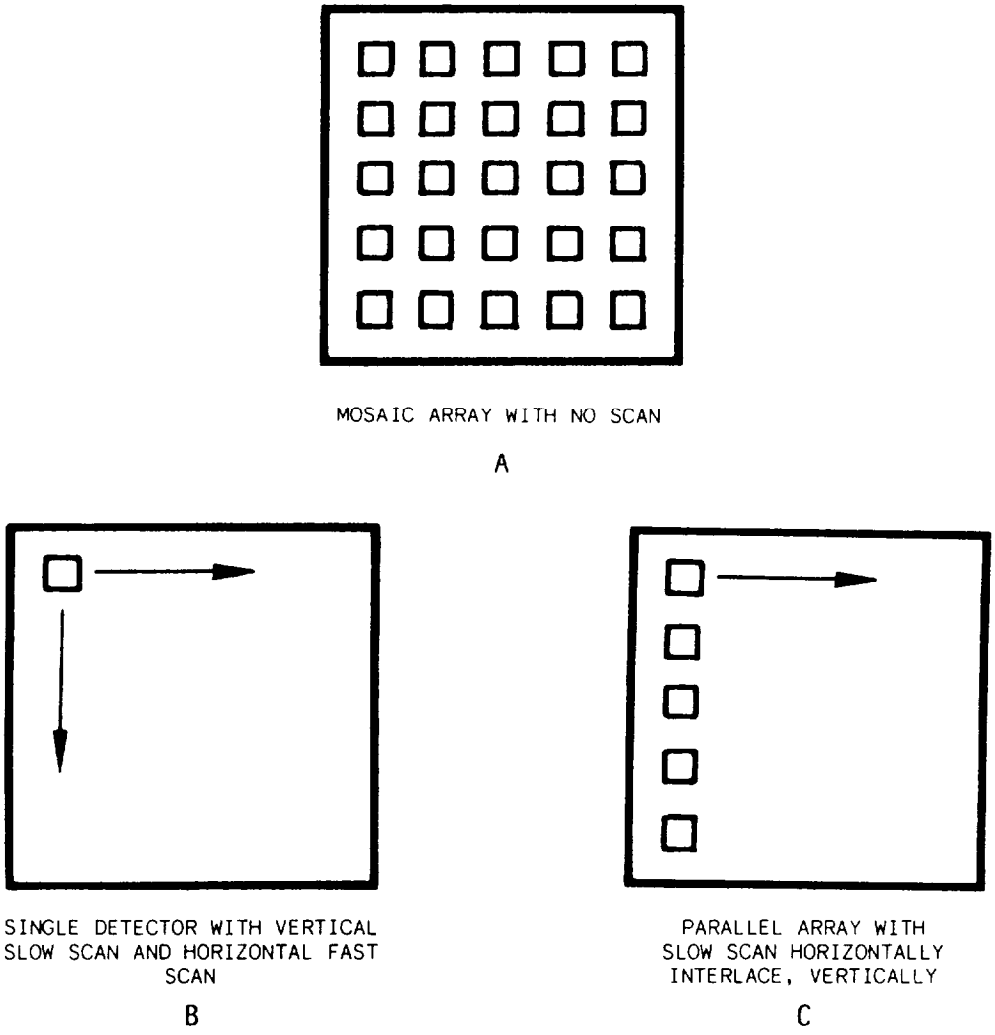


Figure 7-26.-Detector arrays.

single detector with one axis of fast scan and one axis of slow scan is used, the scene is scanned rapidly in the horizontal direction and slowly in the vertical direction. As a result, the line is scanned horizontally; then the next line is scanned horizontally, and so forth.

A vertical linear array is scanned rapidly in the horizontal direction. One detector element scans one line of the image. In the linear array, there is a space, one element wide, between each element. The scan is one axis with an interlace being used. A vertical linear array is scanned rapidly in the horizontal direction. After each horizontal scan, the mechanism shifts the image upward or downward one detector element width so that on the next scan, the lines that were missed are covered.

Each system has an optimum configuration of detector array and image dissection. If the number of elements in the detector are increased, the system becomes more complicated. The cost of the system is increased, and the reliability of the system is decreased. If the number of detectors is decreased, the amount of information that can be processed is reduced. A compromise between a large number of elements (increased cost) and a smaller number of elements (reduced information) is to use a linear array that is scanned in one direction only. Each detector scans one line of the scene image. The complexity of the electronics is reduced, and the amount of information that is processed is increased. Thus, the size of the scene to be viewed and the detail of the scene is increased.

There are many types of mechanisms that can be used to scan the scene. When scanning using two axes, the two scanning motions must be synchronized. The electronic signal that controls the sampling of the detectors must also be synchronized with the scanning motions.

Front End Optics

The front end optics collect the incoming radiant energy and focus the image at the detectors. The optics may be reflective or refractive, or a combination of both. Many systems offer a zoom capability, allowing a continuous change in magnification of the image without changing the focus. Spectral filters are used to restrict the wavelength of light entering the system. This prevents unwanted wavelengths of light from reaching the detector and interfering with the imaging process.

Refrigeration System

A refrigeration system is needed in imaging systems because many types of infrared detectors require low temperatures if they are to operate properly. The two types of detector coding that are used are the open-cycle and closed-cycle.

In the open-cycle type of cooling, a reservoir of liquified cryogenic gas is provided. The liquid is forced to travel to the detects, where it is allowed to revert to a gas. As it changes from a liquid to a gas, it absorbs a great deal of heat from the surrounding area and the detector.

In the closed-cycle type of cooling, the gas is compressed, and the heat generated by the compression is radiated away by the use of a heat exchanger. The gas is then returned to the compressor and the cycle repeats itself.

Image Processing Systems

The image processing system is used to convert the data collected by the detectors into a video display. Data from the detectors is multiplexed so that it can be handled by one set of electronics. Then it is processed so that the information coming from the detectors is in the correct order of serial transference to the video display. At this point, any other information that is to be displayed is added.

In other image processing systems, the signals from the detectors are amplified and sent to an LED display, or they are optically amplified by photomultiplier tubes and projected on a phosphorescent screen.

FORWARD LOOKING INFRARED SYSTEM

Learning Objective: *Recognize components and operating principles of a typical forward looking infrared (FLIR) system.*

A forward looking infrared (FLIR) system is an infrared detecting set (IRDS). The IRDS is a passive device that operates on the infrared (IR) principles of emissivity. The terms FLIR and IRDS are synonymous so far as system operation is concerned. Only the azimuth coverage differs. The FLIR system scans an operator-selected portion of the terrain along the aircraft's flight path and displays a televised image of the IR patterns of the terrain. The primary function of the FLIR system is to give the operator an improved capability to detect, identify, and classify targets of interest that would otherwise be concealed from either visual observation or radar detection. The concealment could be due to darkness or camouflage or from radar

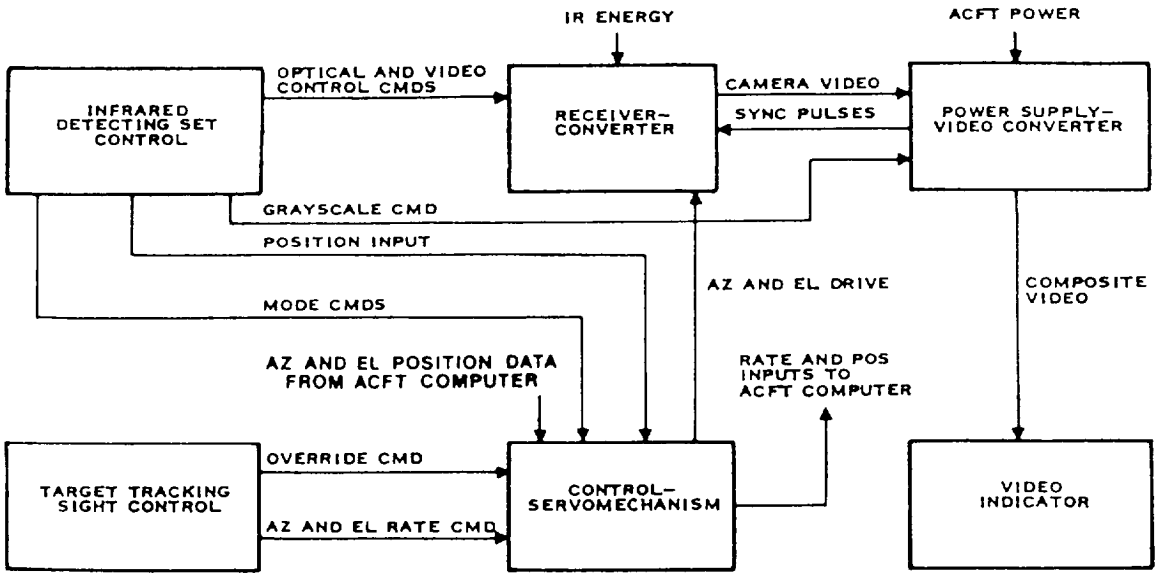


Figure 7-27.-FLIR system block diagram.

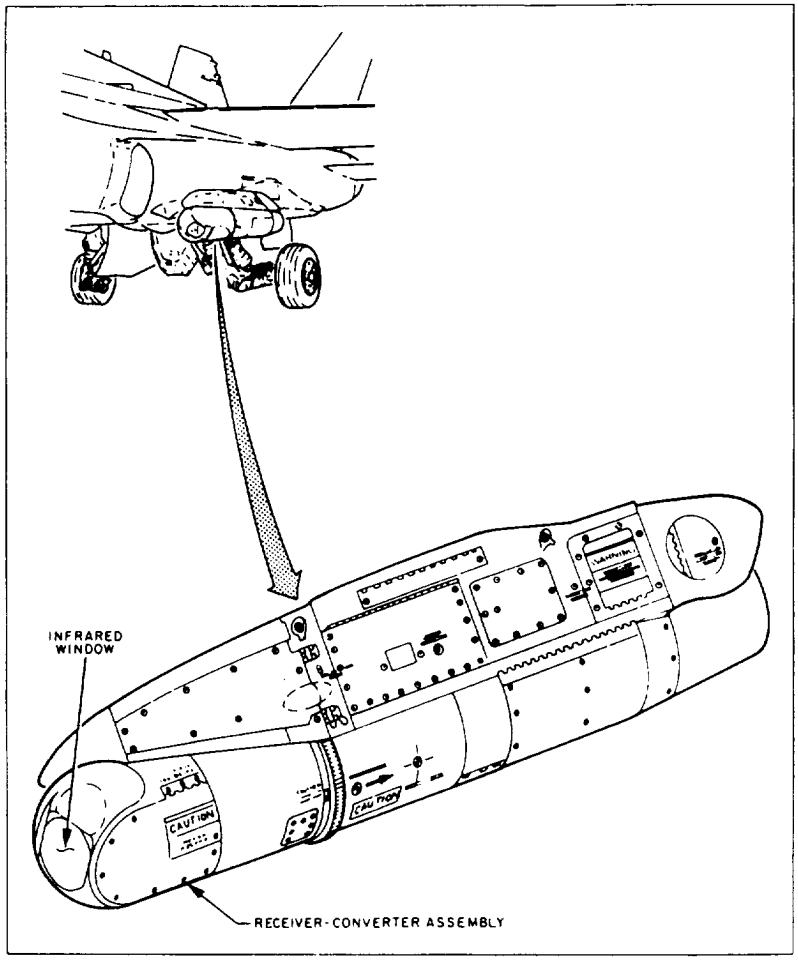


Figure 7-28.-Station-mounted FLIR pod.

detection due to extreme scope clutter caused by inclement weather, rough seas, or electronic jamming. Additionally, the FLIP system emits no transmission for detection by an enemy. Although there are various types of FLIR systems used in the Navy, their principles of operation are basically the same.

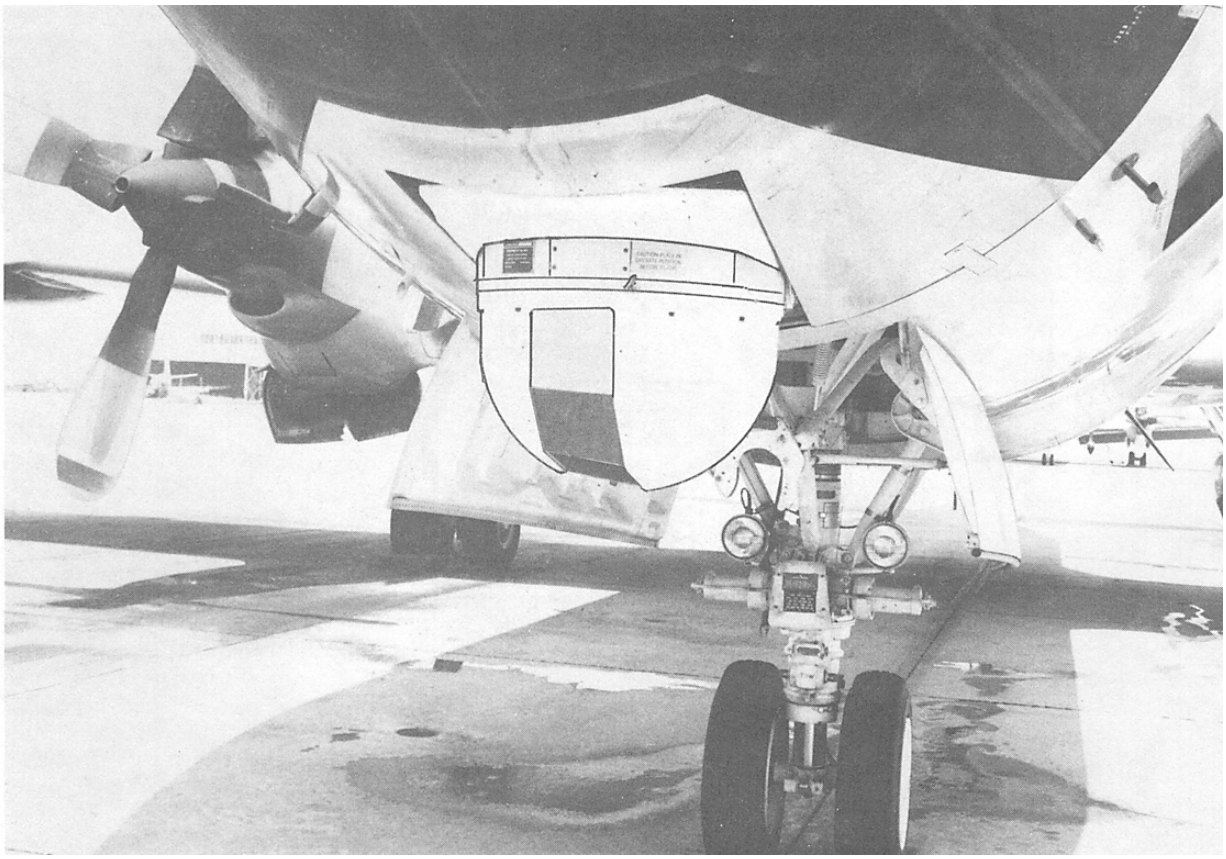
A typical FLIR system is comprised of weapons replaceable assemblies (WRAs). The WRA nomenclatures are as follows: receiver-converter, power supply-video converter, control servomechanism, target tracking sight control, infrared detecting set control, and video indicator. Figure 7-27 shows the block diagram of a typical FLIR system. You should refer to the diagram as you read about the FLIR system assemblies.

RECEIVER-CONVERTER ASSEMBLY

The receiver-assembly contains all of the optics and electronics used to detect and convert IR energy into a single-channel video output to be processed by the power supply video converter assembly. The processed video output is applied to the video indicator. The receiver-converter assembly contains gyros, gimbals, and drive motors to aim and stabilize the receiver section in azimuth and elevation. It also contains a heat exchanger to circulate conditioned air throughout the assembly and a refrigerator unit to keep the IR detectors cooled to cryogenic temperatures. The receiver-converter assembly is housed in either the forward section of a station-mounted FLIR pod (fig. 7-28) or in its own pod, mounted on the forward lower part of the aircraft's fuselage (fig. 7-29). The housing used depends on the aircraft model.

Functionally, the receiver-converter breaks down into four functionally subsystems as follows:

1. IR to composite video conversion
2. Temperature control
3. Positioning and stabilization
4. BITE (built-in test equipment)



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Figure 7-29.—Receiver-converter pod shown in operating position.

IR to Composite Video Conversion Subsystem

Figure 7-30 is a diagram showing the optics and electronic components of a typical receiver-converter required to perform IR detection and conversion into a usable video signal. Although the signal optical path and conversion coincide, each path is discussed separately for simplicity. You should refer to figure 7-30 as you read the following paragraphs.

SIGNAL OPTICAL PATH.– Incoming IR energy from a target enters the receiver through a window (fig. 7-28) and strikes one side of a double-sided scan mirror. This mirror is controlled by a scanner module. The IR signal either strikes the mirror directly or is focused onto the mirror by optical lenses contained in what is called an afocal optics unit. The operation (in-out) of the afocal unit is governed by a field of view (FOV) switch on the operator's infrared detecting set control (IRDSC). In the wide FOV mode of operation, the afocal optic lenses are not in the signal optical path. In the narrow FOV mode of operation, the lenses are in the path shown in figure 7-30. The lenses are focused by a motor that is controlled by a focusing module in the receiver-converter assembly. The focusing module has two inputs to control the operation of the focusing motor. One input is a target range scale set by the operator on the IRDSC. The other

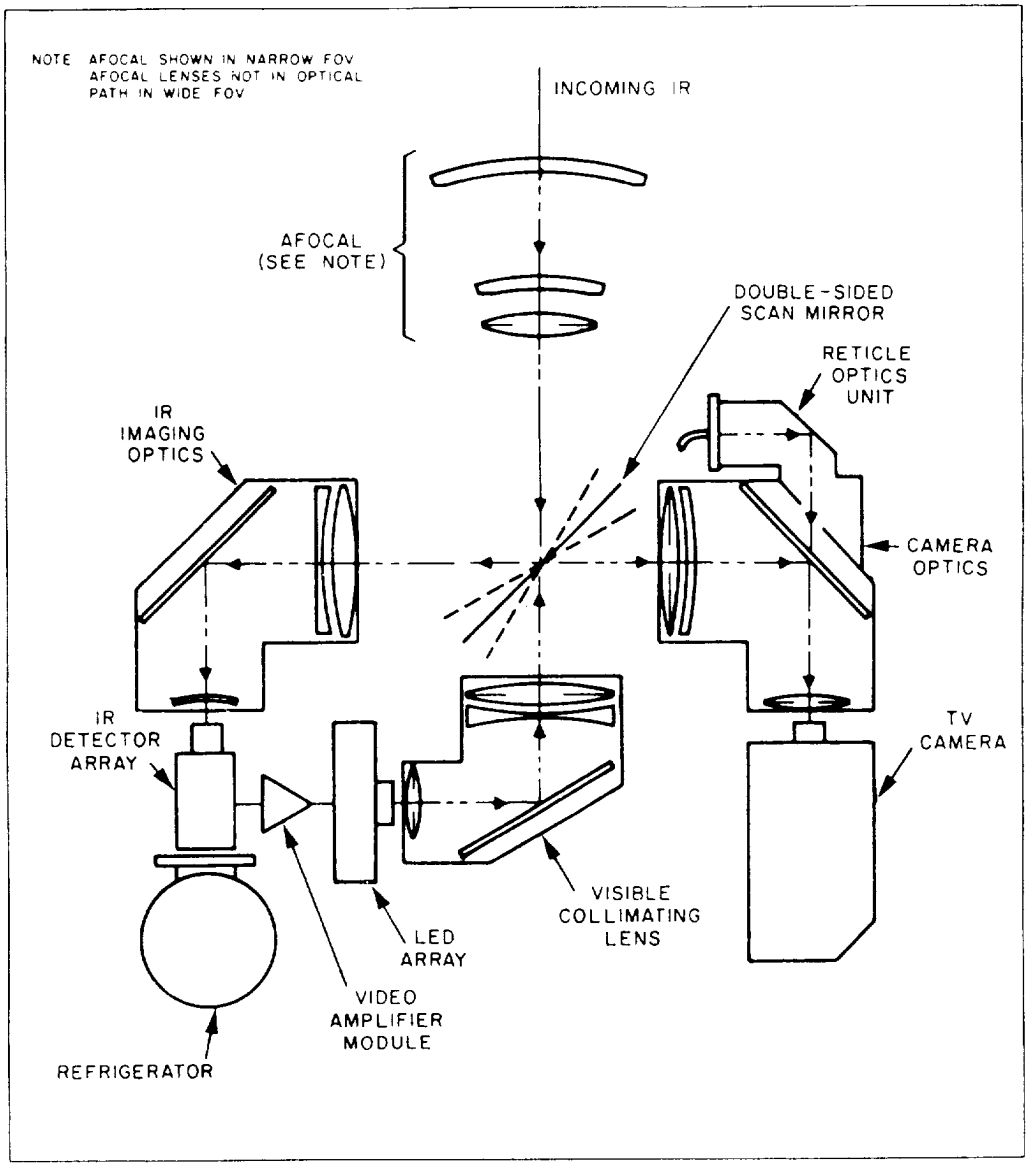


Figure 7-30.-FLIR optical path and IR processing.

input is a feedback signal from a temperature-sensing circuit in the afocal optics unit. Because the index of refraction of an optical lens changes with changes in temperature, the focusing module monitors the temperature of the lens and maintains proper focusing of the IR signal onto the scan mirror.

The scan mirror is controlled by a scanner module. It is also positioned in line-of-sight (LOS) position, along with the entire receiving head, by signals from the control servomechanism. As shown at the top of figure 7-30, one side scans the incoming IR energy and reflects the signals into the IR imaging optics, while the bottom side simultaneously scans visible light signals from the collimating lens and reflects the signal into the TV camera optics. The mirror scans the horizontal axis and is indexed in the vertical axis to provide a 2 to 1 ratio interlace scan. The mirror scan is synchronized to the TV camera by sync signals.

IR TO VIDEO PROCESSING.– The IR signals reflected from the scan mirror into the imaging optics are focused onto an IR detector array located behind an imaging lens. The lens is kept in focus much the same as the afocal lenses. The IR detector array consists of approximately 180 individual detectors arranged in a linear array with a space between each to allow for 2:1 interlacing. The scan mirror scans the image across the detectors, and each detector produces a single horizontal line of IR video. The scan mirror is indexed one line width in the vertical direction, making a total of 360 lines of video with only 180 detectors and amplifier channels. Because of the 2:1 interlacing, two full scans of the mirror are required to reproduce the entire image. Each detector conducts according to the amount of IR energy impressed on it from the scan mirror. The resultant output of the IR detector array is 180 parallel signals representing 360 video lines of IR energy scanned by the scan mirror.

The IR detectors are kept at cryogenic (very low) temperatures by the refrigerator unit; they are biased to process the incoming IR energy into a usable multichannel IR video signal. The low-level output of the detector array is fed to a video amplifier module. The module consists of one preamplifier and three postamplifiers for each of the IR detectors in the array. This is required to increase the IR signal to a usable level. The output of the postamplifiers are controlled by a dc level (+ or -) whose polarity is controlled by a polarity switch on the IRDSC. The purpose of the dc polarity is to have “hot” IR targets appear either white or black on the video indicator. Also, the outputs of the video amplifiers are gated on or off in synchronization with the scan mirror and the TV camera sweep.

Each video amplifier feeds a light-emitting diode (LED) of the LED array. The LED array duplicates the IR detector array; the visible light intensity output of one LED is proportional to the IR output of the corresponding IR detector in the detector array. The resultant output of the LED array comprises 180 parallel, visible light beams (signals) representative of the IR energy scanned by the scan mirror. In other words, the resultant output of the LED array represents the terrain or targets scanned.

The output of the LED array is applied to the collimating lens unit. This unit focuses the visible light while maintaining the light beams parallel to each other rather than converging them to a focal point, as shown in figure 7-31, views A and B.

The output of the collimating lens unit is scanned by the back side of the scan mirror, and is reflected into the camera optics unit that focuses the light for the TV vidicon camera. Also, a reticle light signal from the

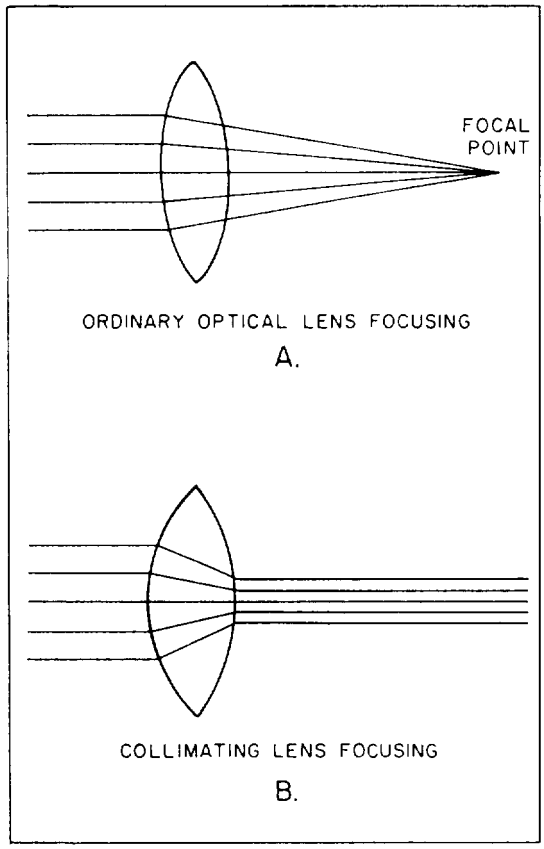


Figure 7-31.-Lens focusing patterns.

reticle optics unit is applied simultaneously to the TV camera to give an indication of the position of the receiving head. The TV vidicon camera processes the visible light signals into a signals channel video signal. The video output of the camera is fed to the power supply video converter assembly for further processing.

Temperature Control Subsystem

Figure 7-32 is a block diagram of a receiver-converter heat exchanger. The heat exchanger supplies conditioned air to the receiver-converter assembly for environmental control. The exchanger shown consists of two blowers, six heaters (three connected in a wye configuration and three connected in a delta configuration), three temperature-sensitive switches (usually mounted on the receiver casting), and an air-to-air heat exchanger. Blowers B1 and B2 circulate cooling air and heating air within the receiver-converter. Internal blower B1 operates whenever the FLIR system is turned on. If the temperature of the assembly compartment goes above approximately 77°F (25°C), thermal switch RT3 operates relay K3, applying power to external blower B2. Blower B2 draws cool external air through the air-to-air heat exchanger. Internal blower B1 circulates the cool air and cools the compartment. If the temperature in the compartment goes below 68°F (20°C), thermal switch RT2 operates relay K1, applying power to the wye-configured heaters. The heaters operate at approximately 365 watts. If the compartment temperature drops below 50°F (10°C), thermal switch RT1 operates relay K2, applying power to the delta-configured heaters. This increases the wattage used by the heaters to approximately 1,200 watts. Thus, the heat exchanger is able to maintain the receiver-converter compartment at a temperature between 68°F and 77°F at all times.

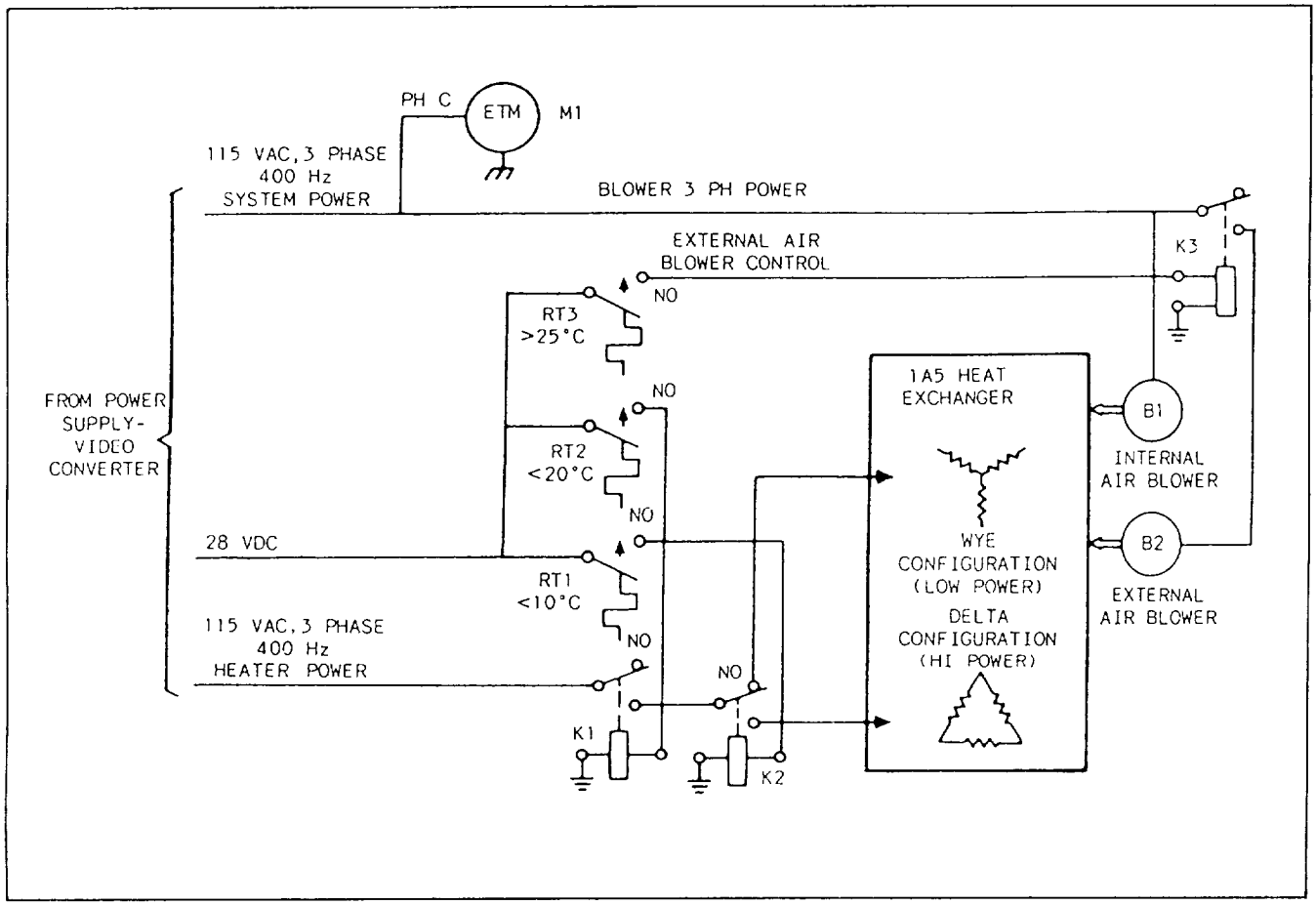


Figure 7-32.-Receiver-converter heat exchanger simplified block diagram.

Positioning and Stabilization Subsystem

The function of the stabilized gimbal subsystem of the receiver-converter is to allow the operator to select the line of sight (LOS) desired for the receiver unit, and to maintain a steady image of the IR patterns of the area viewed regardless of sporadic movement of the aircraft. This critical stability is obtained by means of gyros mounted on the receiver assembly. There is one azimuth and one elevation gyro. Figure 7-33 is a simplified block diagram of a typical FLIR positioning and stabilization subsystem.

The LOS of the receiver is selected in one of three ways:

1. By a program in the aircraft's computer (or a video tracking computer when the system is used with a laser system)

2. By inputs from the target tracking sight control

3. By operator adjustment of the azimuth and elevation controls on the control box

The azimuth and elevation commands from either input are processed in the control servomechanism weapons replaceable assembly (WRA); drive signals are applied to the receiver head positioning motors and gimbals. The receiving head is aligned to the desired LOS. If the aircraft should pitch or roll, the gyros (mounted on the receiver head) initiate error signals to a capture loop. This loop creates azimuth and elevation rate signals. The rate signals are fed back to the control servomechanism (CS). The rate signals are processed by the CS and are applied as azimuth and elevation drive signals to keep the receiver head at the correct LOS. In manual mode (which uses the target tracking sight control), the stabilization system is inhibited in the control servomechanism assembly.

Receiver-Converter BITE Subsystem

Most FLIR systems are equipped with a built-in test equipment (BITE) subsystem. The receiver-converter BITE subsystem monitors status signals from the camera video, refrigerator unit, gimbal, scan mirror and

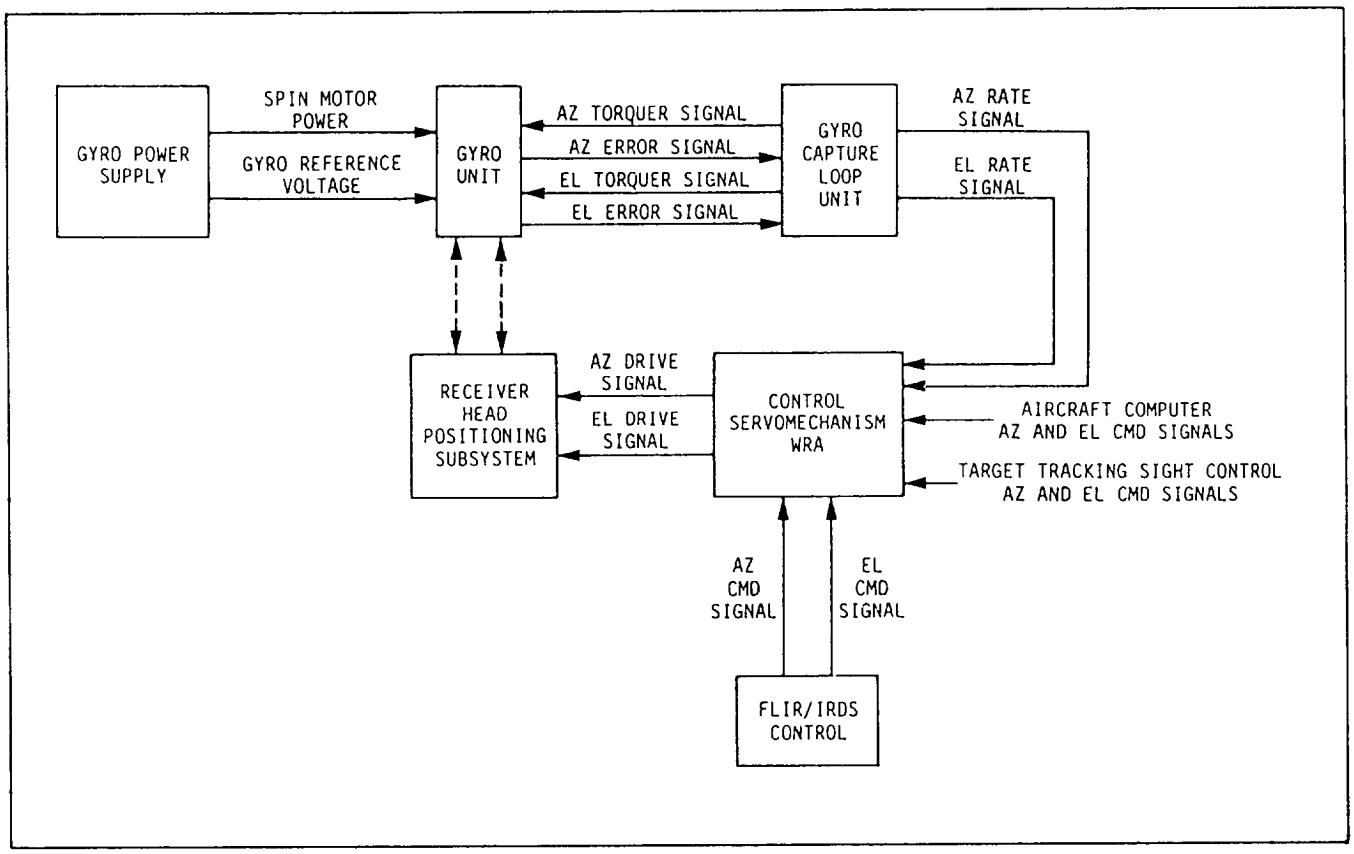


Figure 7-33.-FLIR position/stabilization system simplified block diagram.

TV camera scan focus drive of lenses, heat exchanger, afocal optics assembly, and the IR detectors. Figure 7-34 is a simplified block diagram of a receiver-converter BITE subsystem. The BITE signals go to a BITE logic module in the power supply video converter WRA, where they are combined and sent to the IRDSC to light various indicators.

POWER SUPPLY VIDEO CONVERTER ASSEMBLY

Functionally, the power supply video converter assembly breaks down into three subsystems–the power supply, video processing, and BITE subsystems.

The power supply subsystem is a typical power supply; it filters aircraft power for use by the receiver-converter assembly. It also develops all of the operating voltages for both the receiver-converter and power supply video converter circuits.

The video processing subsystem generates the synchronizing drive and timing signals for the receiver-converter. It also converts the TV camera video from the receiver-converter assembly into a composite video format consisting of camera video and receiver position information for presentation on the video indicator.

The BITE subsystem monitors and evaluates BITE signals from the receiver-converter, control

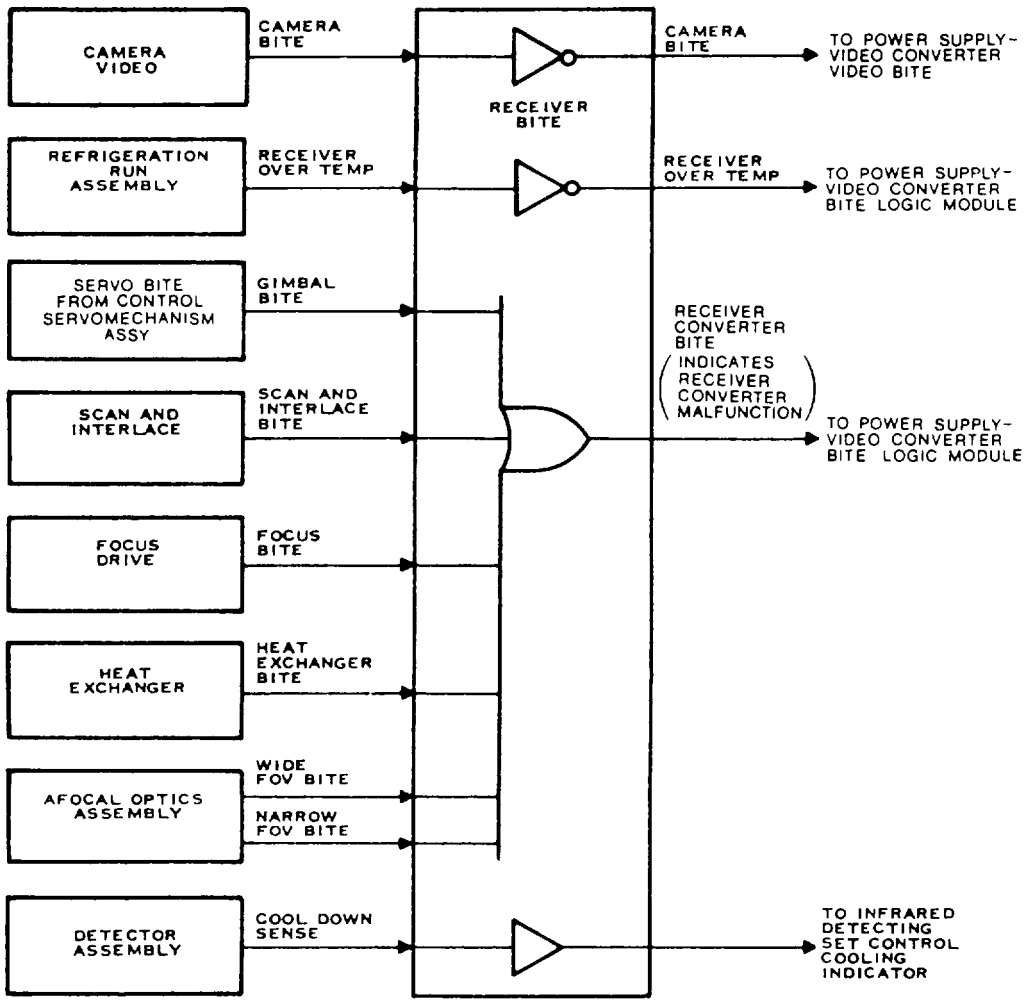


Figure 7-34.-Receiver-converter BITE subsystem simplified block diagram.

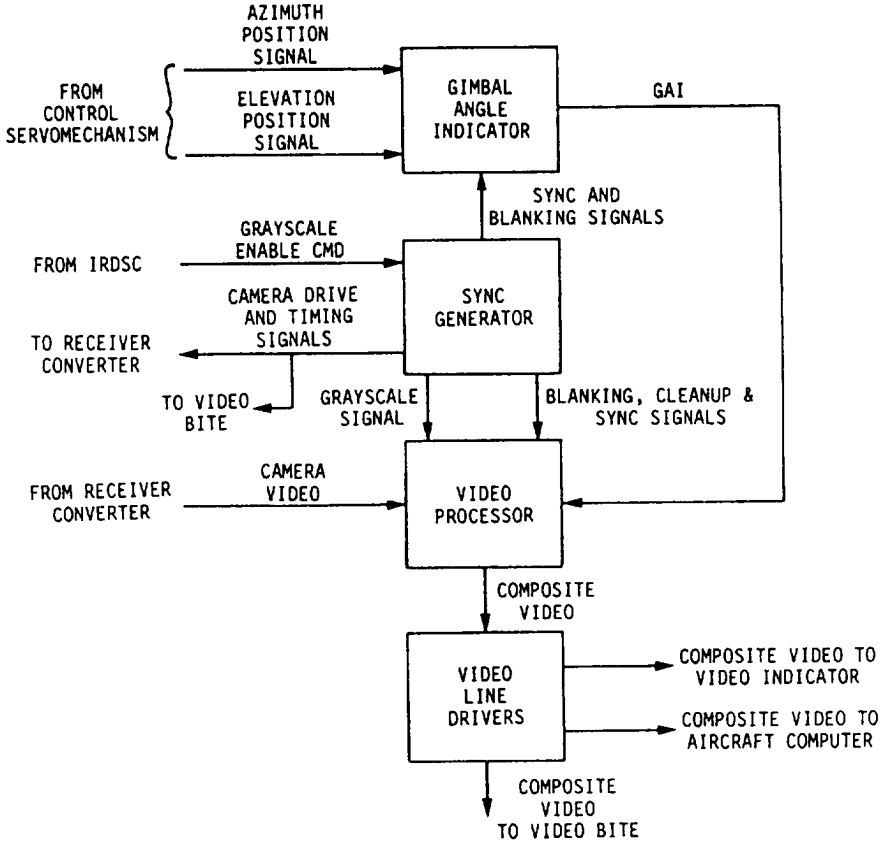


Figure 7-35.-Video converter video processing circuits simplified block diagram.

servomechanism, IRDSC, video, and power supply circuits within the assembly. These BITE signals initiate status-indicating signals for the IRDSC.

Video Processing

Figure 7-35 is a simplified block diagram of video processing circuits found in a FLIR power supply video converter WRA. The gimbal angle indicator unit receives linear signals from the control servomechanism azimuth and elevation gimbal potentiometers. These potentiometers (their inputs are from gimbals in the receiver-converter) generate synthetic video signals. The signals present short, bright line segments along calibrated scales on the video indicator (fig. 7-36) to indicate the receiver head position.

The sync generator module contains a crystal-controlled clock. The module generates all timing (sync), clamping, and drive signals for the receiver-converter and TV camera. It also generates all timing, gating, clamping and blanking signals for the

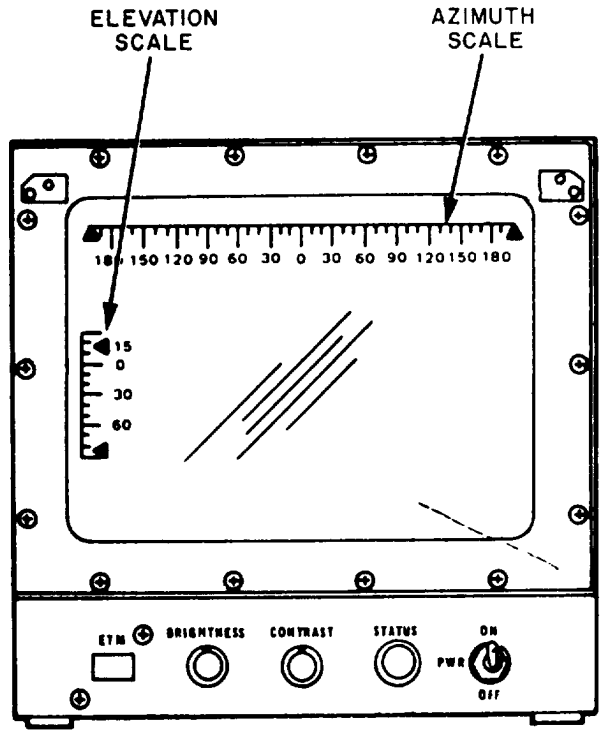


Figure 7-36.-Azimuth/elevation scales on a video indicator.

Table 7-3.-Sync Generator Outputs

SYNC GEN output	Unit Supplied
Beam gate only	TV camera
Cathode drive	TV camera
Dark current drive	TV camera
Horizontal deflection drive	TV camera
Vertical deflection drive	TV camera
Composite clamping	TV camera
Scanner sync	Scanning optic unit
Horizontal sync	Gimbal angle indicator
Composite blanking	Gimbal angle indicator, video processor
Vertical blanking	Gimbal angle indicator
Composite sync	Video processor
Grayscale signal	Video processor
Horizontal clamping	Video processor

video processor and gimbal angle indicator (GAI) modules. These signals are listed in table 7-3.

The sync generator module also generates a grayscale signal upon receipt of a grayscale command signal from the IRDSC when the grayscale switch is turned on. This grayscale signal presents a grayscale pattern (fig. 7-37) on the video indicator. The pattern serves to aid the operator in adjusting the level and gain controls on the IRDSC. There is a total of 10 different shades. Each shade represents a different IR temperature range to which the operator can compare the target intensity and estimate the temperature of the target. The temperature is indicative of the type of target material.

The video processor receives raw video signals from the TV camera. It also receives gimbal angle indicator (GAI) synthetic video, grayscale signals, and composite sync, composite blanking, and horizontal clamping signals from the sync generator. The video

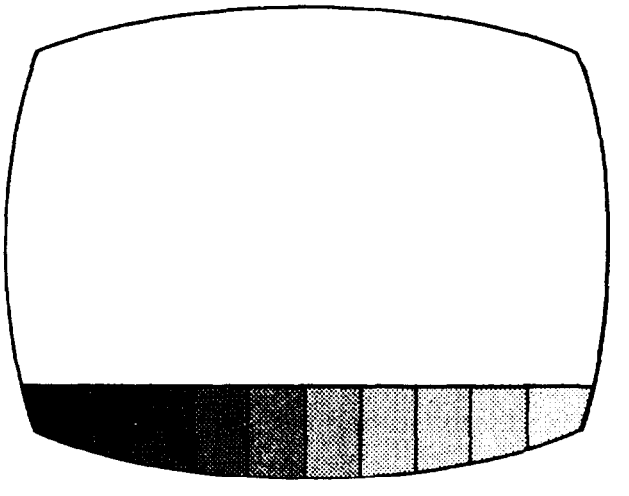


Figure 7-37.-Grayscale video indicator presentation.

processor combines all of these signals into a composite video signal, which is fed to the video line drivers.

The video line drivers amplify the composite video signal and provide three separate outputs. One output goes to the video indicator, one to the aircraft computer, and one to the video BITE module.

BITE Subsystem

Figure 7-38 is a simplified block diagram of the BITE subsystem of a power supply video converter WRA. When BITE is initiated on the IRDSC, a BITE command signal is outputted from the IRDSC and received by the BITE logic module. The signal causes the BITE logic module to generate and send a grayscale enable signal to the sync generator. This signal overrides the grayscale switch on the IRDSC and causes the sync generator to output a grayscale signal in addition to the sync, blanking, clamping, and drive signals it generates. BITE logic also sends a BITE initiate signal to the control servomechanism to initiate servo BITE and a signal (BITE ON) to the IRDSC to light the BITE ON light. This light indicates the BITE mode is operating.

The TV video BITE module monitors the output of the sync generator, the video line drivers, and the camera BITE from the receiver-converter WRA. If a failure occurs in these circuits, the video BITE module generates a TV fail signal to the power supply BITE module. If no failure (error) occurs, no TV fail signal is generated.

The power supply BITE module monitors all of the power supply voltages. If any voltage is not correct, a power supply malfunction signal is generated and sent to BITE logic, which sends a power supply fail signal to the IRDSC to light the power supply fail light. If power supply BITE receives a TV fail signal from video BITE, it generates a power supply malfunction signal that goes to the IRDSC to light the power supply fail light.

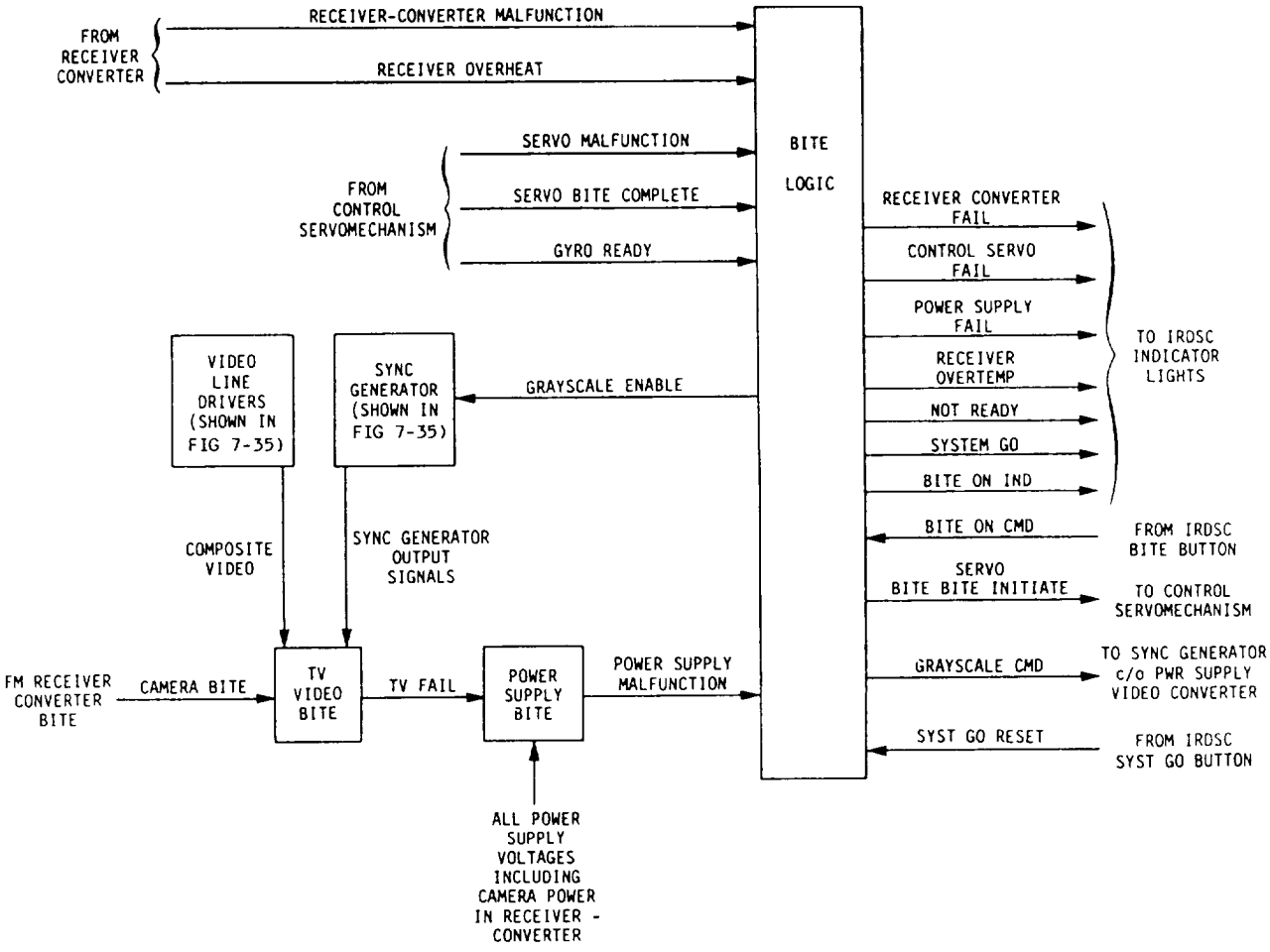


Figure 7-38.-Power supply video converter BITE simplified block diagram.

BITE logic also monitors incoming signals from the receiver-converter and control servomechanism and initiates signals to the control box (IRDSC) to operate various indicator lights.

CONTROL SERVOMECHANISM ASSEMBLY

The control servomechanism assembly processes line-of-sight (LOS) position and rate commands from the IRDSC, the target tracking sight control (TTSC), or the aircraft computer, depending on the operational mode selected. These commands are processed as analog drive signals for slewing (steering) the receiver-converter drive motors and gimbals to position the receiver head. Functionally, the control servomechanism assembly breaks down into four subsystems–the power supply, azimuth drive, elevation drive, and BITE subsystems.

The power supply is a typical one. It filters aircraft power and develops all of the operating voltages for the control servomechanism assembly and the target tracking sight control assembly circuits.

Although the azimuth and elevation drive signals are processed simultaneously in a given module for simplicity, each subsystem is discussed separately. Notice the modules are labeled in their respective block diagrams as azimuth (fig. 7-39) or elevation (fig. 7-40) as appropriate. Keep in mind that, in actual practice, a module (such as mode logic) is shared by both azimuth and elevation drive signals. The following discussion refers to the module as azimuth mode logic or elevation mode logic.

Azimuth Drive Subsystem

Figure 7-39 is a simplified block diagram of a typical azimuth drive subsystem used in a control servomechanism assembly. The azimuth mode logic module receives one of four operational mode commands (position, FWD, manual track or computer track) from the FLIR control box (IRDSC), as selected by the operator.

POSITION MODE.– The position mode command signal is processed by the mode logic module, which outputs a position (POS) command signal to the azimuth position compensation module. This module processes the signal, enabling it to receive azimuth position (LOS) information from the IRDSC azimuth control. It outputs position loop command and gimbal angle signals to the azimuth rate compensation module. This module outputs an azimuth drive signal to the azimuth heat sink module. This module, in turn, amplifies the signal to develop enough motor drive power to steer the receiver head to the azimuth bearing selected on the IRDSC by the operator.

Four feedback signals are involved in ensuring the receiver head maintains the correct LOS. An azimuth tachometer signal from the receiver-converter azimuth drive unit is fed back to the mode logic module. The mode logic module produces a gimbal rate signal for the azimuth rate compensation module. An azimuth position signal from the same azimuth drive unit is fed back to the azimuth position compensation module. The mode logic module compares the azimuth position signals with the IRDSC azimuth input and, as applicable, outputs a position error signal to the azimuth rate compensation module. An azimuth gyro rate signal from the receiver-converter's gyro unit is fed to the azimuth rate compensation module. Also, an azimuth current drive signal is fed back to the rate compensation module. The rate compensation module processes all of the feedback signals (position error, gimbal rate, gyro rate, and current drive signals) and develops any necessary azimuth drive output required to maintain the receiver head at the correct selected heading.

FORWARD MODE (FWD).– The FWD mode command is processed by the azimuth mode logic module. This module outputs a FWD mode command signal to the azimuth position compensation module. The signal is processed, and the module outputs a position loop command and a gimbal angle signal to the azimuth rate compensation module. This module then outputs an azimuth drive signal to steer the receiver head to 0-degree azimuth. The azimuth drive signal is amplified by the azimuth heat sink module to produce motor power to drive the motors in the receiver-converter. The stabilization/positioning feedback circuits work the same as the circuits in the position mode previously explained.

MANUAL TRACK MODE.– The manual track mode command signal is processed by the mode logic circuit. This circuit outputs a manual track command signal and an azimuth rate inhibit signal to the azimuth rate compensation module. These signals cause the circuits to accept only azimuth rate signals from the target tracking sight control (TTSC) assembly. The TTSC assembly is a pistol grip unit. The operator uses the thumb control on the pistol grip unit to aim the receiver head to the desired LOS. The azimuth drive output signal from the circuit is controlled by the TTSC, and no feedback is used for stabilization.

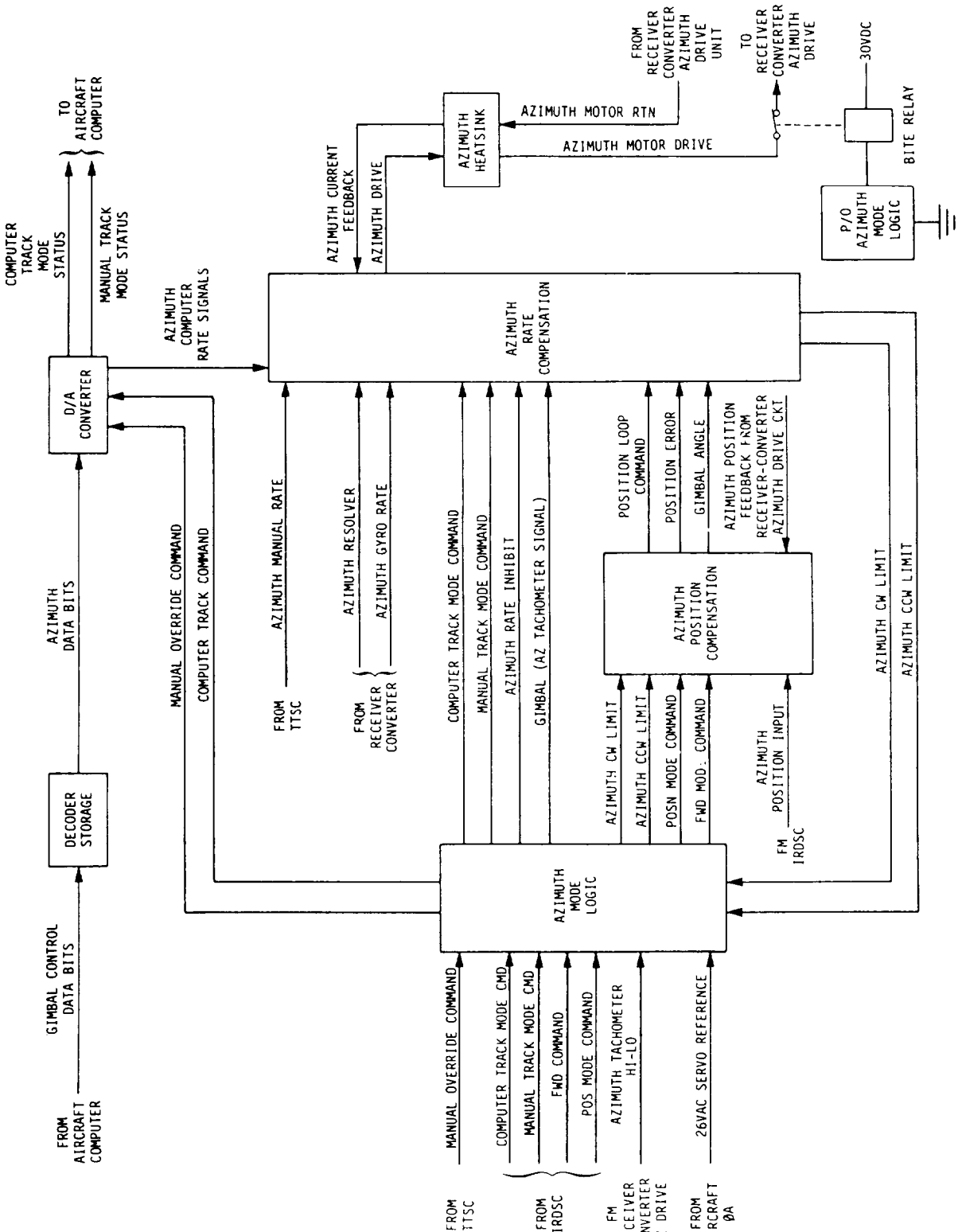


Figure 7-39.–Azimuth drive subsystem simplified block diagram.

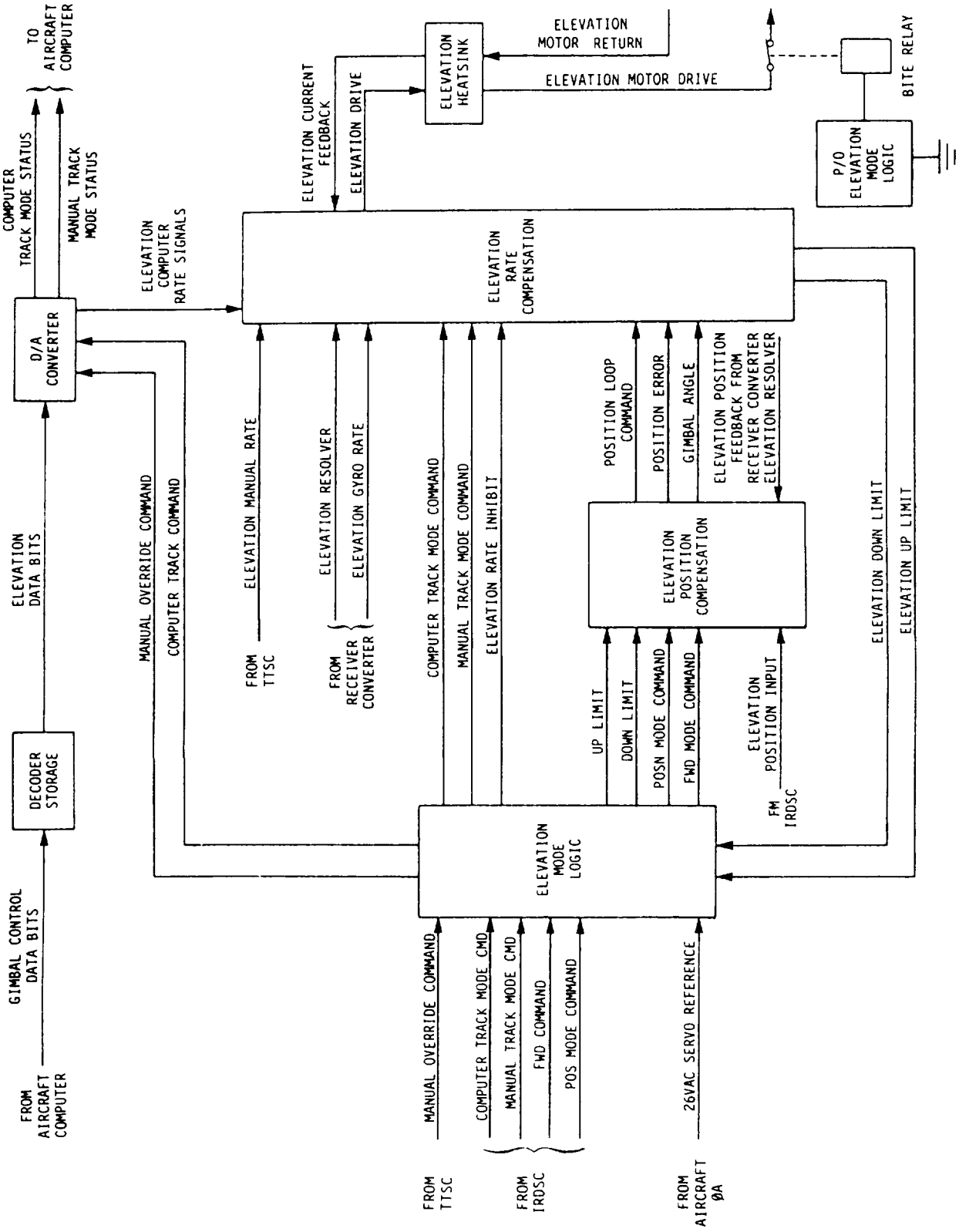


Figure 7-40.–Elevation drive subsystem simplified block diagram.

Action of a comparator circuit in the azimuth rate compensation module determines when the receiver-converter gimbals reach their electrical limits and produce limit signals (CW and CCW). These limit signals prevent the manual track mode and computer track mode commands from developing the azimuth drive signal. The limit signals are fed back to the mode logic module that outputs a CW or CCW signal to the azimuth position compensation module. This signal causes the module to develop an error signal that, in turn, develops appropriate azimuth drive.

COMPUTER TRACK MODE.– The aircraft computer supplies gimbal control data bits (azimuth and elevation position rate commands) from its program to the decoder storage module. This module demultiplexes and stores 12-bit azimuth and elevation rate commands and provides azimuth and elevation data outputs to the digital-to-analog (D/A) converter module. The purpose of storage is to allow the decoder to output data bits to the D/A converter while the computer updates itself from feedback information before issuing new gimbal-control signals to keep the receiver head at the programmed LOS.

When the computer track mode is selected on the control box, the computer track command signal is processed by the mode logic module. This module outputs a computer track mode command to the D/A converter and to the azimuth rate compensation module. The signal enables the D/A converter to process azimuth data and elevation data bits into analog azimuth position rate signals, which are fed to the azimuth rate compensation module. The D/A converter also outputs a computer track mode status signal to inform the aircraft computer that the D/A converter is operating in the computer mode. The computer track command signal enables the azimuth rate compensation module to only accept azimuth position rate signals from the D/A converter. The azimuth rate compensation module disables inputs from the azimuth position compensation module. This means no feedback information can be processed in the computer mode. The rate compensation circuit processes the azimuth position rate signal from the D/A converter and outputs an azimuth drive signal. This signal goes through the heat sink circuit to slew the receiver head to the azimuth position programmed into the computer.

An azimuth resolver in the receiver-converter feeds back four-wire resolver position information signals to the azimuth rate compensation module (not shown in the block diagram), where they are converted into three-wire syncro information signals. These signals are fed back to the aircraft computer logic circuits to update the computer and develop gimbal control data bits to maintain the receiver head at the correct LOS. If the system is operating in either the position mode, forward mode, or the computer mode (as selected on the control box) and the operator desires to quickly shift to the manual mode, the operator is able to do so by use of a manual override function on the target tracking sight control (TTSC). When manual override is initiated, a manual override command signal from the TTSC is processed by the azimuth mode logic module. This module outputs a manual override signal to the D/A converter and manual track command signals, as explained previously for manual track mode. The FLIR system functions in the manual track mode regardless of the position of the control box mode select switch.

The D/A converter is disabled by the manual override signal, and it outputs a manual track mode status signal to the aircraft computer logic to prevent the computer from operating should the control box have the computer mode selected.

Elevation Drive Subsystem

Figure 7-40 is a simplified block diagram of an elevation drive subsystem. Compare figures 7-39 and 7-40, Notice they are the same except for the elevation vice azimuth labels on the various modules. The subsystems are basically the same because the servomechanism elevation drive circuits operate the same, process the same signals, and develop the same drive signal as the azimuth drive circuits previously explained. There are only two differences–(1) no tachometer feedback is used in the elevation circuits, and (2) all signals go to or come from elevation circuits instead of azimuth circuits of other WRAs, such as IRDSC and the receiver-converter.

In the FWD mode of operation (recall that the receiver head was slewed to 0-degree azimuth), the elevation circuits slew the receive head to -4 degrees (down tilt) elevation. All other modes operate the same except the receiver head is slewed in elevation instead of in azimuth.

CS BITE Subsystem

The control servomechanism (CS) BITE subsystem automatically determines whether a servo-system failure is in the CS WRA or in the receiver-converter WRA. Figure 7-41 is a simplified block diagram of a CS BITE subsystem. For ease of signal tracing, some of the modules have been duplicated at various locations on

the diagram. These modules, such as the mode logic, have alphanumeric designators (A1, A2, A3, etc.). However, the modules are all part of one mode logic module. The alphanumeric designations are used to show where signals enter/leave a particular module.

A temperature sensor in the receiver-converter monitors the gyro operating temperature. When the gyros are operating properly, the sensor develops a "temp ready" signal, regardless of which operational mode is selected on the control box. The "temp ready" signal is fed to the mode logic module (A1) of the CS. The mode logic circuit outputs a "rcvr ready" signal to the servo BITE module (B1). The "rcvr ready" signal enables servo BITE (B1). When BITE is initiated on the control box, a "BITE initiate" signal is received by the servo BITE (B1) module from the power supply video converter BITE logic module. (See fig. 7-38.). The "BITE initiate" signal initiates a series of four tests sequenced as follows:. fault isolate, BITE 1, BITE 2, and BITE 3. All tests are controlled by a 10-Hz clock in the servo BITE module. A test sequence takes 10 to 12 seconds to complete.

FAULT ISOLATE TEST.– Initially (when "rcvr ready" and "BITE initiate" are received), the servo BITE module (B1) generates a BITE fault isolate signal and a digital computer interface (DCI) fault isolate signal. The BITE fault isolate signal goes to the following modules:. azimuth position compensation (C1), elevation position compensation (D1), mode logic (A3), azimuth rate compensation (E1), and elevation rate compensation (F1). The signal enables all of these modules and causes the azimuth position compensation module (C2) and elevation position compensation module (D2) to generate BITE motor drive signals.

The DCI fault isolate signal goes to the decoder storage module (for use in the BITE 3 test) and to mode logic (A2). The DCI signal causes mode logic (A2) to open the BITE relay drive line, de-energizing the BITE relay (shown in the de-energized position). Opening the relay removes azimuth and elevation motor drive from the motors in the receiver-converter. Instead, the azimuth and elevation heat sink motor drive output is routed to the azimuth position compensation module (C1) and elevation position compensation module (D1). The BITE motor drive signals generated by azimuth position compensation (C2) and elevation position compensation (D2) are routed by way of AZ POSN and EL POSN lines to azimuth rate compensation (E1) and elevation rate compensation (F1), respectively. The drive signals out of these modules go to the azimuth/elevation heat sink module. From this module the signals go through the BITE relay and back into azimuth position compensation module (C1) and elevation position compensation module (D1). Modules C1 and D1 output azimuth and elevation BITE position signals to servo BITE (B2). The servo BITE module (B2) outputs a servo control fail signal to the power supply video converter BITE logic circuit, which is labeled a servo malfunction in figure 7-38. This logic circuit, in turn, outputs a "not ready" signal to light the NOT READY light on the control box.

With the BITE relay de-energized (as shown), an azimuth loop is formed. This loop consists of azimuth position compensation (C2), azimuth rate compensation (El), a heat sink module and BITE relay, azimuth rate compensation (E2), and azimuth position compensation (C1). Likewise, an elevation loop is formed. This elevation loop consists of elevation position compensation (D2), elevation rate compensation (F1), a heat sink module and BITE relay, elevation rate compensation (F2), and elevation position compensation (D1). This allows the BITE motor drive signals (developed in C2 and D2) to continue around the loop. The BITE motor drive signals are monitored by frequency and amplitude detectors in servo BITE (B2). The inputs to these detectors are the azimuth and elevation BITE position signals from C1 and D1, which represent the signals in the loops. If an error or failure occurs, servo BITE (B2) generates a servo control fail signal. This signal is sent to the power supply video converter BITE logic, labeled servo malfunction in figure 7-38. From here a control servo fail signal is sent to the control box to light the CONTROL SERVO FAIL light.

If no errors or failures are present during the fault isolation test, which takes approximately 10 to 12 seconds, a BITE 1 signal is generated by servo BITE (B1), terminating the fault isolation test, and BITE 1 testing is initiated.

BITE 1 TEST.– When servo BITE (B1) generates a BITE 1 signal, the signal is fed to mode logic (A2). This causes module A2 to energize the BITE relay (via the BITE relay drive line), which reconnects the heat sink module output to the drive motors in the receiver-converter WRA and opens up the azimuth and elevation loops. Mode logic (A2) also generates a FWD command signal. This signal is fed to the azimuth position compensation module (C2) and elevation position compensation module (D2). These two modules (D2 and C2) generate BITE motor drive signals; they are fed by way of azimuth and elevation rate compensation modules (E1 and F1) to the heat sink

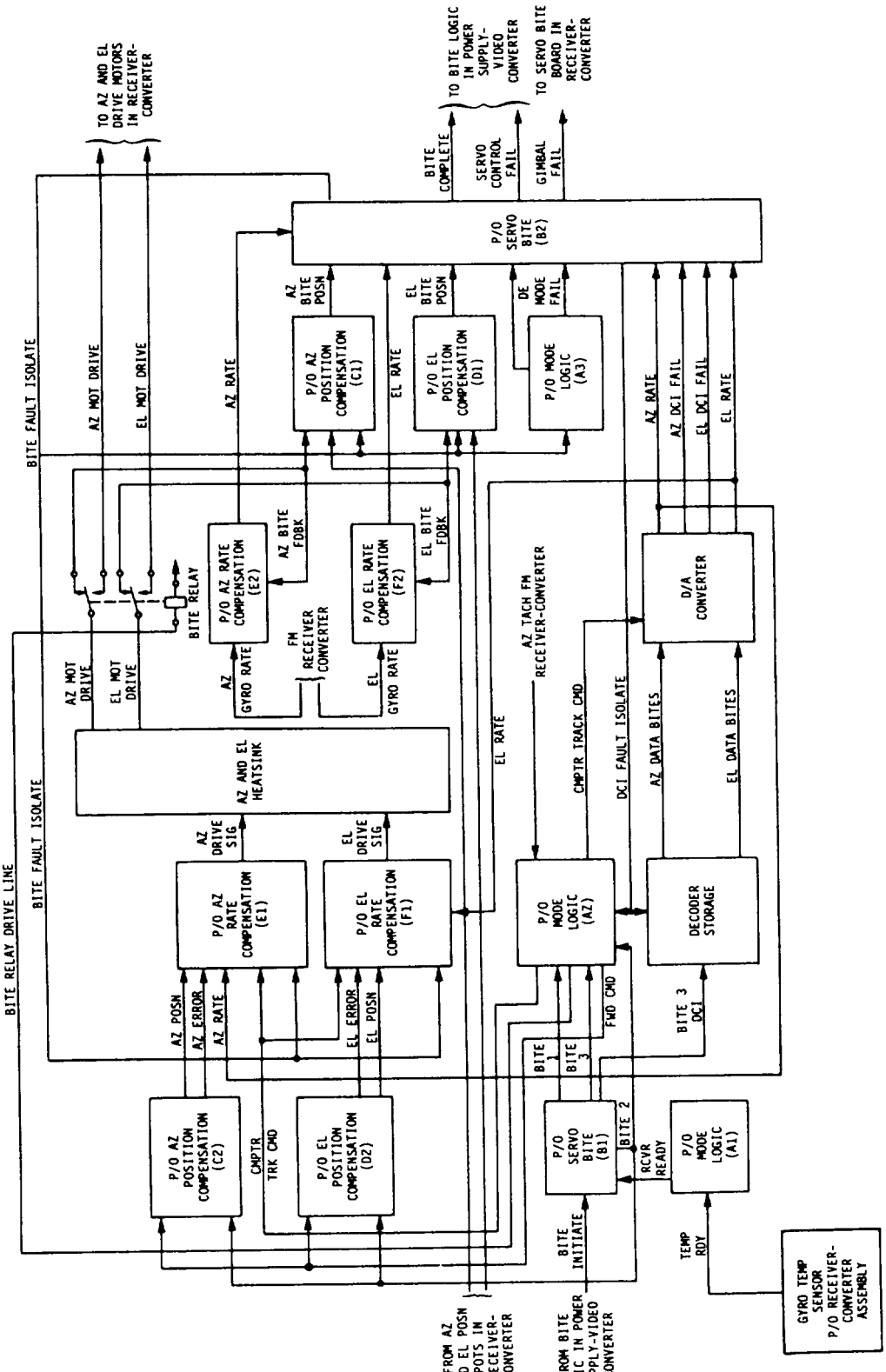


Figure 7-41.–Control servomechanism BITE simplified block diagram.

module and the now energized BITE relay to position the receiver head in the receiver-converter to 0-degree azimuth and -4 degrees of elevation. Position feedback signals from the receiver-converter are fed to the servo BITE (B2) module, where they are monitored. If there is an error/failure, a gimbal fail signal is generated and fed to the servo BITE board in the receiver-converter (fig. 7-34). This action causes the receiver BITE circuit to generate and send a receiver-converter malfunction signal to the control box to light the RCVR CONV FAIL light.

If the feedback signals to B2 are correct (for 10 to 12 seconds), a BITE 2 signal is generated by servo BITE (B1). The BITE 2 signal terminates the BITE 1 test and initiates the BITE 2 test.

BITE 2 TEST.– When servo BITE (B2) generates a BITE 2 signal, the signal is fed to the mode logic module (A2) to cancel the FWD command signal. The BITE 2 signal is also fed to the azimuth position compensation module (C2) and the elevation position compensation module (D2). The BITE 2 signal causes these modules to develop and send error motor drive signals to the receiver head by way of the same signal path as the BITE 1 test signal. These signals drive the receiver head to 130 degrees of azimuth and -60 degrees of elevation. Feedback signals are monitored by servo BITE (B2). If an error is present, B2 generates a gimbal fail signal to light the RCVR CONV FAIL light on the control box (using the same signal path as BITE 1). If the feedback signals are correct, servo BITE (B1) generates a BITE 3 signal to terminate the BITE 2 test and to initiate the BITE 3 test.

BITE 3 TEST.– When servo BITE (B 1) generates a BITE 3 signal, the module (B1) also generates a BITE 3 DCI signal (simulated computer data bit). This signal is sent to the decoder storage module. Simultaneously, a BITE 3 signal is sent to mode logic (A2). This module initiates a computer track command signal and sends it to the azimuth position compensation module (C2), the elevation position compensation module (D2), and the D/A converter module. The computer track command signal enables these modules for the computer track mode.

The BITE 3 DCI signals from the decoder storage module are processed by the D/A converter (now enabled). The D/A converter outputs azimuth and elevation rate signals. A circuit in the D/A converter monitors the amplitude and frequency of these rate signals. If the amplitude and frequency are incorrect, the D/A converter generates a DCI fail (either azimuth or

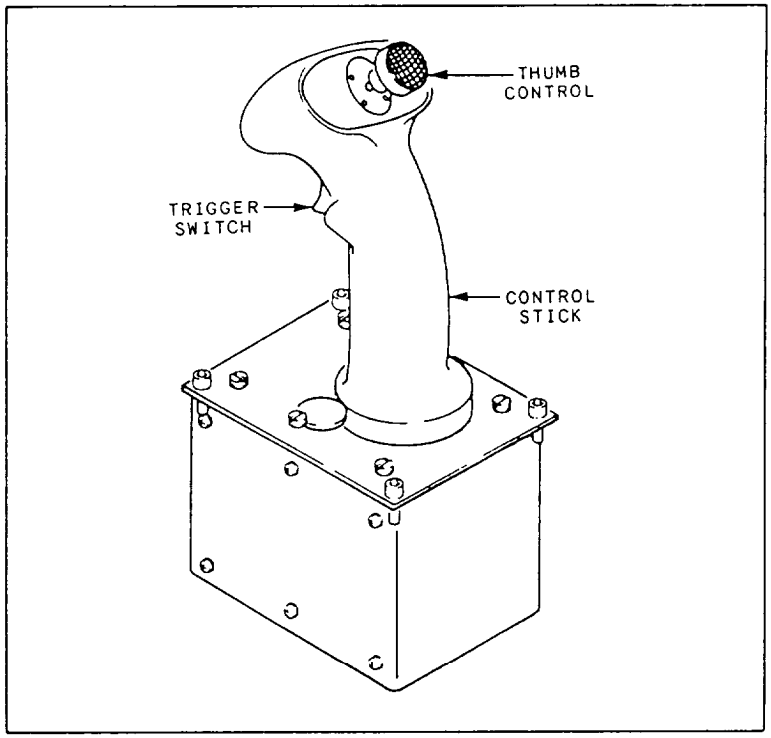


Figure 7-42.-Target tracking sight control.

elevation) signal to the servo BITE (B2). This module outputs a servo control fail signal, which, in turn, lights the CONTROL SERVO FAIL light on the control box. If the signals are correct, they are fed to the azimuth and elevation rate compensation modules (E1 and F1) to develop motor drive signals to slew the receiver head gimbals maximum CW and up. The rate feedback signals (gyro rate to E2 and F2 and azimuth tachometer to mode logic A2 and A3) are fed to servo BITE (B2). These signals are compared to the signals (azimuth/elevation rate) from the D/A converter. Should an error exist, a DCI fault isolate signal is generated by servo BITE (B2) and fed to mode logic (A2), which sends a known tachometer signal to the tachometer demodulator. If the demodulator is bad, a DEMOD fail signal is generated and sent to B2 that causes a control servo fail output. If the demodulator is good, but a rate error still exists, B2 outputs a gimbal fail signal that signifies the receiver-converter is bad. If the rate comparison shows no error, a BITE complete signal is generated by servo BITE (B2) and sent to the power supply video converter (fig. 7-38). If no error has occurred during BITE, the BITE logic module initiates a "system go" signal. This signal is sent to the control box to light the SYS GO light.

TARGET TRACKING SIGHT CONTROL

As mentioned earlier in this chapter, the target tracking sight control (TTSC) is the manual control used in the manual mode of operation to position the receiver-converter to the desired LOS. Figure 7-42 is a drawing of a FLIR TTSC. The TTSC consists of a stationary control stick (A) and the electronics for producing azimuth and elevation dc rate command signals. A thumb control (B) is used in conjunction with an angle transducer to steer the receiver head. A trigger-type switch (C) is used to provide manual override.

Figure 7-43 is a simplified block diagram of a TTSC. A voltage regulator regulates the 15-volt dc input from the control servomechanism WRA and provides +6 volts of dc and -6 volts of dc to an angle transducer. Adjustment of the thumb control produces a voltage output of the elevation and azimuth angle transducers. These outputs are amplified and sent to the control servomechanism, where they are processed (as previously explained) to position the receiver head to the desired LOS.

Should the computer tracking mode (forward or position mode of operation) be selected on the control box, depressing the trigger switch initiates a manual override command signal for the mode logic and D/A

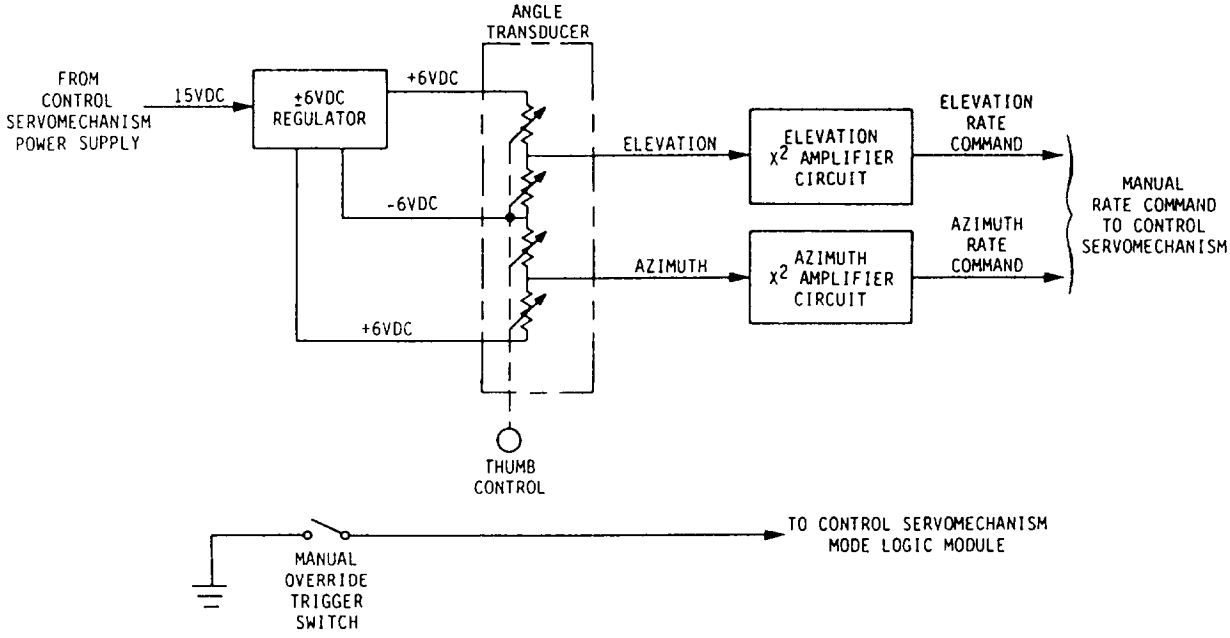


Figure 7-43.-TTSC simplified block diagram.

converter modules in the control servomechanism. This places the system in the manual mode of operation.

INFRARED DETECTING SET CONTROL (IRDSC)

Figure 7-44 is a drawing of an IRDSC/FLIR control box. Each control has been numbered, and the corresponding number is listed in table 7-4. Refer to table 7-4 for an explanation of the name, description, and function of each control and indicator.

VIDEO INDICATOR

The typical video indicator (fig. 7-36) is an 875-line, 30 frames-per-second, closed-circuit TV monitor. On the front of the WRA is an OFF-ON power switch, an elapsed time meter, brightness and contrast controls, and a status indicator. Operation of the indicator is similar to that of the TV monitors. Figure 7-45 is a simplified block diagram of a video indicator's signal processing circuits.

Composite video from the video line driver module of the power supply video converter (fig. 7-35) is applied to the video amplifier/sync stripper module. This module separates the video signals (IR, RETICLE, GAI, and grayscale) from the sync signals (blanking, clamping, and sync). The module amplifies the video signals and provides the video output to the CRT for display. The contrast control is also injected in the video amplifier/sync stripper module. The module also sends the composite sync signals to the vertical, sync, CRT protect, and brightness control module.

The video amplifier/sync stripper module processes the composite sync signal; it provides vertical and horizontal sync signals to the vertical and horizontal sweep module; and it provides blanking and clamping signals to the CRT. The brightness control is injected in this module. The module also monitors the vertical sweep (yoke) and horizontal flyback signals, removes the 300-volt dc operating voltage from the CRT, and extinguishes the STATUS light when a failure occurs.

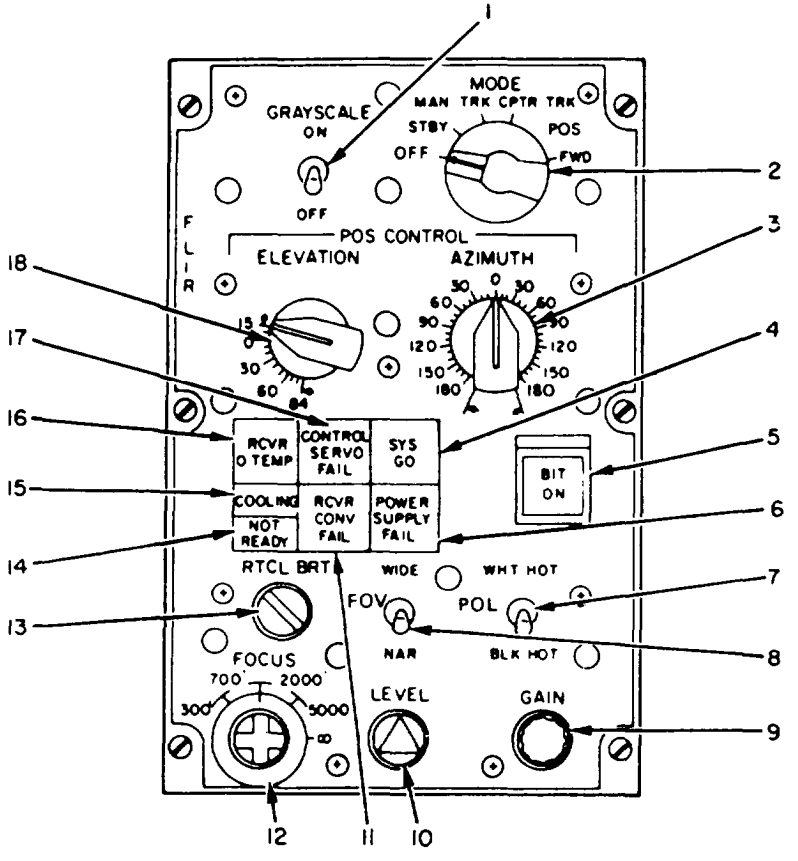


Figure 7-44.-IRDSC/FLIR control box.

The vertical and horizontal sweep module generates the vertical and horizontal sweeps used to drive the CRT yoke; it provides these drive outputs to the sweep heat sink module. The sweep module receives feedback signals from the sweep heat sink module to update and maintain the proper drive outputs.

The sweep heat sink module receives the vertical and horizontal drive signals and provides the proper level of yoke drive to the CRT. The sweep heat sink module also sends feedback signals to the vertical and horizontal sweep module and to the monitor circuits in the vertical, sync, CRT protect, and brightness control module.

The image presented on the video indicator is a TV picture of the IR energy scanned by the receiver. The video indicator is independent of the BITE subsystems of the other WRAs. Only the STATUS light and picture are indicative of a properly operating video indicator.

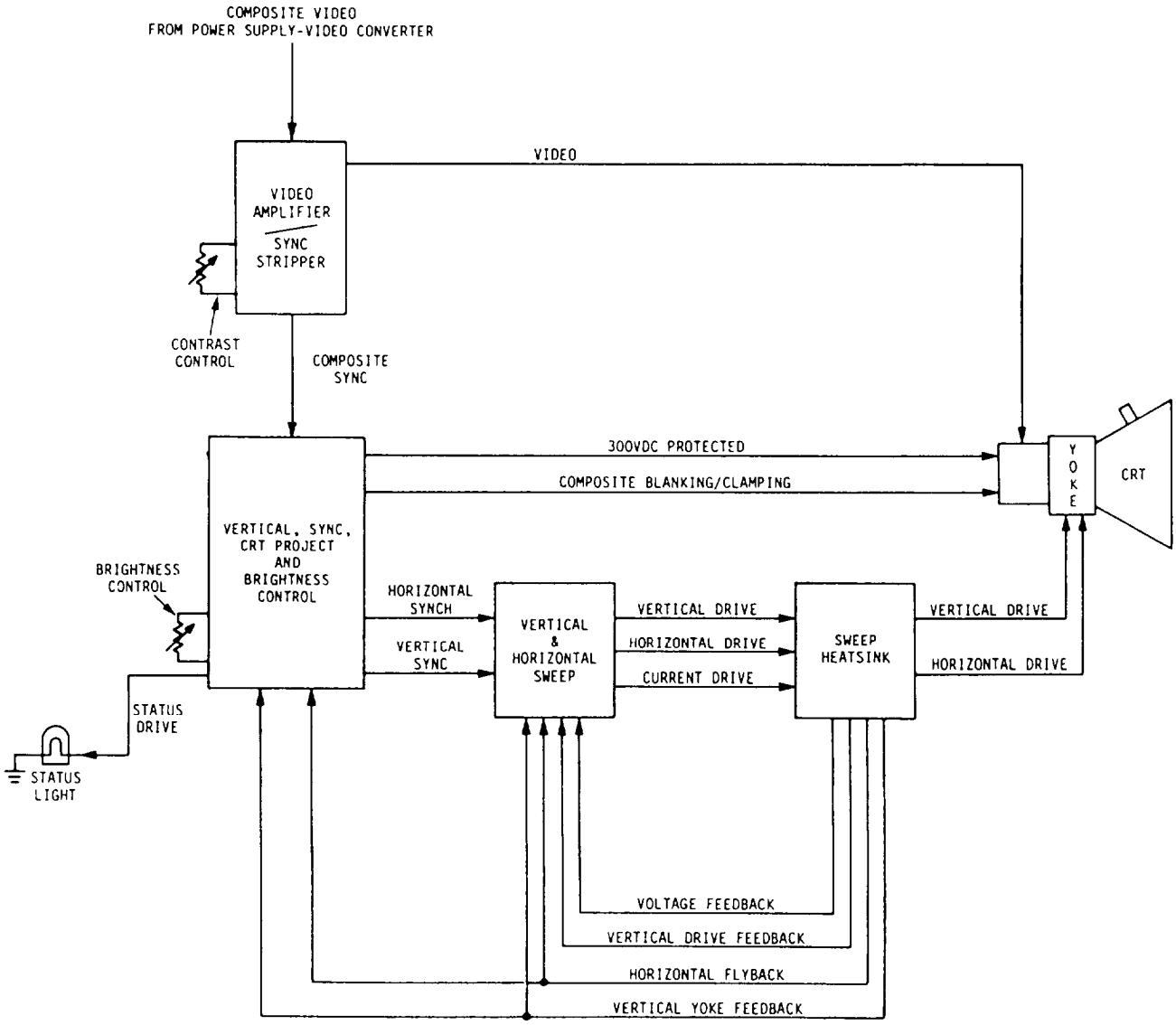


Figure 7-45.-Video indicator simplified block diagram.

Table 7-4.-IRDSC/FLIR Control Box Controls and Their Functions

1	GRAYSCALE	2- position toggle switch	Energizes circuit in power supply-video converter WRA to present eleven shades of gray across the bottom of the video indicator
2	MODE Switch	6-position rotary switch	
	STBY		Establishes and maintains system operational readiness by activating air conditioning and cryogenic cooling of detectors. Positions the receiver head in the stow position (CCW and up limits)
	MANTRK		Establishes manual track mode of operation. System responds to slew signals from TTSC only
	*CPTRTRK		Establishes computer track mode of operation. System responds to slew signals from aircraft computer
	*POS		Establishes position mode of operation. System responds to slew signals from controls 3 and 18
	*FWD		Establishes forward mode of operation. Initiates slew signals to position receiver head to 0° azimuth and –4° elevation with respect to aircraft flight path
3	AZIMUTH	Potentiometer	Controls receiver slew signals (azimuth) in the position mode of operation
4	SYST GO	Back-lighted momentary action pushbutton switch	Illuminates to indicate a good system at completion of BITE. Can be extinguished by pressing the indicator
5	BIT ON	Back-lighted momentary action pushbutton switch	Initiates BITE test sequence. Light is on while BITE is in progress. Extinguishes upon completion of BITE
6	POWER SUPPLY FAIL	Indicator fails BITE	Illuminates if power supply-video converter
7	POL	2-position toggle switch	Selects polarity of video signal from postamplifiers of the receivers converter detector array video amplifier circuit
	WHT HOT		In WHT HOT position, hot targets appear white on the video indicator
	BLK HOT		In BLK HOT position, hot targets appear black on the video indicator

*NOTE:	CPTRTRK, POS, FWD modes can be overridden by use of the override trigger switch on the TTSC. When the switch is depressed, the system is placed in manual track mode of operation and stays in this mode for as long as the switch is depressed

Table 7-4.-IRDSC/FLIR Control Box Controls and Their Function–continued

8	FOV	2-position toggle switch	Selects wide or narrow field of view by switching a focal lenses assembly in or out of optical path of receiver
9	GAIN	Potentiometer	Adjust dc level of video output to video indicator and affects contrast of targets on indicator
10	LEVEL	Potentiometer	Adjusts dc level of video output of the receiver-converter and affects brightness of background on the video indicator
11	RCVR CONV FAIL	Indicator	Illuminates if receiver-converter fails BITE test
12	FOCUS	4 position rotary switch	Selects target range for focusing in the narrow FOV (focuses a focal lenses)
13	RTCL BRT	Potentiometer	Controls brightness of reticle superimposed on video signal applied to video indicator
14	NOT READY	Indicator	Illuminates when receiver-converter is not at operating temperature, when gyro spin up is not complete, or when bite is in the fault-isolate mode
15	COOLING	Indicator	Illuminates to indicate IR detectors in receiver-converter WRA have not reached operating temperature
16	RCVR O TEMP	Indicator	Indicates excessive temperature within the receiver-converter WRA
17	SERVO FAIL	Indicator	Illuminates if control servomechanism fails BITE test
18	ELEVATION	Potentiometer	Controls receiver slew signals (elevation) in the position mode of operation

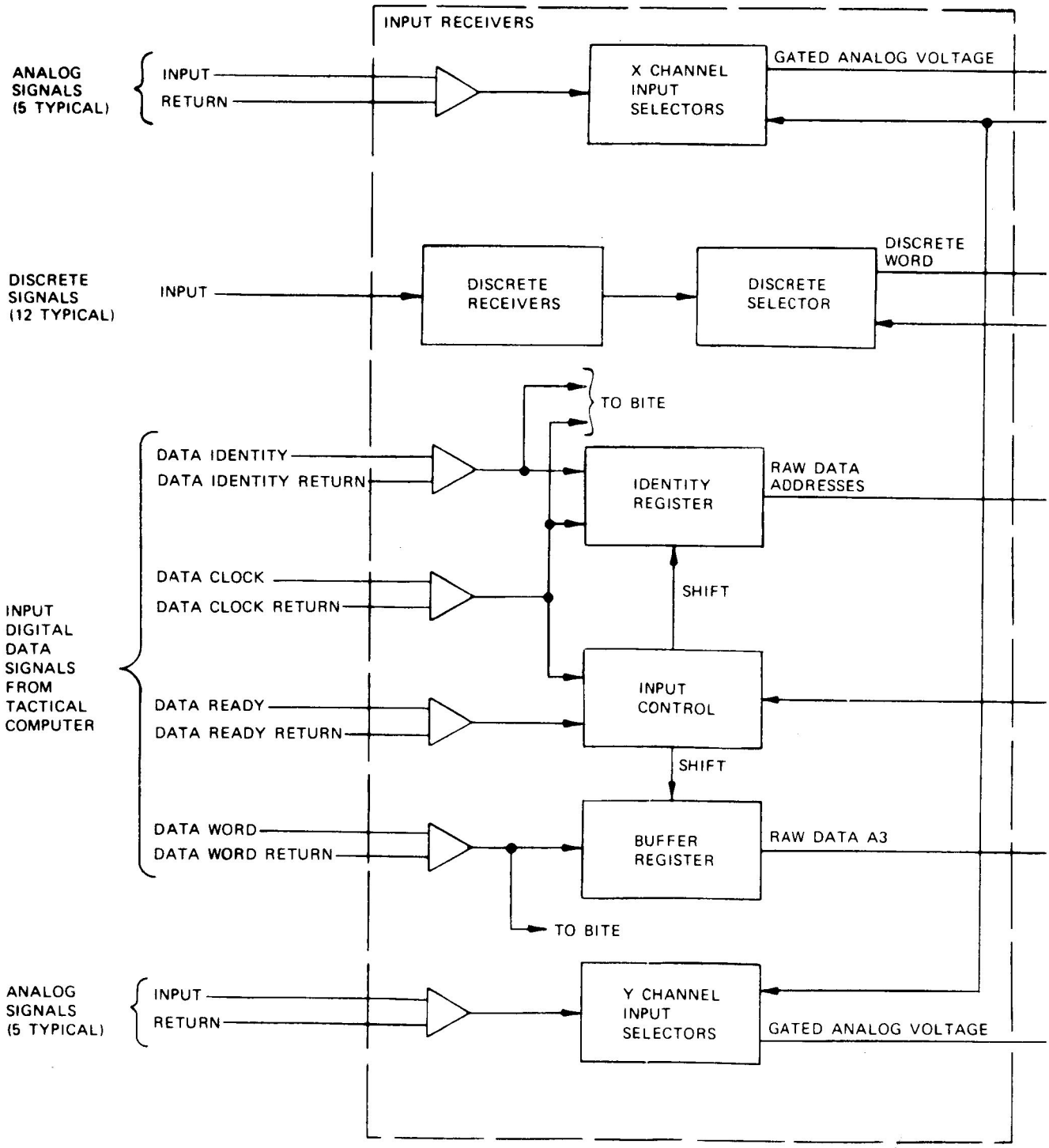


Figure 7-5a.—Signal data processor signal flow diagram.

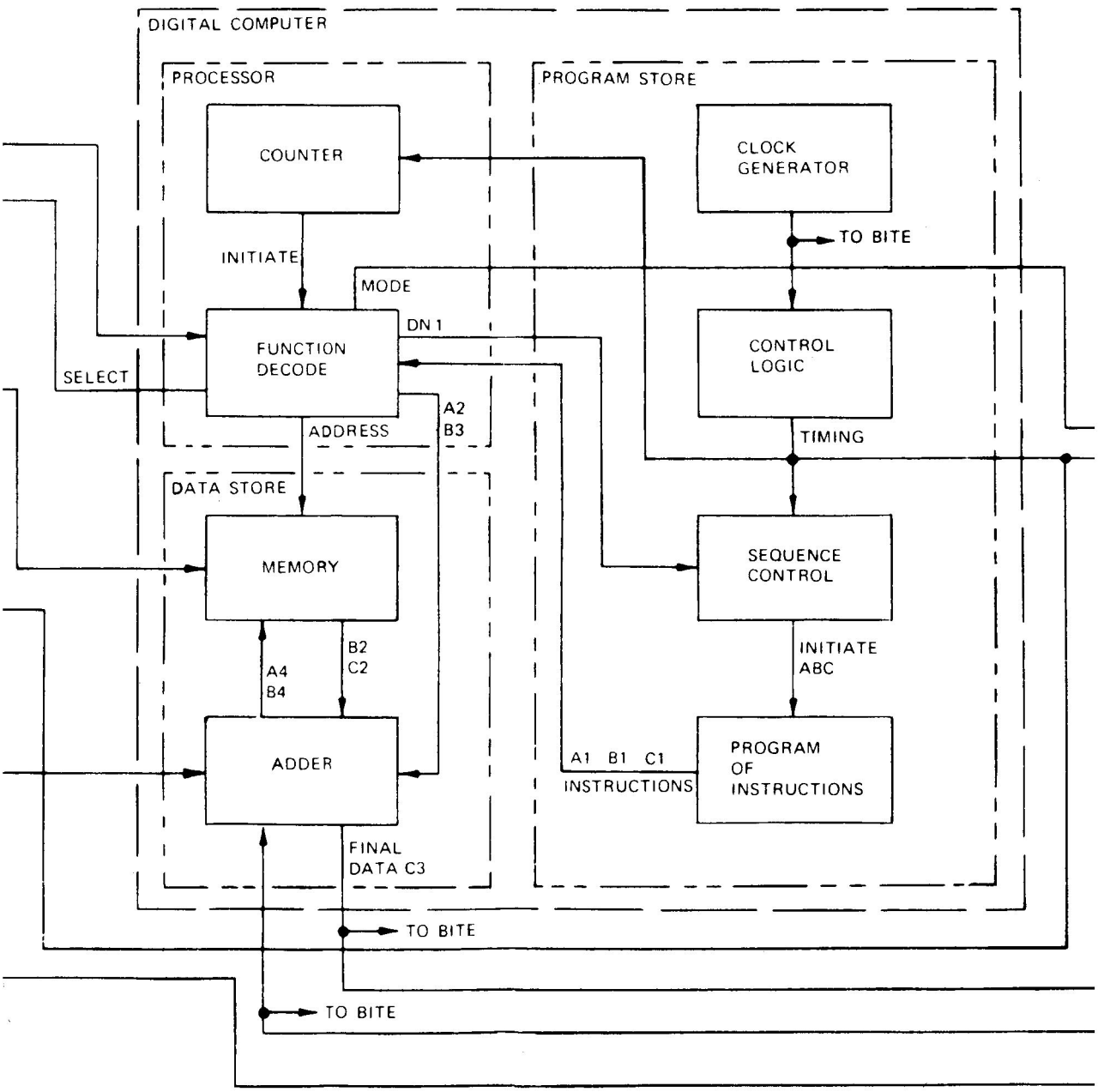


Figure 7-5b.—Signal data processor signal flow diagram. (Cont'd)

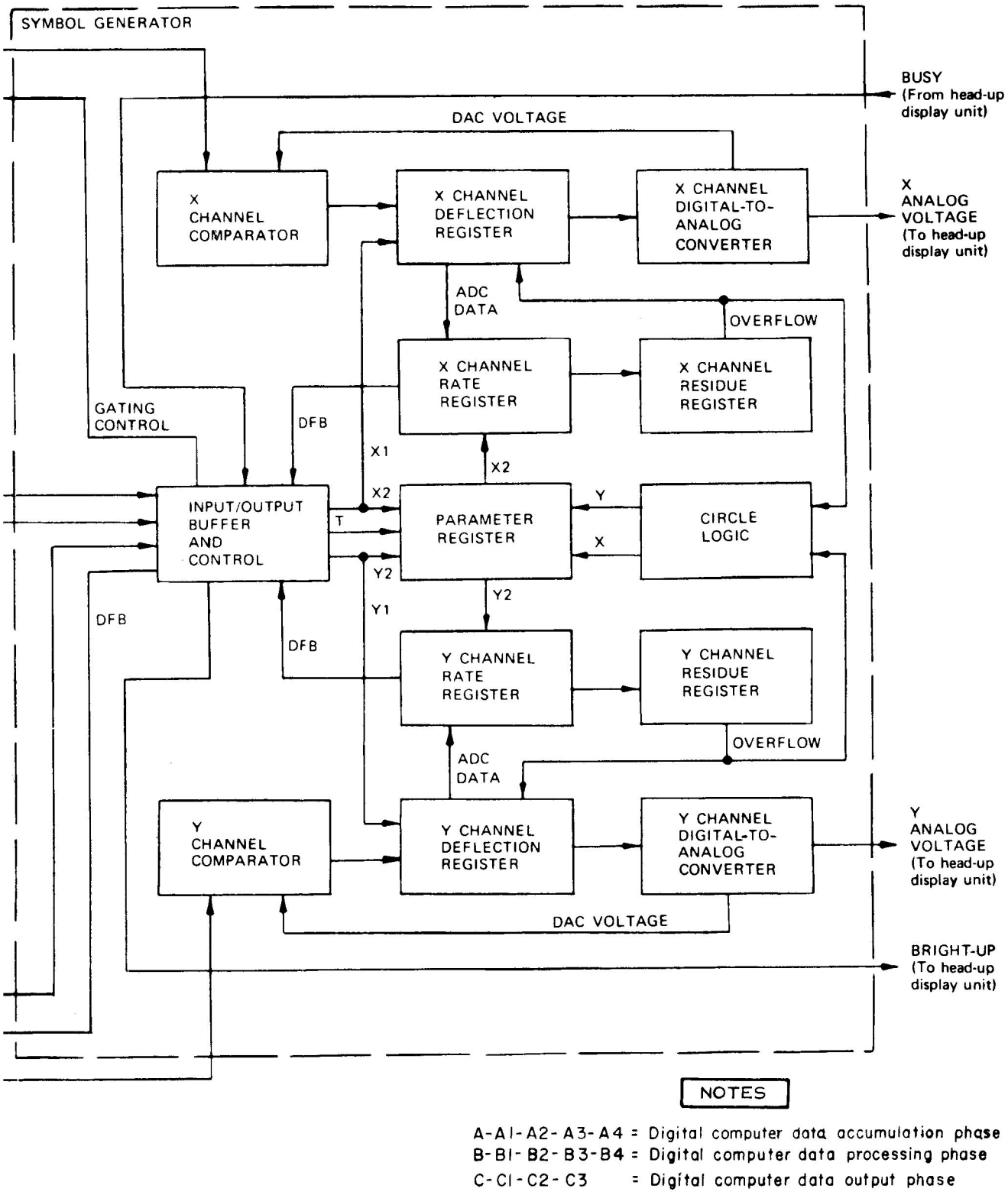


Figure 7-5c.—Signal data processor signal flow diagram. (Cont'd)

CHAPTER 8

TELEVISION

Chapter Objective: *Identify the fundamental operating principles of television to include camera tubes, camera circuits, composite video, the control unit, receivers, sound systems, picture tubes, co for circuits, and power supplies.*

Television (TV) is defined as the transmission and reception of visual images by means of electrical signals from one point to another. All TV contains the following basic elements:

1. A pickup device (transducer) to convert an optical image into an electrical signal.

2. A means of transmitting the electrical signal.

3. A reproducing device to reconvert the electrical signal into an optical image.

Pickup devices, or camera tubes, include image orthicons, image isocons, vidicons, and secondary electron conduction (SEC) tubes. Transmitting devices include coaxial lines, low- or high-power transmitters, or microwave relay links. The reproducing device is a TV receiver, a TV monitor, or some form of electrooptical projection system.

The first large-scale application of TV was in commercial broadcasting. Technical requirements imposed on terminal equipment for commercial broadcasting are standardized by the Electronics Industries Association (EIA) (formerly the Radio-Electronics Television Manufacturers Association [RETMA]). Standardization is necessary because the objective of each TV station is the same–to produce a program suitable for transmission to receivers.

In the closed-circuit television (CCTV) industry, technical requirements vary widely from one application to another. Technical standards for closed-circuit TV do not necessary have to conform to specifications governing commercial broadcasting. Requirements for picture quality may be much lower or higher than for broadcast TV. For example, when you need to read a meter at a distance, transmission standards may be lower than when the transmission involves highly detailed map information.

Another difference between commercial and closed-circuit TV is the greater variety of equipment and facilities used in closed-circuit TV. Installations vary from a simple camera-monitor combination to highly complex combinations of cameras, distribution and switching equipment, and viewing monitors. Also, closed-circuit TV equipment must operate under a wide variety of environmental conditions that cannot always be controlled.

As an Aviation Electronics Technician (AT), you must be familiar with TV systems used in several different applications. These include special weapons guidance, weapon prelaunch information, and the use of TV as an attack display in many modern aircraft.

TELEVISION FUNDAMENTALS

Learning Objective: *Identify the fundamentals of TV to include the basic systems, scanning signals, and other systems.*

TV transmission of a picture from one point to another involves a process in which light, reflected from an objector scene, is converted into electrical impulses of varying magnitude. This process is called synchronized scanning. In synchronized scanning, the picture is examined by the camera and reproduced by the viewing monitor. This is done point-by-point and in regular pattern. The process is carried out so rapidly that the entire picture is scanned many times each second, and the eye sees it as a single complete image.

THE BASIC TELEVISION SYSTEM

The basic TV system is shown in figure 8-1. It consists of a transmitter and a monitor. An image of the scene is focused on the camera pickup device. An electron beam in the device scans the optical image and produces an electrical signal which varies in amplitude with the amount of light that falls on each point of the image. A synchronizing signal is added to the electrical picture signal from the camera. The resultant composite video signal is transmitted to the display monitor.

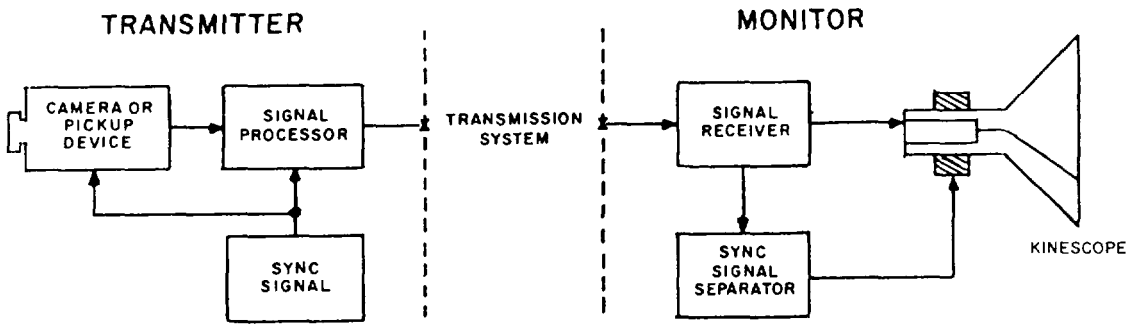


Figure 8-1.-Basic television system.

Within the monitor, the synchronizing signal causes the beam to scan the kinescope (TV picture tube) faceplate in sync with the camera scanning beam. The intensity of the kinescope beam is varied with the picture signal, and the image appears on the face of the kinescope.

The time required for one vertical scan of the picture in broadcast TV systems in the United States is 1/60 of a second (60 hertz) or a multiple or submultiple thereof. The rate of 60 hertz per second was chosen because most commercial electrical power sources in the United States operate at a frequency of 60 Hz. Synchronization with the power frequency reduces the visible effects of hum and simplifies the problem of synchronizing film projectors with scanning.

SCANNING

The number of scanning lines determines the maximum ability of the system to resolve fine detail in the vertical direction. Also, the number of scanning lines is related to the resolution ability in the horizontal direction. Resolution is determined by the number of scanning lines because, for a given video bandwidth and frame time, horizontal resolution is inversely proportional to the number of scanning lines. Therefore, as the number of scanning lines are increased, the bandwidth of the system must also be increased in the same ratio to maintain the same resolution in the horizontal direction.

Maintaining approximately equal values of horizontal and vertical resolution is ideal. The bandwidth requirements increase as the square of the number of scanning lines. The present system of 525 lines was chosen for broadcast TV as the most suitable compromise between channel width and picture resolution. This number of lines has also been found satisfactory for, and used in, many closed-circuit TV systems.

Noninterlaced Scanning

The simplest scanning method is called noninterlaced or sequential scanning (sometimes called "progressive"). This scanning method uses an electron beam that moves very rapidly from left to right on an essentially horizontal line, while it travels slowly from the top to the bottom of the picture. When the electron beam reaches the end of a line, a blanking voltage is applied that shuts off the beam. This period of time is known as the horizontal retrace period or "flyback" time. Similarly, when the beam reaches the bottom of the picture, the beam is blanked out and reappears at the top of the picture.

Interlaced Scanning

An important variation of the scanning method discussed above is called interlaced scanning. It is used in broadcast TV and in most closed-circuit TV equipment. With interlaced scanning, it is possible to reduce the video bandwidth by a factor of two without reducing resolution or seriously increasing flicker.

In the standard two-to-one method of interlacing, alternate lines are scanned consecutively from top to bottom. Then, the remaining alternate lines are scanned. This principle is illustrated in figure 8-2, which shows interlaced scanning with 13 scanning lines. In this kind of scanning, each of the two groups of alternate lines is called a "field." The complete set of lines, consisting of two consecutive fields, is called a "frame." Interlacing is accomplished by making the total number of lines in a frame an odd integer. Thus, the number of lines in each of the two fields is an even number plus one-half line. This results in consecutive fields that are displaced in space with respect to each other by one-half of a line. Thus, interlacing of the lines is produced. In the actual method used for broadcast TV and most closed-circuit TV, the total number of lines is 525, the total per field is

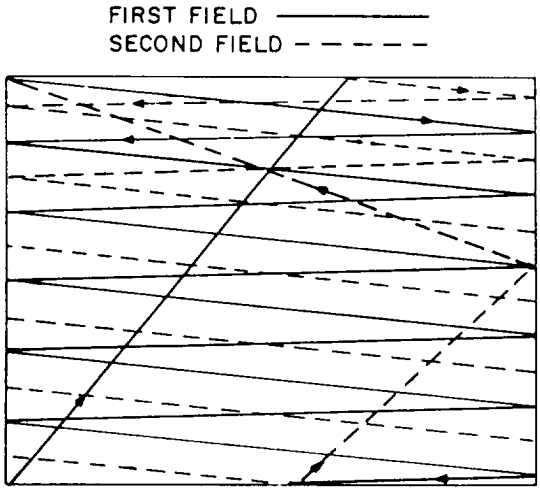


Figure 8-2.-Interlaced scanning.

262 1/2, the vertical scanning frequency is 60 Hz, the number of complete pictures (frames) per second is 30, and the horizontal scanning frequency is 15,750 Hz (60 x 262 1/2).

TELEVISION SIGNALS

The standard TV signal consists of four elements:

1. The picture information generated during active scanning time
2. The picture blanking pulses
3. The picture average dc component
4. The picture synchronizing pulses

Picture Information

The basic part of the signal, the picture information, is a series of waves and pulses generated during active line scanning of the pickup or camera tube. A scanning line traverses (travels back and forth) the face of the pickup tube. It is modulated in amplitude in proportion to the brightness variations in the scene it is scanning. The signal produced varies in amplitude proportionally with the brightness of the scene. For commercial broadcasting, the amplitude variations are such that the maximum video amplitude produces black and the minimum video amplitude produces white. Ordinarily the maximum and minimum video amplitude values represent 75 percent and 15 percent of the maximum carrier voltage, respectively.

Picture Blanking Pulses

To prevent undesirable signals from entering the picture during retrace time, blanking pulses are applied to the scanning beams in both the camera tube and the receiver kinescope. Camera blanking pulses are used only in the pickup device. They serve only to close the scanning aperture on the camera tube during retrace periods and never actually appear in the final signal sent to the receiver. In some systems, the same pulse that triggers the scanning circuit and blanks the kinescope also closes the camera scanning aperture.

The function of the kinescope blanking pulses is to suppress the scanning beam in the kinescope during both vertical and horizontal retrace time. They are simple rectangular pulses, somewhat wider than the corresponding camera blanking pulses. They have a duration slightly longer than the actual retrace periods in order to trim up the edges of the picture and provide a clean noise-free period during retrace. The complete video signal shown in figure 8-3 contains pulses for the removal of visible lines during horizontal retrace periods only. The horizontal pulses recur at intervals of

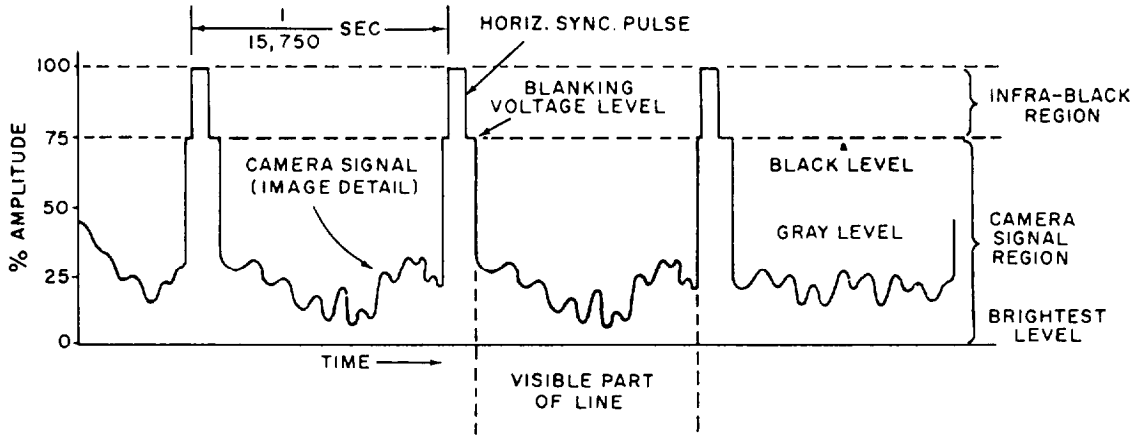


Figure 8-3.-The complete video signal for three scanned lines.

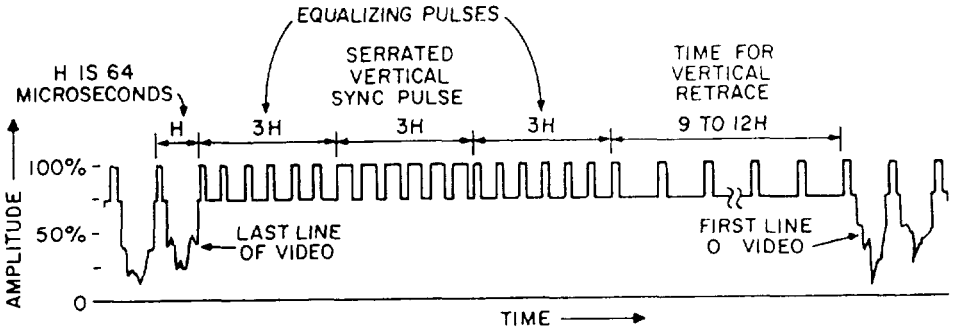


Figure 8-4.-Vertical synchronizing and equalizing pulses.

1/15,750 of a second. At the bottom of the picture they are replaced by vertical blanking pulses, which are similar to the horizontal pulses, except they are of much longer duration (approximately 15 scanning lines) and have a periodic recurrence of 1/60 of a second.

The blanking pulses (and synchronizing pulses) are added at a relatively high-level point in the transmitter and are, therefore, considered to be noise-free. The importance of noise-free blanking and synchronizing pulses should not be underestimated. They determine the stability of the viewed picture or the degree to which a picture will remain locked-in on a kinescope even under the most adverse transmission conditions. This point is especially important when considering the use of TV for closed-circuit applications. Extreme environmental conditions can seriously degrade the picture signal, making it difficult to synchronize or lock-in a picture unless the original blanking-to-picture and signal-to-noise ratios are high.

Picture Average dc Component

If a TV picture is to be transmitted successfully with the necessary fidelity, it needs the dc component of the picture signal. This component is a result of slow changes in light intensity. The loss of the dc component occurs in ac or capacitive coupling circuits, and is evidenced by the picture signal tending to adjust itself about its own ac axis. The dc component is returned to the video signal by a dc restorer or inserter circuit.

Synchronizing

Synchronization of the scanning beams in the camera and the receiver must be exact at all times to provide a viewable picture. To accomplish this, synchronizing information is provided in the form of electrical pulses in the retrace intervals between successive lines and between successive pictures (fig. 8-3). The retrace periods (which are as short as circuit considerations permit) are in areas in which synchronization pulses may be inserted without interfering with the picture.

Synchronizing pulses are generated at the program origin end of the TV system in the equipment that controls the timing of the scanning beam in the pickup tube. These pulses become a part of the complete signal transmitted to the receiver or monitor. In this manner, scanning operations at both ends of the TV system are always instep with each other. In general, synchronizing signals should provide positive synchronization of both horizontal and vertical sweep circuits. The signals should be separable by simple electronic circuits to recover the vertical and horizontal components of the composite sync signal. They should also be able to be combined simply with the picture and blanking signals to produce a standard composite TV signal.

Most TV systems produce synchronizing information that conforms to the basic requirements of synchronization. Figure 8-3 shows how the synchronizing signal waveform is added to the picture information and blanking signals to form a complete composite picture signal ready to be transmitted. Note that the duration of the horizontal sync pulses is considerably shorter than that of the blanking pulses. Vertical sync pulses are rectangular, but they are of much shorter duration than the horizontal pulses. Thus, vertical sync pulses provide the necessary means for frequency discrimination.

Synchronization presents a difficult problem because the largest number of failures occurs as a result of the loss of proper interlacing. Discrepancies in either timing or amplitude of the vertical scanning of alternate fields cause displacement in space of the interlaced fields. The result is nonuniform spacing of the scanning lines. This reduces the vertical resolution and makes the line structure of the picture visible at normal viewing distance. The effect is usually called “pairing.”

To prevent the pairing problem and maintain continuous horizontal synchronizing information throughout the vertical synchronization and blanking interval, another series of pulses is added before and after the vertical sync pulses. These are "equalizing" pulses (fig. 8-4). The time between the last horizontal sync pulse and the first equalizing pulse changes from a full horizontal line interval to one-half of a horizontal line interval every other field. This change is caused by the ratio between 15,750 Hz and 60 Hz. The ratio produces the necessary difference between fields to provide interlaced scanning. Since the horizontal oscillator is adjusted to the frequency of the horizontal sync pulses, it is triggered only by every other equalizing pulse or serration of the vertical sync pulse.

OTHER SYSTEMS

Although commercial broadcasting and many closed-circuit installations adhere closely to the previously mentioned EIA standards, some noncommercial as well as closed-circuit installations use synchronizing signal specifications that are considerably less rigorous. The TV systems that are discussed fall into the following four general categories: (1) random interlace, (2) odd-line interlace, no special sync pulses, (3) odd-line interlace, modified sync pulses, and (4) slow-speed scan.

Random Interlace, No Special Sync Pulses

The random interlace is the simplest TV method. It provides no special sync pulses and no fixed relationship between the horizontal and vertical scanning raster. "Lock-in" or synchronization information at the receiver or monitor is obtained from the horizontal and vertical blanking pulses contained in the video signal, as shown in figure 8-5, view A. Usually at the camera control location, sufficient blanking signal or "setup" is added to the video signal to provide an adequately long and steep transition at both vertical and horizontal frequencies to provide lock-in.

Some receivers and monitors may have difficulty, however, in synchronizing with this information. Electrical noise is a possible condition when the camera and viewing device are separated by a great distance. The lock-in of such a signal becomes extremely difficult. Note in figure 8-5, view A, that there is no horizontal sync signal during vertical blanking time. The horizontal frequency circuit in the monitor, even though it has some tendency to keep on the correct frequency, will be essentially free-running during this period. This

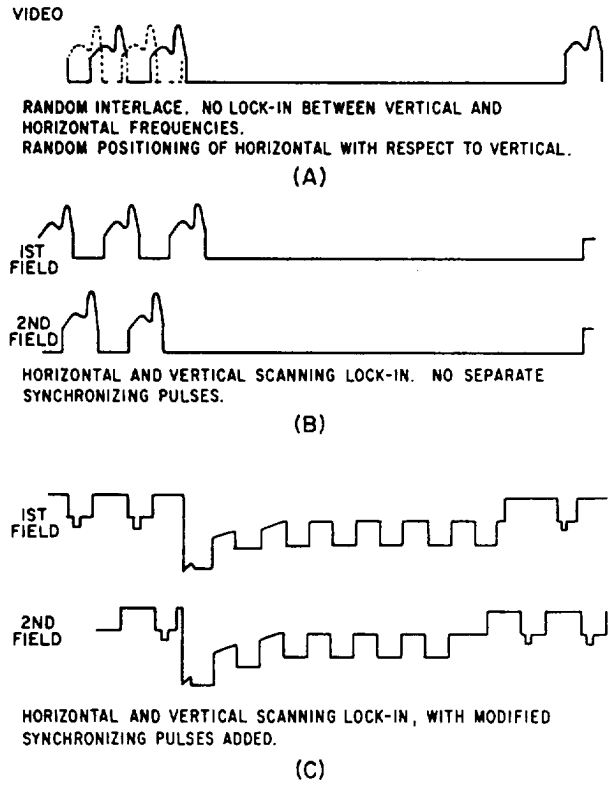


Figure 8-5.-Nonstandardized television waveforms (positive picture phase).

may not have a harmful effect during blanking or retrace time. However, when the horizontal sync information returns, the receiver horizontal circuits may have trouble synchronizing to the new information.

The most undesirable characteristic of this TV system is insufficient resolution caused by lack of interlace. Good interlace is not possible because an absolute frequency relationship between the horizontal and vertical frequencies is lacking. The nominal vertical frequency is usually 60 Hz (lock-in to the 60-Hz power line), while the horizontal frequency (usually established by a free-running oscillator) is nominally 15.75 kHz. Thus, there is no direct relationship between the two frequencies (horizontal frequency should bean odd multiple of one-half the field rate) as required for satisfactory interlace. The advantages of such a system are reduced cost and greater simplicity of circuits. However, marginal resolution capabilities, incompatibility, marginal stability, and general reduction of system performance limit its application and use.

Odd-Line Interlace, No Special Sync Pulses

The odd-line interlace system has a distinct advantage over the previous system. A definite

relationship exists between the horizontal scan frequency and the vertical field rate. This system, whose waveform is shown in figure 8-5, view B, can effectively use the 2:1 odd-line interlace technique. Therefore, it provides a considerable improvement in the resolution capabilities of the system. In theory, the vertical resolution should be double that of the previous system. In practice, however, the improvement in resolution is somewhat less since it is difficult to obtain perfect interlace (no pairing).

Like the random interlace, this system does not provide a special synchronizing signal, and it is subject to the same synchronizing or lock-in limitations as were discussed previously. These limitations become an important problem in the more elaborate installations where a series of cameras and monitors separated by wide distances might be used. For this reason, installations using these systems are usually limited to smaller, less complex applications where stability and reliability may not be an important factor.

Odd-Line Interlace, Modified Sync Pulses

The odd-line interlace method provides further advantages over the previous two systems, but it has a considerable number of limitations when compared with the EIA system. In this system, a form of special synchronizing signal has been added to the video waveform (fig. 8-5, view C). Note that the synchronizing signal has been added to the tip of each horizontal blanking pulse. The sync pulses also continue through the vertical blanking interval. They provide synchronizing information for the monitor horizontal frequency-locking circuits at all times. These circuits are therefore no longer free-running during the vertical blanking interval. Addition of the special sync pulse information greatly improves the lock-in ability of the composite video signal under adverse conditions of noise and false signals.

Slow-Speed Scan

A TV system that is being used more frequently uses the slow-speed scan technique. This technique represents a radical departure from nominal scanning standards. It permits a scene, which contains a limited amount of action or movement and a great deal of redundancy, to be picked up and transmitted successfully from one location to another. It affords fair resolution and fidelity in signals transmitted over relatively economical narrow-band transmission facilities. For example, a “slow-speed” camera located in a bank a message center, or a newspaper office can scan printed information and transmit it to a distant location over ordinary telephone line facilities. Some methods are able to transmit pictures having more action, such as a person talking, with reasonable clarity. Such methods, however, require somewhat greater bandwidth.

Most slow-speed scan systems use a much slower scanning rate, with a correspondingly narrower bandwidth, than present telecasting standards. Broadcasting systems transmit a picture every 1/30 of a second, with a 4-MHz bandwidth. Slow-scan systems transmit a picture in 1/10 of a second to 2 seconds, with a video bandwidth ranging from approximately 250 kHz to as low as 500 Hz.

Slow-speed scan systems are practical where time is available for transmission. For example, the information contained in a 5-minute commercial TV program requires several hours of time to be transmitted with comparable detail by the average slow-speed scanning technique. The advantages of the slow-speed scanning system are greatly simplified equipment and relatively inexpensive transmission facilities, as compared to the complex relay systems required for broadcast TV. The disadvantages are that the scene content is limited to relatively immobile objects, resolution is marginal, and the system is incompatible with standard TV systems. Complex scan conversion equipment is required to make the two systems compatible. Except for certain special applications, slow-speed scan systems are inferior in performance and cannot be used successfully where a high degree of resolution and detail is required.

CAMERA TUBES

Learning Objective: *Identify the basic types of camera tubes to include vidicon, plumbicon, image orthicon, and secondary electron conduction.*

The type of camera tube used is determined by the intended use of the camera and the amount of available illumination. The amount of light required by a camera tube is rated in candelas. The minimum number of candelas required by a camera tube is a measure of the tube's sensitivity. The following types of camera tubes are used in modem CCTV systems.

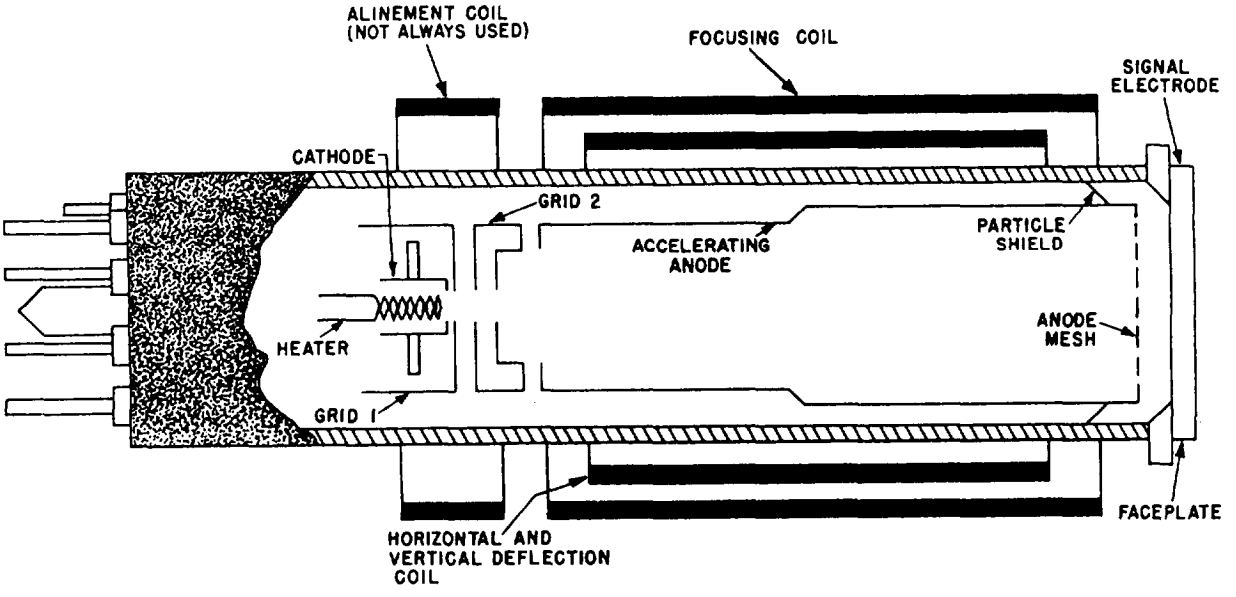


Figure 8-6.-Structure of vidicon.

VIDICON TUBE

The vidicon camera tube (fig. 8-6) has a transparent conductive coating, called the signal electrode, on the inner surface of the faceplate; a layer of photoconductive material deposited on the signal electrode; an accelerating anode; a focusing coil; and a cathode emitter for producing a beam of electrons. Associated with the vidicon tube are the alignment coil, the focus coil, and the deflection coils.

The alignment coil produces a magnetic field that is variable in both magnitude and direction, and is used to adjust the direction of the electron beam so that it is parallel to the field of the focus coil. Control of the alignment coil current is accomplished in the control unit.

The focus coil surrounds the vidicon tube and establishes a magnetic field along the axis of the tube. It is connected between the +300-volt and the +150-volt power supplies through a resistor located in the rotatable section of the base unit.

The vertical and horizontal deflection coils are excited by linear sawtooth currents from the control unit. These currents produce the field that causes the beam to scan the photosensitive layer.

The beam of electrons is directed toward the layer of photoconductive material (on the cathode side of the signal electrode) at a medium velocity because of the relatively low accelerating potential between the cathode and the accelerating electrode. A sharp beam is formed by the electrostatic field of the focus electrode, the axial magnetic field of the focus electrode, and the axial magnetic field of the focus coil surrounding the tube. The electron beam is deflected by the deflection coil in such a way as to scan the photoconductive layer. When no light is permitted to reach this layer, its resistivity is extremely high. One side of the layer is maintained at a small positive potential of 0 to 70 volts by direct contact with the signal electrode.

When light from the scene being televised passes through the faceplate and is focused on the photoconductive layer, the resistivity of this material (which has been extremely high) is reduced in proportion to the amount of light reaching it. Because the potential gradient between adjacent elements in the photoconductive layer is much less than the potential gradient between opposite sides of the layer, electrons from the beam side of the layer leak by conduction to the other side between scans of the electron beam. Consequently, the potential of each element on the beam side approaches the potential of the signal electrode side, and it reaches a value that varies with the amount of light falling on the element. On the next scan, the electron stream replaces a number of electrons on each element just sufficient to return it to the potential of the cathode. Because each element is effectively a small capacitor, a capacitive current is produced in the signal-electrode circuit that corresponds to the electrons deposited as the element is scanned. When these electrons flow through the load resistor in the signal-electrode circuit, a voltage, which becomes the video signal, is produced.

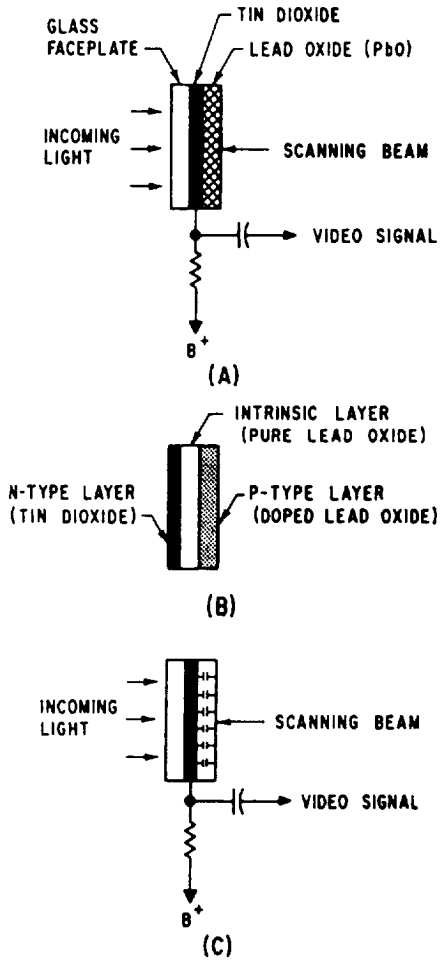


Figure 8-7.-Plumbicon target.

PLUMBICON

The plumbicon tube is similar in appearance and operation to the vidicon. It has several advantages over the vidicon. The plumbicon tube has a more rapid response and produces high quality pictures at lower light levels. Because of its small size and low power consumption, the plumbicon tube is well suited for use in transistorized TV cameras. Its simplicity and spectral response to primary colors make it particularly useful in color cameras.

A unique feature of the plumbicon is that its color response can be varied by the manufacturer. It is therefore available with spectral responses for each of the primary colors. The color response of each tube is identified by the letter R (red), G (green), or B (blue) following its basic number. For example, a plumbicon for a green channel maybe designated 55875G.

Figure 8-7 provides a simplified diagram of a plumbicon target. The glass faceplate (fig. 8-7, view A) has its inner surface coated with tin dioxide. This thin, transparent layer is the signal plate of the target. The tin dioxide itself is a strong N-type semiconductor. Two layers of lead oxide are deposited on the scanning side of the target. The first of these two layers is almost pure lead oxide, an intrinsic semiconductor. The second layer of lead oxide is doped to form a P-type semiconductor. As shown in figure 8-7, view B, the three layers form a P-I-N junction.

Light from the televised scene passes through the layer of tin dioxide and is focused on the

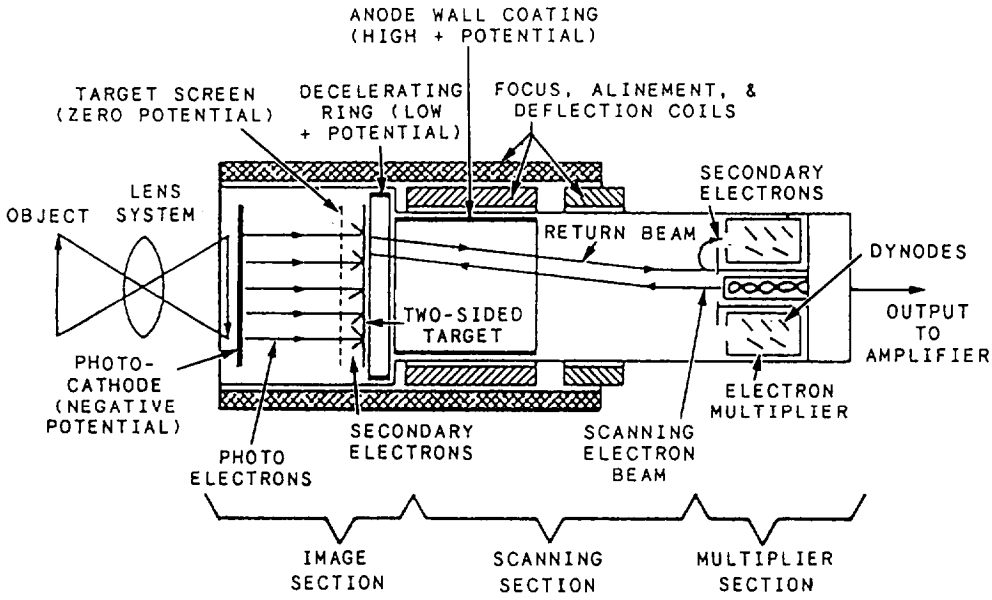


Figure 8-8.-Structure of image orthicon.

photoconductive lead oxide. Note in figure 8-7, view C, that each picture element charge acts like a capacitor whose positive plate faces the scanning beam. The target signal plate forms the negative plate. As the low-velocity scanning beam strikes each charged element, it releases electrons that neutralize the "capacitors."

IMAGE ORTHICON

The image orthicon (fig. 8-8) is an ultrasensitive TV camera pickup tube. The tube requires only 8 to 40 candelas for light, and is used in modem conventional and closed-circuit TV systems. When this tube is used, a light image from the subject (arrow at extreme left) is picked up by the camera lens and focused on the light-sensitive face of the tube, releasing electrons from each of the thousands of tiny globules in proportion to the intensity of the light striking it.

These electrons are directed on parallel courses from the back of the tube face to the target, from which each striking electron liberates several more electrons, leaving a pattern of proportionate positive charges on the front of the target. When the back of the target is scanned by the beam from the electron gun in the base of the tube, enough electrons are deposited at each point to neutralize the positive charges. The rest of the beam returns, as shown in figure 8-8, to a series of electron multiplier stages or dynodes surrounding the gun.

Each dynode is a metallic disk with openings similar to a pinwheel and operates at a positive potential of 200 to 300 volts greater than the preceding dynode. Multiplication occurs through secondary emission at each dynode. If five dynode stages are used, each having again of 4, again of 4 x 4 x 4 x 4 x 4 or approximately 1,000 is realized in the multiplier section. Considering the gain of 5 in the image section, the overall gain of the image orthicon is 5,000. This gain allows this particular pickup tube to operate with relatively less light than the plumbicon or vidicon. The electrons from the last dynode are routed through a signal-developing resistor to an extremely high B+ voltage. The output signal is then coupled to the first stage of video amplification.

SECONDARY ELECTRON CONDUCTION (SEC) TUBE

This is a vidicon-like tube with a special target that uses secondary electron conduction (SEC). In this tube, light is focused on the photocathode, which emits electrons into the tube. These electrons are focused to form an image of electron streams that strike the SEC target. The electrons are accelerated to approximately

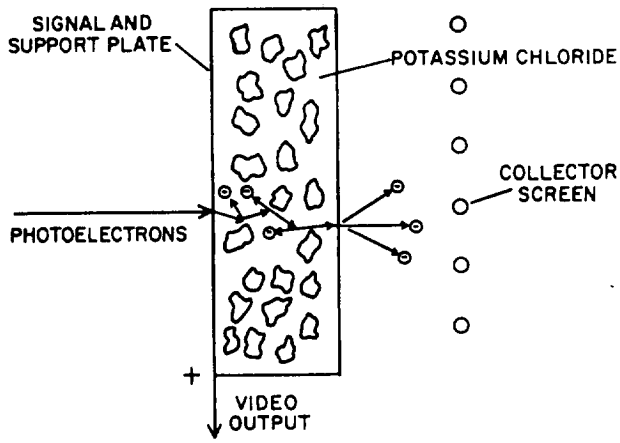


Figure 8-9.-SEC camera tube target operation.

10,000 electron volts by the time they strike the target. The SEC target intercepts these streams of electrons. A resulting great number of secondary electrons from each electron striking the target are emitted and stored (fig. 8-9). The photoelectrons possess sufficient energy to penetrate the thin, metallic signal and support plate. As they travel through the porous potassium-chloride layer, many secondary electrons are emitted as the beam strikes the interlined particles. These secondary electrons either escape to the positive collector screen or travel through the spaces of the porous layer to the positive collector plate. This loss of electrons produces a positive charge on the scanned side of the target. Several hundred secondary electrons are emitted for each incident electron, producing a substantial gain at the target.

The video signal is developed from the target by the scanning beam discharging the positively charged areas of the target in the same manner as in a vidicon tube. This charging current, flowing out of the signal plate connection, is then amplified by an external amplifier.

The SEC tube has applications in extremely low-light, nighttime military TV systems where a high internal amplification and a fast speed or response to moving images are important.

CAMERA CIRCUITS

Learning Objective: *Recall the purpose of camera circuits to include block diagram analysis and camera circuit analysis. Recall the difference between black and white and color cameras.*

The purpose of camera amplifier circuits is to amplify the extremely low output signal from the

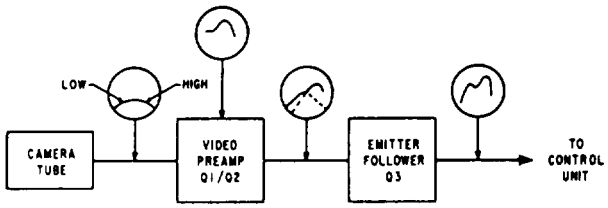


Figure 8-10.-Typical camera unit block diagram.

camera tube. The video signals must be increased in amplitude to overcome tube noise, random circuit noise, and hum that may be injected into the video output.

Camera circuits are carefully designed to amplify high- and low-frequency signals equally. The output of the camera tube (fig. 8-10) must be as large as possible. The output circuit contains high resistance, providing good gain at low-frequencies but low gain at high frequencies. Because a video signal contains both high- and low-frequency components, the video stages are designed to amplify more at some frequencies than at others.

BLOCK DIAGRAM ANALYSIS

Video preamplifier Q1/Q2 (fig. 8-10) serves as a "peaker" circuit because it provides greater amplification to the higher video frequencies. Note that the resultant frequency response curve at the output of Q1/Q2 is reasonably flat over a wide band of video frequencies.

The video signal from Q1/Q2 is routed through emitter follower Q3 and developed across a resistor in the camera control unit. Emitter follower Q3 is used to match the output impedance of the camera unit to the input impedance of the control unit.

CAMERA CIRCUIT ANALYSIS

Figure 8-11 illustrates the circuits in a typical camera unit. The vidicon is represented schematically

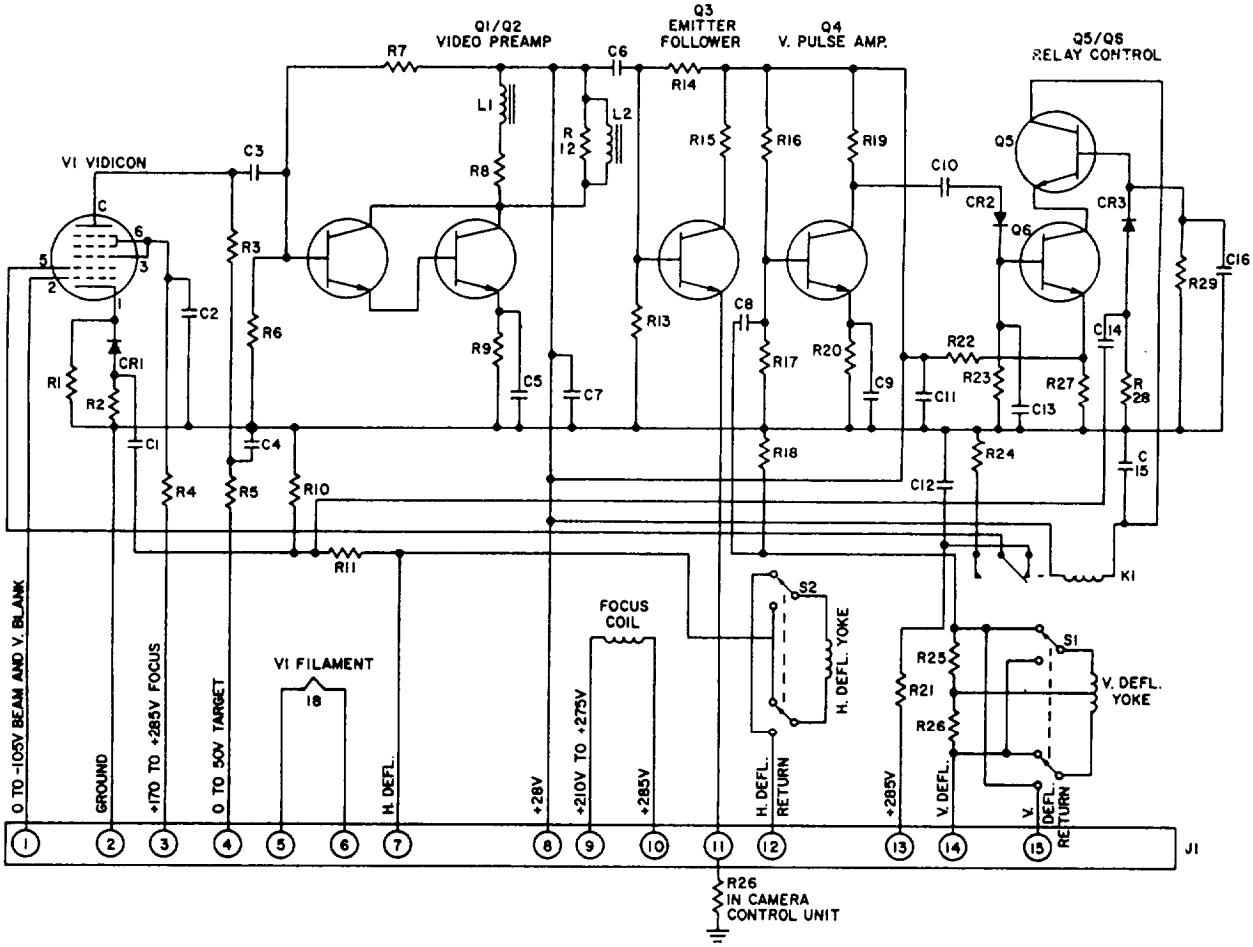


Figure 8-11.-Typical camera circuit.

by V1. The elements in a vidicon tube have functions similar to those in an ordinary electron tube; however, the vidicon tube terminology differs somewhat because the control grid is called the beam control, and the plate the target control.

When the camera circuit is operating, the video signal leaves the vidicon through pin C and is developed across R3. A variable dc focusing voltage is applied to pin 6 of V1 via decoupling network R4 and C2. A fixed positive accelerating voltage is applied from pin 13 of J1, through the contacts of K1, to pin 5 of V1. K1 is normally energized while the camera is operating. (K1 is shown energized.) Vertical blanking pulses and a variable beam control bias are applied from pin 1 of J1 to pin 2 of V1. Pulses from the horizontal deflection yoke are coupled through S2 (shown in normal position), differentiated by C1 and R2, clipped by CR1 and R1, and applied as sharp positive horizontal blanking pulses to the cathode of V1.

Video output from V1 is coupled through C3 to the base of Q1. Q1 and Q2 comprise a compound-connected transistor circuit known as a Darlington amplifier. This type of circuit was selected for use as the preamplifier stage because it has high input impedance, very high current gain, and excellent frequency response. Forward bias for the stage is provided by voltage divider components R6 and R7. R9 and C5 are primarily intended for temperature compensation. The amplified video output from the collector of Q2 is developed across R8 and L1 (a shunt-speaking circuit), and coupled through R12 and L2 (a series-peaking circuit) to the base of emitter follower Q3. This stage provides an impedance match to allow coupling of the video via coaxial cable to emitter load resistor R26 in the camera control unit.

Vidicon Protection Circuitry

Q4, Q5, and Q6 are used in circuitry designed for vidicon tube protection. Recall from your knowledge of cathode-ray tubes that it is undesirable to allow the electron beam to remain too long on the same spot on the CRT screen. If this occurs or the beam retraces the same line too often, the phosphor coating on the CRT screen will "burn." A vidicon target is comparable to a CRT screen, because if the vertical and horizontal sweep signals are not present, the vidicon target could become permanently damaged.

Pulsed amplifier Q4 serves to amplify a sampling of the vertical sweep signal, which enters the camera on pin 14 of J1. This signal is attenuated by R26 and R25 and coupled through C8 to the base of Q4. After being amplified by Q4, the signal is coupled through C10, rectified by CR2, and faltered by R23 and C13. The resultant positive dc voltage is used as forward bias for Q6. Note Q6 is connected in series with Q5, forming (in terms of digital computers) an AND gate. To allow current to flow through both Q5 and Q6, forward bias for Q5 must also be provided.

Forward bias for Q5 is developed, logically enough, by action of the horizontal sweep signal, which consists of pulses at 15,750 Hz. This signal enters the camera unit through pin 7 of J1. The pulses are attenuated by R11, coupled through C 14 to the anode of CR3, and then rectified R29 and C16 filter the resultant dc, which is used as forward bias for Q5.

Remember that Q6 is forward biased only as long as a vertical deflection signal is present in the camera unit, and Q5 is forward biased only as long as a horizontal deflection signal is present. Therefore, BOTH the vertical and horizontal deflection signals must be present for current to flow through Q5 and Q6.

Note that the coil of current-sensitive relay K1 is in the collector circuit of Q5. If Q5 and Q6 are conducting, K1 is energized. Look at pin 13 of J1. A +285 volts dc is routed from pin 13 through R21 and the energized contacts of K1 to pin 5 of the vidicon. If the +285 volts dc is NOT present, the vidicon is cut off. If K1 de-energizes, a ground is applied through R24 to pin 5 of the vidicon tube, assuring that the tube is cut off.

To summarize, if either the horizontal or vertical deflection signal is not present, K1 will de-energize, causing the vidicon to cut off. This provision, of course, serves to protect the vidicon target from being "burned" if sweep signals are lost.

Special Effects

Switches S1 and S2, connected across the vertical and horizontal deflection yokes, respectively, are used to initiate special video effects. Closing S1 causes the image to invert because the direction of current flow is reversed through the vertical yoke. Similarly, closing S2 results in a mirror image because current now flows in the opposite direction through the horizontal yoke.

COLOR CAMERAS

The camera unit used for color pickup is similar to the black and white camera previously discussed. In the

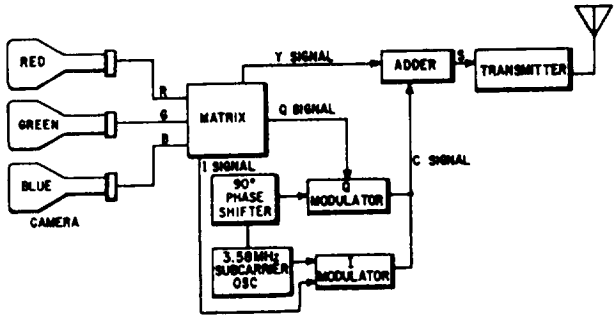


Figure 8-12.-Color TV transmitter block diagrams.

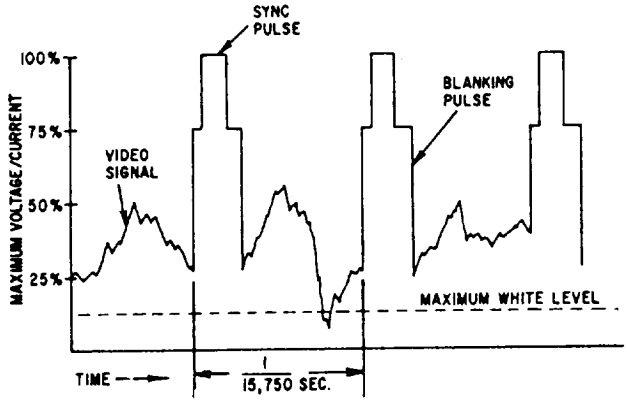


Figure 8-13.-Composite video signal.

color camera, the camera unit (fig. 8-12) has three pickup tubes (one for each of the primary colors). Filters and mirrors are used to direct the right color of light to its respective pickup tube. The output of the camera unit provides a red, green, and blue video signal to the matrix system (a circuit that proportions the primary signals to produce the correct brightness and chrominance colors). The three primary colors are identified as *R* for red, *G* for green, and *B* for blue.

The matrix section is essentially a resistive voltage-divider circuit that proportions the primary color signals, as required, to produce the brightness and chrominance signals. With red, green, and blue color video voltages as inputs, the three video signal output combinations formed are the following:

1. A luminance signal, designated the Y signal, which contains the brightness variations of the picture information.

2. A color video signal, designated the Q signal, which corresponds to either green or purple picture information.

3. A color video signal, designated the I signal, which corresponds to either orange or cyan picture information.

The I and Q signals together contain the color information for the chrominance (hue and saturation) signal.

Additional information concerning the color video signal and its transmission is presented later in this chapter as part of the coverage of the composite video signal.

COMPOSITE VIDEO

Learning Objective: *Identify the characteristics and purpose of the composite video signal.*

The composite video signal contains all the information needed to reproduce the picture. Its contents include the video from the camera unit, synchronizing pulses to synchronize the transmitter signal with the receiver/monitor, and blanking pulses to obliterate the retrace signals from the picture tube. The video signal is combined with the blanking pulse (fig. 8-13), and the sync pulse is placed on top of the blanking pulse.

In the composite video signal, successive values of voltage and current amplitudes are shown against values of time during the scanning of three horizontal lines. During the time when blanking and sync pulses are being transmitted, no video appears. The overall signal amplitude is divided into two parts: the lower 75 percent is for video, and the upper 25 percent is devoted to

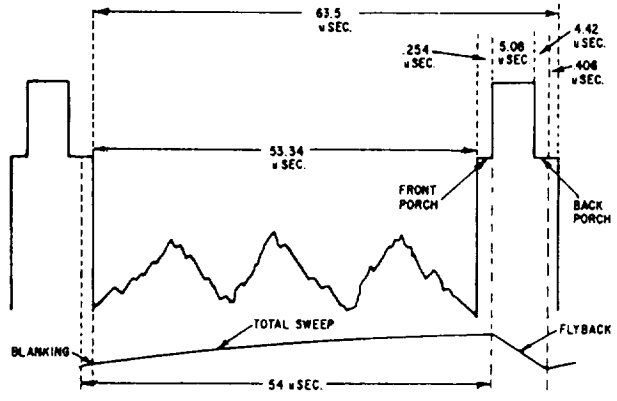


Figure 8-14.-Composite video time allocations.

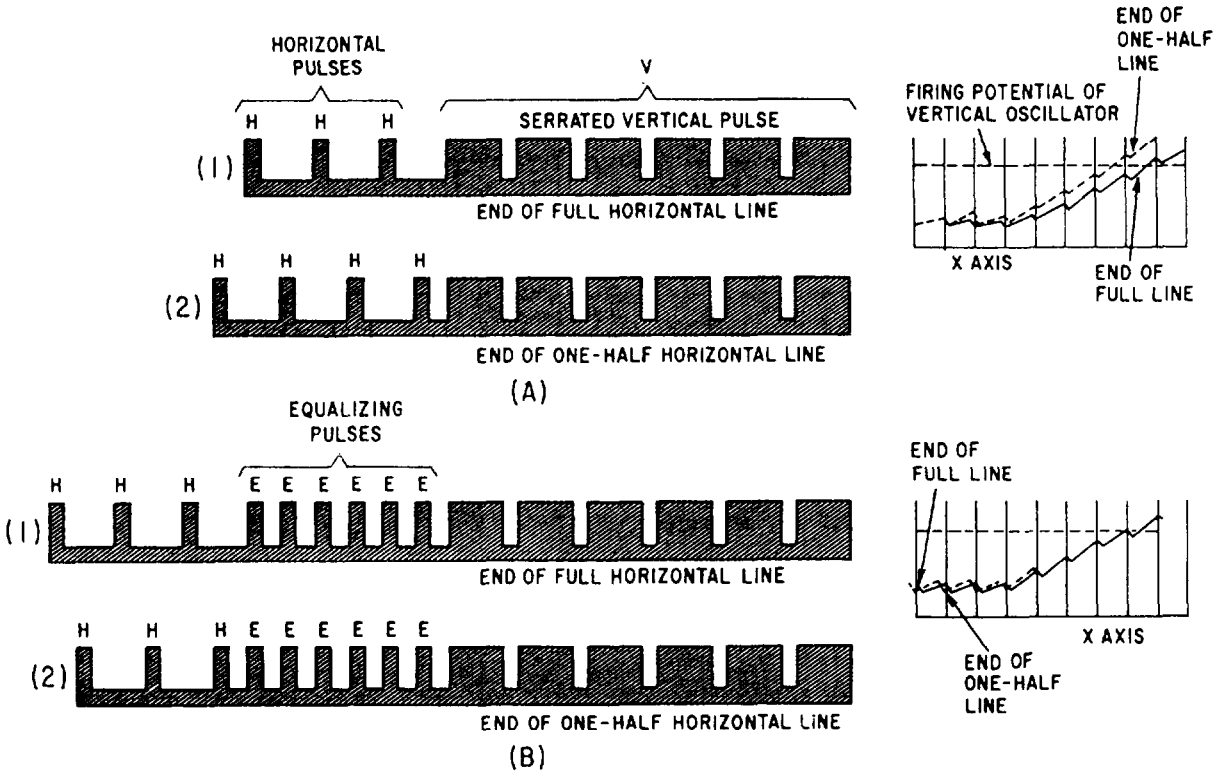


Figure 8-15.-Sychronizing pulse forms. (A) Without equalizing pulses; (B) with equalizing pulses.

synchronizing pulses. Standardization is necessary to ensure the transmitted signal is suitable for all receiver/monitors.

The lowest amplitudes correspond to the whitest parts of the picture. The picture becomes blacker as amplitude increases toward 75 percent. This standard of transmission is called "negative transmission," which is defined as decreasing signal amplitude for decreasing light intensities. As the level reaches 75 percent, the grid becomes negative, cutting off the picture tube. The absence of light establishes the blackest level, which is the case when the blanking level occurs.

Details of the horizontal blanking and sync pulse are shown in figure 8-14. The interval of the complete scan is 63.5 microseconds since the horizontal frequently is 15,750 Hz. The horizontal blanking pulse is 10.16 microseconds, or about 16 percent of total sweep. The sync pulse, superimposed on the pedestal, occupies 5.08 microseconds or one-half the blanking time.

The part of the pedestal just before the sync pulse (0.254 microsecond) is called the "front porch"; the portion following the sync pulse (4.826 microseconds) is called the "back porch." The front porch blanks the right side of the picture screen just before the sync pulse begins. Flyback occurs with the leading edge of the sync pulse and continues for 4.42 microseconds of the back porch. The next sweep starts, but the left side of the screen is blanked for 0.406 microsecond of the starting sweep. This action strives to maintain a straight left edge.

The vertical blanking pulse blanks the picture during retrace time when the electron beam has completed one field and is returned to the top ready to start the next field. Immediately following the last active line, the video signal is brought up to the black level by the vertical retrace.

So far in the discussion, little has been said about the special form of the combined synchronizing pulses. The form and the timing of the synchronizing pulses are such that the horizontal and vertical oscillators are triggered at exactly the right instant to keep the sweep in the camera tube and the sweep in the picture tube locked in step. Because the horizontal oscillator must be triggered during the vertical sync pulse (to prevent the horizontal oscillator from drifting out of control), the vertical pulse is serrated, as shown in figure 8-15. As a result, the vertical pulse is chopped into six pieces. The fluctuations resulting from serrations do not affect the operation of the vertical oscillator, but only keep the horizontal oscillator properly triggered.

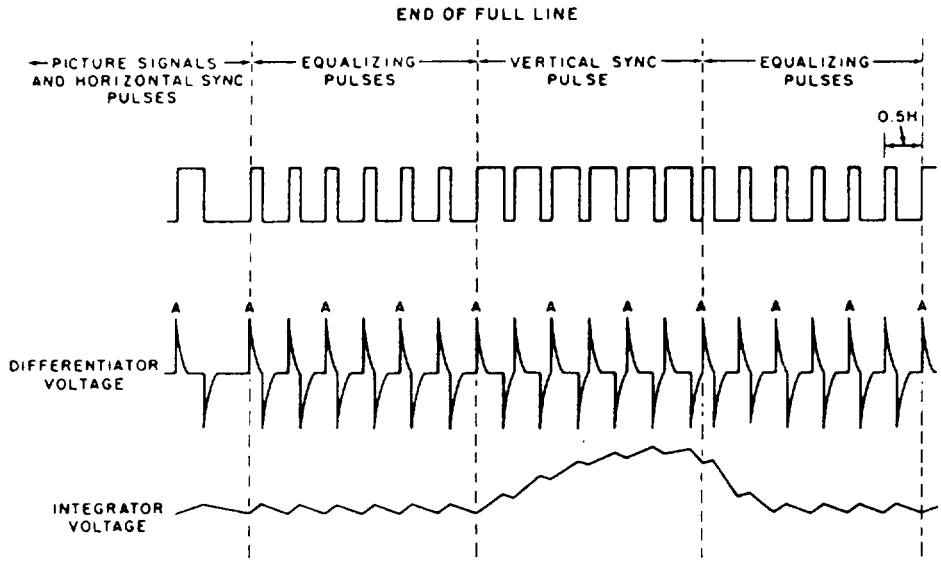


Figure 8-16.–Form of sync pulses comprising the output integrator and differentiator voltages.

In addition to the serrations in the vertical pulse, equalizing pulses are necessary before and after each vertical pulse. The necessity for equalizing pulses (labeled E in figure 8-15, view B) may be explained as follows:

First, assume that no equalizing pulses are used, as shown in figure 8-15, view A. If the vertical pulse is inserted at the end of the field, which occurs simultaneously with the end of a full horizontal line, part (1), the firing potential of the vertical oscillator is reached at the correct time to produce the desired interlaced scan. If the vertical pulse is inserted at the end of a field, which recurs simultaneously with the end of a half horizontal line, part (2), the firing potential of the vertical oscillator is reached too early. This happens because the slight charge on the capacitor in the triggering circuit of the vertical oscillator (due to each horizontal pulse) does not have time to leak off before the vertical pulse arrives. The residual voltage across this capacitor, plus the voltage due to the vertical pulse, causes the vertical oscillator to fire too soon.

Second, the situation is corrected, as shown in figure 8-15, view B, by the use of equalizing pulses. The buildup of the vertical pulse across the capacitor now begins at the same point, whether the vertical pulse arrives at the end of a full line or at the end of a half line. In other words, the equalizing pulses cause the potential on the capacitor to be at the same level (at the same time the vertical pulse arrives) whether the vertical pulse occurs at the end of a half line or at the end of a full line. Although not shown in the figure, equalizing pulses are also used after the vertical pulses.

To get a better idea of the form of sync pulses, study figure 8-16. You can see that a number of horizontal scans are lost during the vertical flyback time.

CONTROL UNIT

Learning Objective: *Describe the function of the control unit to include the sync generator circuits and sync and blanking insertion.*

Sync generators in the control unit provide reference signals to keep the scanning signals in the monitor in step with those in the camera. Another function of the control unit is to maintain a stable phase relationship between the vertical and horizontal scanning signals. If the phase relationship between these signals is allowed to vary, the 2:1 interlaced scan will not be stable. To obtain the required phase lock, both the horizontal and vertical sync signals are generated in a common master oscillator. Then, they are converted to signals having the required frequencies by the use of appropriate count-down circuits.

Figure 8-17 shows the block diagram of a typical pulse-counter type of sync generator. Master oscillator Q1 generates the 31.5-kHz signal from which both the horizontal and vertical drive signals (which will later become sync signals) are derived. The frequency and phase of the master oscillator signal are stabilized by phase detector Q9. This stage compares both the

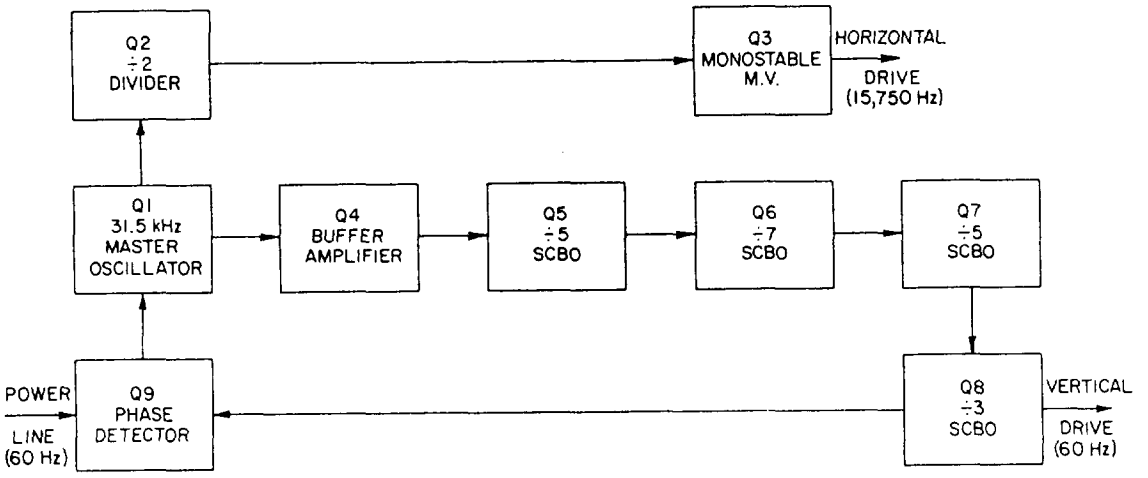


Figure 8-17.-Sync generation using pulse counters.

frequency and phase of the vertical drive signal with those of the 60-Hz power-line reference signal. Frequency divider Q2 serves to halve the master oscillator frequency to 15,750 Hz. This signal is used to trigger Q3, the horizontal-drive multi vibrator.

Buffer amplifier Q4 is an isolation stage used to shield the master oscillator from the loading effects of Q5, a single-cycle blocking oscillator (SCBO). Q5 is the first of four count-down circuits (Q4, Q6, Q7, and Q8) used to produce the 60-Hz vertical drive signal. Note that Q8 has two output signals, one of which is the vertical drive signal. The other output signal is routed back to phase detector Q9 where (as stated before) its frequency and phase are compared with those of the 60-Hz power-line reference voltage. The vertical and horizontal drive signals are then routed from the sync generator to sync insertion circuits in another section of the control unit.

SYNC GENERATOR CIRCUITS

Sync generators usually consist of only a few basic circuits, with these circuits being repeated in different areas. Simplified versions of some of the more common sync generators are described in this section.

Master Oscillator

Figure 8-18 shows a simplified schematic of a 31.5-kHz master oscillator, in which Q3 is connected to form a modified Armstrong oscillator circuit. The frequency of this free-ruining master oscillator is determined by the values of C6 and T1. The inductance of T1 can be changed by adjusting the position of its movable core.

Q2 and associated circuitry comprise a reactance stage capable of changing the master oscillator frequency as a function of the dc control voltage applied to the base of Q2. The dc control voltage is the output of the phase detector previously discussed in the block diagram analysis.

In general, if the output signal of an amplifier can be made to lead or lag the input signal, that amplifier can function electrically as a variable capacitor. Refer again to figure 8-18 and imagine Q2 and associated circuitry to be a variable capacitor in series with C6. C4 and varicap diodes X1 and X2 form a capacitive voltage divider in the feedback loop of Q2. Being capacitive, the feedback loop causes the correction signal voltage to lag the signal current in the loop. The degree of lag is proportional to the amplitude of the correction signal. The apparent change in capacitance "tunes" the master oscillator to the "desired' frequency.

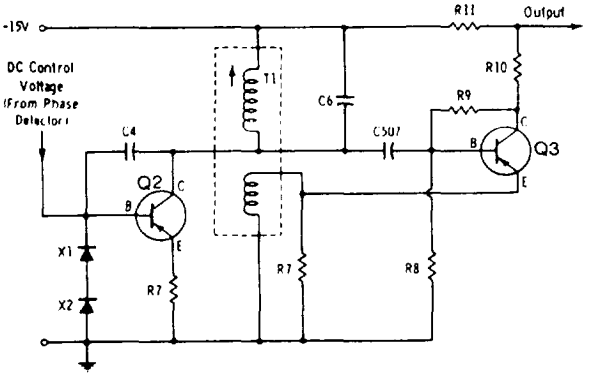


Figure 8-18.-Master oscillator.

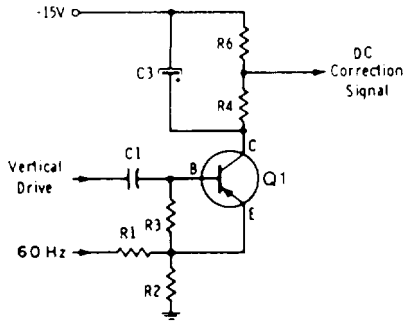


Figure 8-19.-Phase detector.

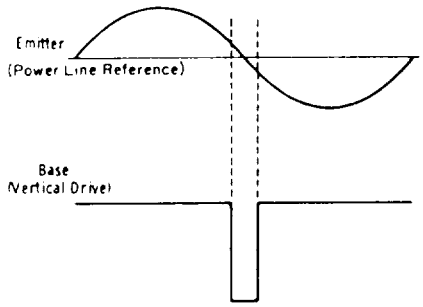


Figure 8-20.-Phase relationships in the phase detector.

Phase Detector

Figure 8-19 shows a phase detector circuit such as used to provide a dc correction voltage to the reactance stage previously discussed and shown in figure 8-18. Note in figure 8-19 that a vertical drive signal is fed to the base of Q1 and a 60-Hz ac reference signal is fed to the emitter. Q1, which is normally cutoff, is brought into conduction by the vertical drive signal (consisting of negative pulses). The normal phase relationship between the vertical drive and the 60-Hz reference signal is shown in figure 8-20. As long as a vertical drive pulse appears when the reference signal passes through zero, a "normal" control voltage will be produced. If, however, the normal phase relationship tends to drift, a larger or smaller correction voltage will be produced. C3 in the collector circuit (fig. 8-19) acts as a filter capacitor, ensuring that a smooth dc correction voltage is applied to the reactance stage.

Frequency Divider

A single-cycle blocking oscillator (SCBO) (fig. 8-21) circuit is often used as a frequency divider circuit (or count-down circuit) in sync generators. In this circuit, Q9 is normally cut off. Input pulses from a preceding stage gradually build up a charge across C17 until a forward bias condition is achieved. As Q9 begins to conduct, a regenerative signal is coupled through T2

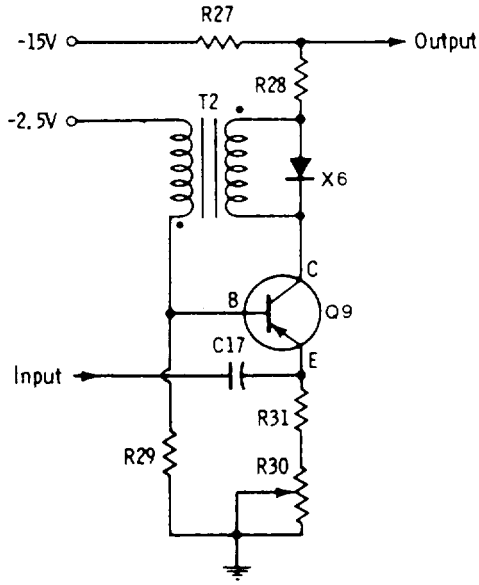


Figure 8-21.-Blocking-oscillator frequency divider.

to the base of Q9. This regenerative signal causes Q9 to reach saturation rapidly. As soon as Q9 saturates, C17 discharges through the transistor. When C17 discharges sufficiently, Q9 abruptly stops conducting. The collapsing field around the secondary of T2 assists in driving Q9 into sharp cutoff. Diode X6 swamps T2, suppressing resonance oscillations. The values of C17, R31, and R30 have been specially chosen to allow the transistor to conduct when the appropriate number of input pulses has been applied. Figure 8-21 shows appropriate values for a 5:1 countdown.

SYNC AND BLANKING INSERTION

The earlier discussion of camera circuits revealed that the video signal is developed in the camera tube and routed through the video amplifier and peaking circuits to the control unit. After entering the control unit, the video signal is subjected to additional amplification and peaking. Video from the camera tube is referred to as "raw" video; that is, no sync or blanking signal is included. Therefore, it is necessary to insert the sync and blanking signals to produce a composite video signal. Sync and blanking insertion are often performed in the control unit. Typical insertion circuits arc discussed in the following paragraphs.

Blanking Insertion

Blanking signal insertion is most often performed before sync signal insertion. Before the blanking signal can be inserted, the proper dc reference level must be established by clamping circuitry. Clamping is needed because a coupling capacitor cannot pass dc, and it also

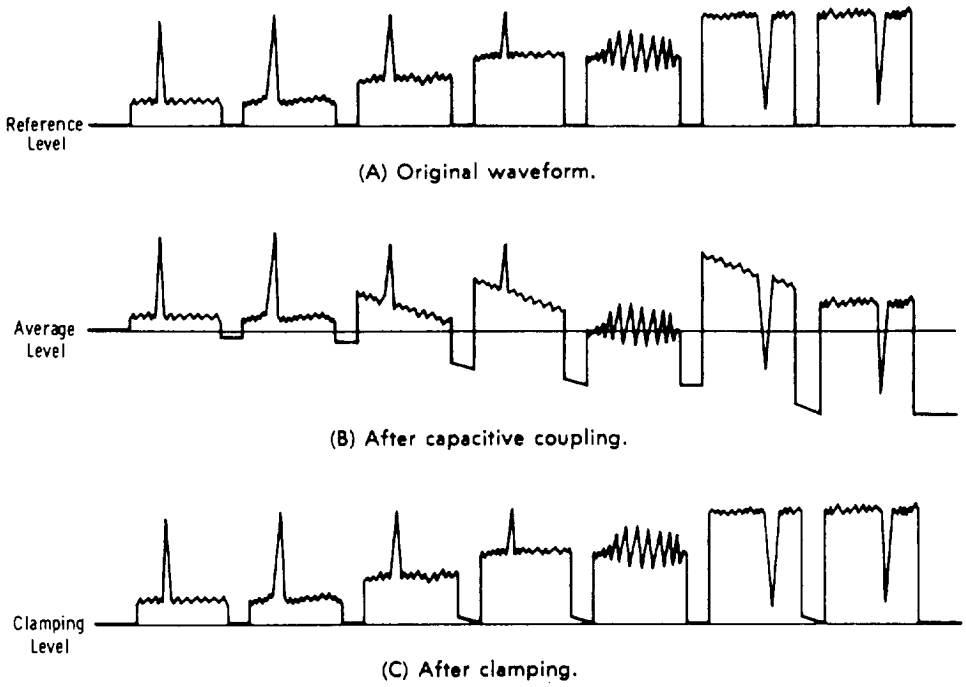


Figure 8-22.-Clamping waveforms.

has difficulty in passing low-frequency ac signals. Note in figure 8-22 that the RC-coupled signal assumes an average level, varying about the bias voltage level of the preceding stage. Keyed clamper circuits are used in most of the modem CCTV systems. Consequently, no clamping occurs unless a keying pulse is present. Typically, the horizontal drive signal is used as a keying pulse.

Figure 8-23 illustrates a blanking insertion circuit. In this circuit, Q8 is forward biased each time a clamping pulse (horizontal drive pulse) is applied to its base.

When Q8 conducts, the video signal is clamped at the blanking level established by R31. The clamped signal (video) is applied to the base of emitter follower Q9, amplified, and directly coupled to the base of Q10, the blanking mixer stage. Note that the level of forward bias for Q10 is determined by the potential at the emitter of Q9, which is itself controlled by adjustment of R31. The blanking signal is therefore clipped (limited) by Q10 at a level established by R31. The output of this stage is routed to a sync insertion circuit.

Sync Insertion

The sync signal is normally inserted into the composite video signal after the blanking signal has been inserted. The sync-insertion circuit used is generally a simple additive mixer similar to the sync-adding network shown in figure 8-24. In this

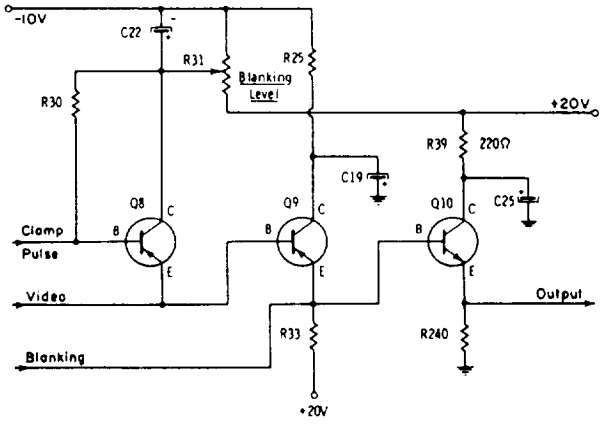


Figure 8-23.-Blanking-insertion circuit.

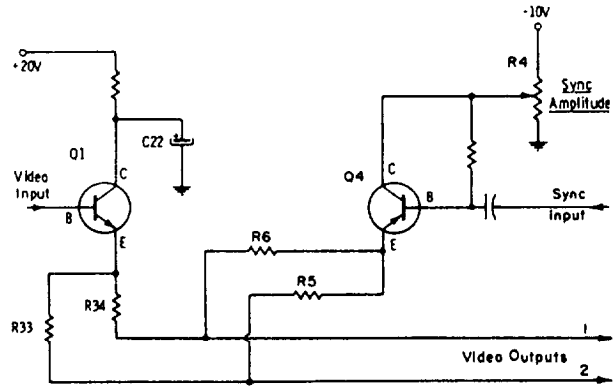


Figure 8-24.-Sync-adding network.

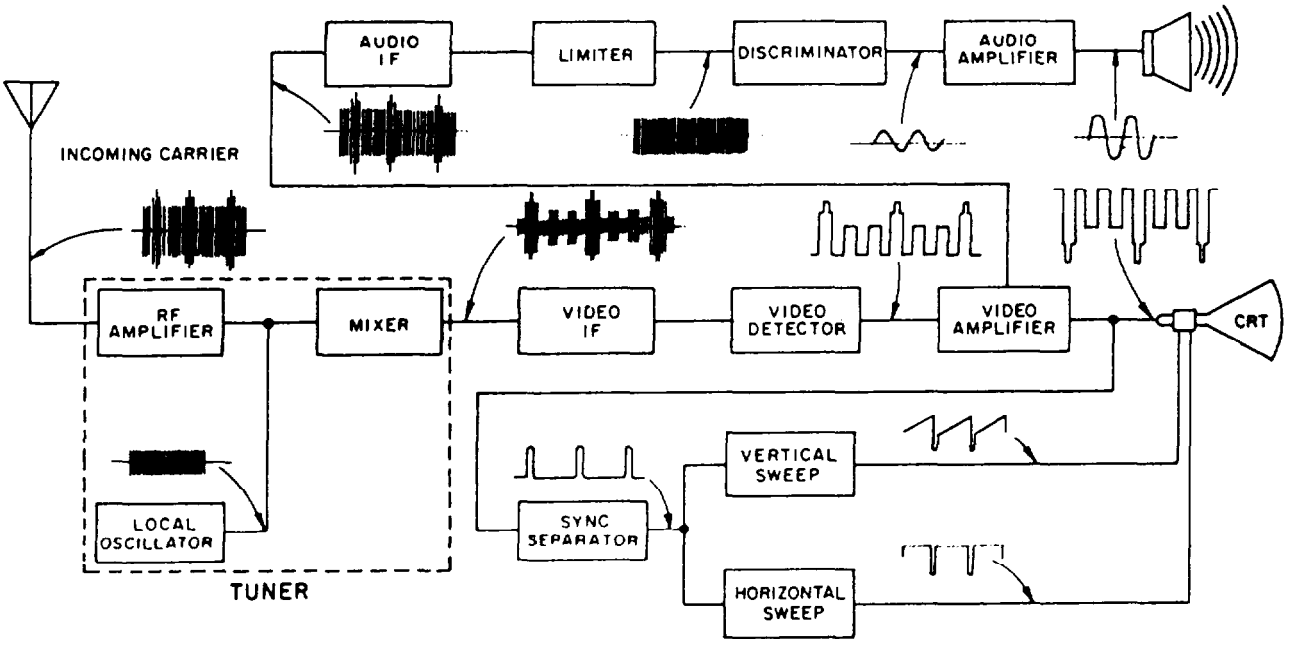


Figure 8-25.-Sectional block diagram with waveforms.

network sync is added (via R5 and R6) to the video signal at the video output terminals.

Isolation amplifier Q4 prevents the coupling of video to the sync input terminals. Adjustment of potentiometer R4 sets the gain of Q4, thus controlling the sync signal amplitude. Transistor Q1 is the video-output stage.

RECEIVERS

Learning Objective: *Recognize the major sections of a TV receiver to include tuners, solid-state tuners, video IF amplifiers, traps, circuits, video detectors, video amplifiers, and dc restorers.*

Figure 8-25 is a sectional block diagram of a TV receiver. Each of the major sections is discussed in the following text.

TELEVISION TUNERS

TV tuners, like those of communications receivers, consist essentially of three parts: the RF amplifier, the oscillator, and the mixer. The function of the tuner, often called the front end, is to amplify the relatively weak input signal, to produce a locally generated RF signal, and to mix these two signals in such a way to produce signals at the chosen audio and video intermediate frequencies. Because of the wide band of frequencies (approximately 6 MHz) that must be passed, overcoupling and loading are employed to increase the frequency response.

A block diagram of a typical front end is shown in figure 8-26. Various methods of tuning are used. Rotary switches that connect the correct capacitors and inductors into the circuit are often employed. In some systems the "tuned line" is used, and in others, separate components are connected for each channel. Turret tuners are also widely used. In this system of tuning, elements pretuned to the desired frequency may be plugged into a clip in the turret. For compactness and precision, printed circuits are sometimes employed.

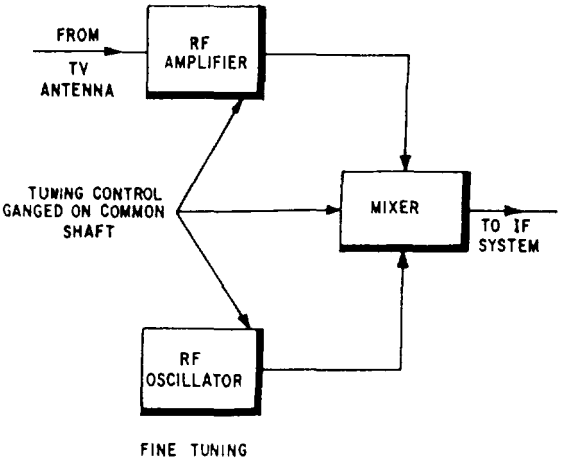


Figure 8-26.-Block diagram of TV front end.

RF Amplifiers

Because the TV signal available at the input of the TV receiver is weak, it is desirable to amplify the signal in a stage of RF amplification before it is applied to the mixer. Various types of RF amplifiers employed in TV receivers include common emitter, pentode, and field-effect transistors (FETs), all of which are discussed in detail in Navy Electricity and Electronics Training Series (NEETS), Module 7. The FETs are being used extensively in TV primarily because of the three advantages discussed in the following paragraphs.

In an ordinary transistor, current flow through the device is dependent upon both the hole flow and the electron flow. Since both polarities of the charges are required for their operation, ordinary transistors are sometimes referred to as bipolar devices. In comparison, a FET is an unipolar device because the flow of current through it is either by hole flow or electron flow, but not by both. Since an electron-hole combination causes noise, a unipolar device generates less noise in its operation. This is a definite advantage in RF amplifiers.

Another advantage of FET operation is its high input impedance. In this regard, it is similar in operation to the vacuum tube. Unlike transistor amplifiers that require current flow in the base circuit, and therefore require input power, a FET circuit may be designed that has virtually no current flow in the gate circuit.

The third advantage of a FET is its square law operation. Current flowing through the FET is directly proportional to the square of the voltage on the gate, which means it is a square law device rather than a linear device.

Figure 8-27 shows a typical (simplified) circuit with a dual-gate FET used as an RF amplifier. An AGC voltage is developed across a voltage divider comprised of resistors R1, R2, and R3. Capacitors C1 and C2 filter the AGC voltage so that only a dc voltage is applied to the gates. The largest part of the AGC voltage is applied to gate G2.

The RF signal from the antenna is applied through C3 and developed across R3. This signal voltage appears on gate G15. Thus, the amount of current through the FET is controlled by two voltages: the AGC voltage and the RF input voltage. The source bias is supplied by R4, and C4 is used to prevent degeneration. The output signal is taken from the secondary of transformer T15 and fed to the mixer stage.

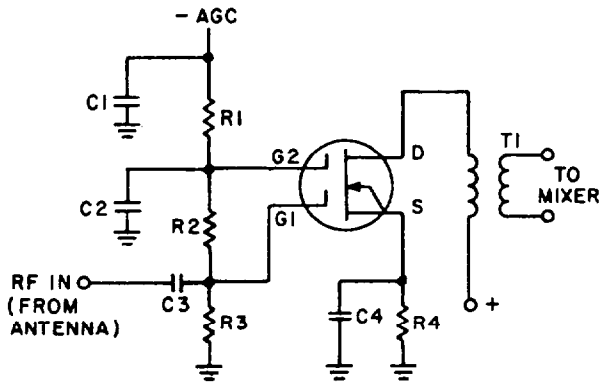


Figure 8-27.-Dual-gate MOSFET RF amplifier.

Local Oscillators

Local oscillators provide a signal that is heterodyned in the mixer with the RF signal. As in the case of superheterodyne radio receivers, the local oscillator frequency is normally above the incoming RF frequency by a frequency that is equal to the receiver IF frequency.

The local oscillator frequency is changed whenever the channel selector is changed/tuned from station to station. The fine-tuning control varies the oscillator frequency over a narrow range. The setting of the fine-tuning control is very important if a satisfactory color picture is to be obtained.

Mixers

As shown in figure 8-26, the mixer receives a signal from both the RF amplifier and the oscillator circuits and converts these to a different frequency output, called the intermediate frequency (IF) signal. This conversion is accomplished by applying both the RF and oscillator voltages to a nonlinear amplifier. If the amplifier is nonlinear, heterodyning will take place, and the output signal will include both the sum and difference frequencies, as well as the two original frequencies. Assuming the frequency difference between the RF and oscillator signals is appreciable, a selective network tuned to the lower frequency component of the output will pass the desired IF only.

The dual-gate MOSFET makes an ideal mixer amplifier. This type of transistor is shown in figure 8-27. Recall it had two gates for inputs. One of the gates could be used for the RF input and the other for the oscillator input, or both inputs can be delivered to one gate, and the other used as a screen to isolate the input and output stages of the mixer.

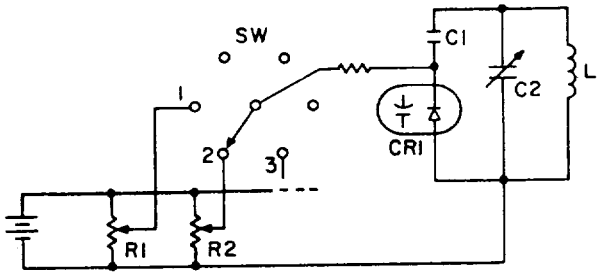


Figure 8-28.-Varactor-controlled tuning circuit.

SOLID-STATE TUNING

The solid-state tuning system is based upon the use of the "varactor" diode. This diode is also known as the voltage-variable capacitor. The varactor is a special solid-state diode (special doping) that acts as a capacitor whose capacitance varies inversely with the amount of reverse bias applied across the diode. All varactors operate only with a reverse bias. The tuned circuits to which the varactor is connected have their resonant frequency changed merely by changing the amount of reverse bias across the varactor. The greater the reverse bias across the varactor, the less the varactor capacitance and the higher the tuned-circuit resonant frequency.

Figure 8-28 is a simplified drawing of a varactor-controlled tuning circuit. In actual practice, the switch (SW) would have positions (and respective variable resistors) for each of the VHF channels. In order for the varactor to control the frequency of the tuned circuit, it is placed across variable capacitor C2. The amount of reverse voltage on the varactor is controlled by the switch position and the setting of the variable resistors. Capacitor C1 is an isolation capacitor used to prevent dc current flow through L.

In actual practice, it is not possible to get a varactor diode to provide a complete capacitor range for tuning all VHF or UHF channels. Therefore, it is common practice to switch in a different inductor for the upper VHF channels.

VIDEO IF AMPLIFIERS

Video IF amplifiers perform essentially the same functions that are performed by the IF amplifiers in superheterodyne radio receivers. However, bandpass considerations, vestigial (pertaining to a remnant or remaining part) sideband transmissions, the necessity for trapping unwanted sound and video beat frequencies, and other factors make the design of video IF amplifiers somewhat complex.

TRAPS

There are five types of traps used in conjunction with video IF amplifiers. These are series, parallel, absorption, degenerative, and bridged-T. For the most part, they are used in both vacuum-tube and solid-state circuits and for either monochrome or color TV reception. The bridged-T trap is the most widely used circuit and is indispensable in color TV receivers.

The series trap is a parallel resonant circuit configuration (fig. 8-29, view A). It is placed between two IF states and tuned to the frequency to be rejected. This type of trap circuit is a sharply tuned network designed to reject one frequency or, at most, a narrow band of frequencies. When a signal voltage at the trap frequency appears at the input of the circuit, the impedance offered by the LC network is very high, and almost all of the undesired voltage is dropped across the trap. A negligible amount of the voltage appears across the input circuit of the following IF amplifier. At all other frequencies, the resonant circuit offers negligible impedance, and the desired signals pass easily.

The parallel trap is a tuned circuit that is placed across, or in shunt with, the circuit. Figure 8-29, view B, shows a series-resonant circuit used in this manner. At the frequency for which it is set, the trap acts as a short circuit, bypassing the resonant frequency to ground and preventing further penetration into the circuit. At other frequencies, the trap circuit presents a relatively high impedance, permitting these signals to proceed to the following stage. It is important that the parallel trap have a very high Q so the circuit will bypass only a narrow range of frequencies.

The absorption trap (fig. 8-29, view C) is a widely used type of rejection circuit. It consists of a coil and a fixed capacitor inductively coupled to the loud inductor of an IF amplifier. When the IF amplifier receives a signal at the resonant frequency of the trap circuit, a high circulating current develops in the trap network as a result of the coupling between the trap and the load inductor. The voltage in the load coil L1 becomes quite low at the trap frequency. Consequently, very little of this interference voltage is permitted to reach the following stage. It is convenient to think of this kind of trap as being able to absorb all of the energy of the frequency to which it is tuned, and therefore no energy at that frequency is left available to pass on into the next stage.

Degenerative traps (fig. 8-29, views D and E) are designed to reduce the gain of an amplifier for frequencies to which the trap is tuned. These traps are

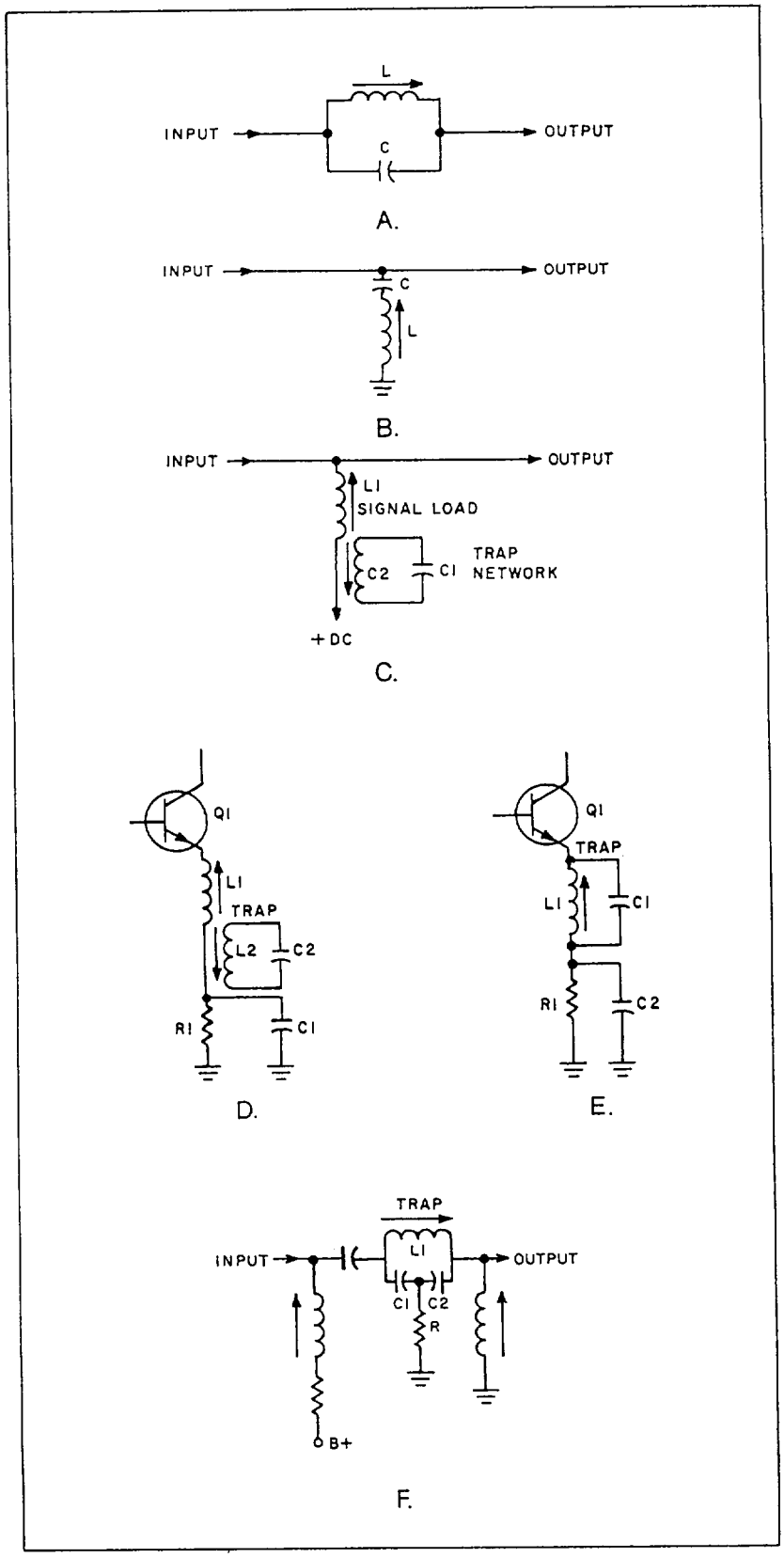


Figure 8-29.-Trap circuits.

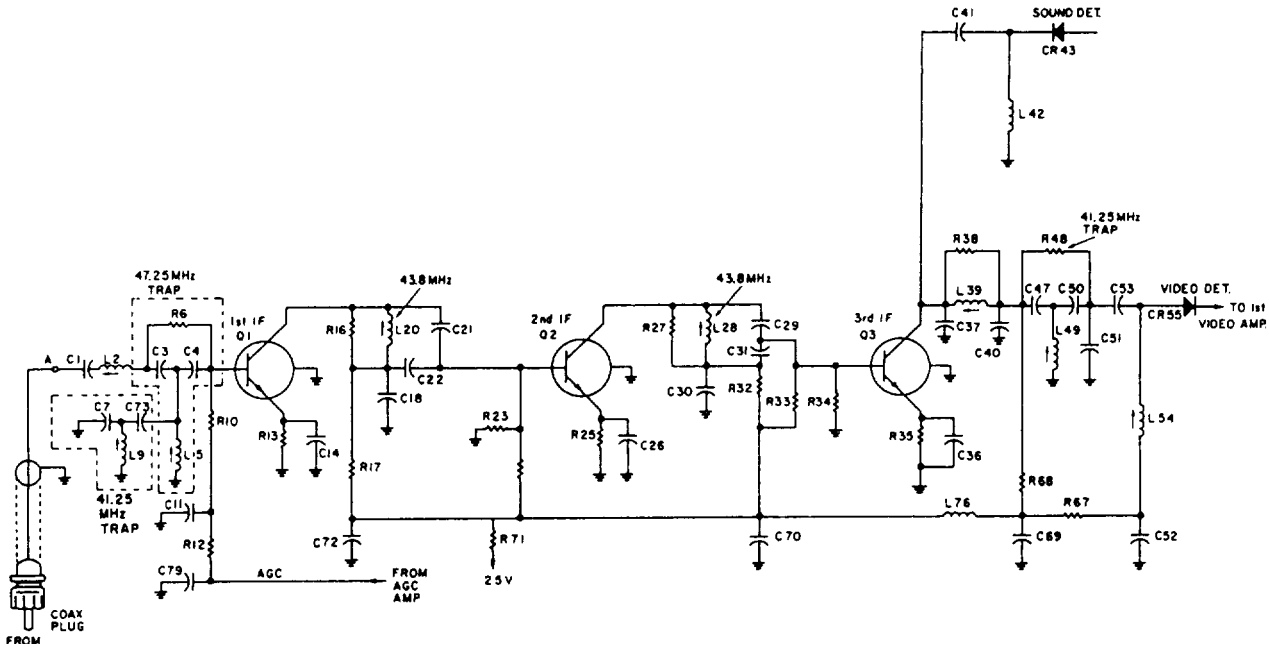


Figure 8-30.-Solid-state color IF amplifier.

used in the emitter circuit of a solid-state amplifier or in the cathode leg of a vacuum-tube circuit. In the latter application, the traps are often called "cathode traps." The two types of traps normally used to provide the degeneration are the absorption type and the series type. Figure 8-29, view D, shows an absorption type in which L1 in series with C1 forms a broadly tuned series-resonant circuit at the frequency to which the amplifier is tuned. This permits the amplifier to function normally for all signals within its frequency range. At the resonant frequency of the trap, however, a high impedance is reflected into the emitter or cathode circuit by the trap, and the gain of the stage is reduced by degeneration.

The series type of degenerative trap (fig. 8-29, view E) places a parallel circuit directly into the emitter or cathode leg. At the resonant frequency of the trap, the impedance in this part of the amplifier circuit will be high, producing a large degenerative voltage, and thus reducing the gain of the amplifier. At all other frequencies, the impedance of this parallel network is low. Only a small degenerative voltage appears, and therefore only a slight loss in gain occurs except at the undesired frequency.

A trap that is more complex than any of the foregoing circuits, but also more effective, is the bridged-T trap, shown in figure 8-29, view F. In this circuit, L1, C1, and C2 are resonated at the frequency of the signal to be rejected. If the resistance of R is properly chosen, the attenuation imposed upon a signal to which L1, C1, and C2 are resonated will be great. Ratios of 50 and 60 to 1 are easily attainable using standard components. This means the strength of the desired signal at the output of the trap will be 50 to 60 times greater than the strength of the undesired signal.

CIRCUITS

While reading this section, you should refer to figure 8-30, which is a schematic diagram of a typical video IF system found in many color TV sets. This system has three NPN transistor stages that are impedance coupled and pass a band of frequencies centered at 43.8 MHz. Input to the IF system is through a plug-in link from the mixer stage in the tuner. The mixer output coil is tuned to position 42.17 MHz at the 50 percent point on the IF response curve, and the impedance of the coil is tapped to match the impedance of the coax link. Q1, the first IF amplifier, is coupled to the input coax link through L2 and C1. L2 is tuned to position 45.75 MHz at the 50 percent point on the opposite side of the response curve from the 42.17 MHz point. Both 50 percent points on the IF response curve are shown in figure 8-31. The mutual coupling of L2 and the mixer output coil provide a wide bandpass through the coax link.

There are two sound traps at the input to Q1-one tuned to reject the adjacent channel sound frequency at 47.25 MHz and the other tuned to suppress the 41.25

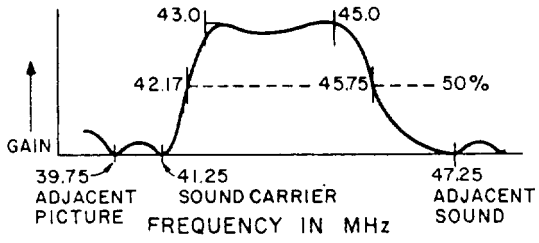


Figure 8-31.-Video IF response curve for a color set.

MHz associated sound carrier. The 47.25 MHz trap consists of R6, C3, C4, and L55. This trap is a bridged-T configuration and delivers, to the base of Q415, two 47.25 MHz voltages that are equal but 180 degrees out of phase. One of these voltages is developed across R6, while the other is developed across C3, C4, and L5. At the base of Q1 the voltages cancel each other, thus eliminating the adjacent channel sound carrier from the IF band. The 41.25 MHz trap is a series-resonant network consisting of C7, C73, and L9. This trap, loosely coupled to the input circuit, improves selectivity and provides better fine tuning.

In this system, AGC is applied only to the first IF amplifier stage. As the AGC voltage increases, forward bias is developed at the base of Q1. This bias increases the current flow through the transistor and, consequently, causes a greater voltage drop across R16, thus reducing the stage gain.

In the collector circuit of Q1, C21, and C22, divide the voltage developed across L20 and couple it to the base of Q25. The junction of the two capacitors matches the impedance of the coil to the base of Q2. L20 is tuned to 43.8 MHz, which is the center of the IF system bandpass. Coupling between Q2 and Q3 is identical to that between Q1 and Q2, with L28 also tuned to 43.8 MHz. The associated sound carrier is tapped off at the collector of Q3, the third IF amplifier.

The output circuit of the IF system is, in effect, very similar to the input circuit. Here, L395 is tuned to position 42.17 MHz at the 50-percent point on the IF response curve, as was indicated in figure 8-31, and L54 positions 47.75 MHz at the same point but on the opposite slope. A bridged-T trap, consisting of R48, C42, C50, and L49, rejects the 41.25 MHz associated sound carrier, thus preventing it from reaching the video detector. The three sound traps in this system, two at 41.25 MHz and one at 47.25 MHz, in addition to suppressing interference, account for most of the selectivity in the IF system.

VIDEO DETECTORS

The video detector in a TV receiver performs essentially the same function as the second detector in a superheterodyne amplitude-modulated radio receiver. It rectifies the signal (video and sync pulses) fed to it by the video IF system, removes the IF components, and feeds the remaining signal and sync information to the video amplifier. Various circuit arrangements, employing either diodes, electron tubes, or crystals, are used.

One type of diode detector is shown in figure 8-32. Two methods of connecting the diodes are shown. The output of CR1 has a negative picture phase, and CR2

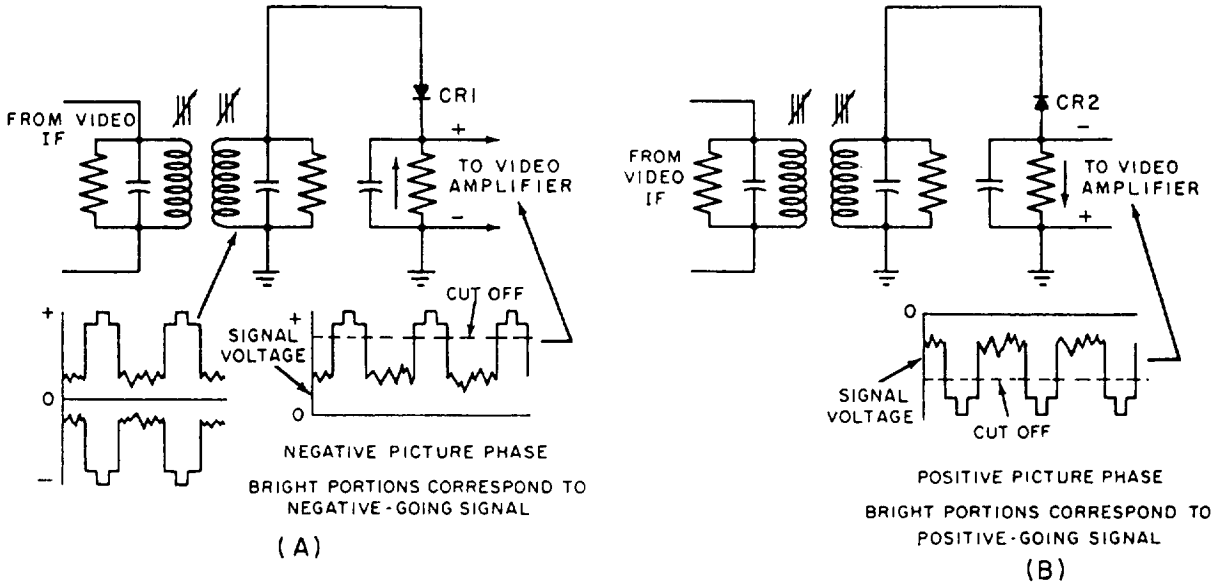


Figure 8-32.-Video diode detectors.

has a positive picture phase. In figure 8-32, view A, the low amplitude positive-going picture signals correspond to the brighter portions of the picture. The higher amplitudes correspond to progressively darker portions of the picture. This corresponds to the negative picture phase.

The low amplitude negative-going picture signals of CR2 correspond to the brighter portions of the picture. The higher amplitudes correspond to progressively darker portions of the picture. This corresponds to the positive picture phase shown in figure 8-32, view B.

The following paragraph will help clarify the concept of the negative picture phase. A video signal having a negative picture phase causes the cathode of the picture tube to be driven in a positive direction beyond cutoff during the blanking pulses. During the darker portions of the picture, the cathode is biased in the positive direction but not to cutoff. As a result, few electrons reach the picture tube screen, and the screen is relatively dark. During the brighter portions of the picture, the cathode is driven only slightly in the positive direction, and many electrons reach the picture tube screen. Therefore, the screen is relatively bright.

The concept of the positive picture phase is presented so that it will be better understood. A video signal having a positive picture phase causes the grid of the picture tube to be driven negative below cutoff during the blanking pulses. During the darker portions of the picture, the grid is driven considerably negative but not to cutoff. Consequently, fewer electrons reach the picture tube screen, and the screen is relatively dark. During the brighter portions of the picture, the grid is driven only slightly negative, and many electrons reach the picture tube screen. Therefore, the screen is relatively bright.

The output of the detector in the receiver may have a positive or a negative picture phase, as indicated in figure 8-32, irrespective of the type of transmission used. In either case, the video amplifiers must supply a signal having a positive picture phase to the grid of the picture tube. If the signal is applied to the cathode of the picture tube, it must have a negative phase.

Look at figure 8-25 for a review of signal processing. Up to this point, you have seen that the TV signal has been received and amplified by an RF stage, converted to another frequency (IF) by means of a mixer, further amplified by the IF stages, and rectified by the diode detector. The strength of the signal at the output of the detector is not sufficient to drive the picture tube; therefore, one or more stages of video amplification are necessary.

VIDEO AMPLIFIERS

After the video signal (containing the video information and the blanking and sync pulses) has been rectified in the second detector, it must be amplified in one or more video amplifiers before it is applied to the picture tube. Because a wide band of frequencies must be passed without discrimination by the video amplifiers, they must be carefully designed. Special high- and low-frequency compensating circuits must be used to extend the approximate range of frequencies passed from 30 hertz to 4 MHz. In other words, frequency distortion must be eliminated as much as possible.

The low-frequency video components include the low-frequency ac variations (represented on the picture screen as portions of the image that does not contain fine detail), the blanking pulses, and the sync pulses (vertical and horizontal). The high-frequency video components are the high-frequency ac variations that produce the fine detail on the picture screen. In addition to the low-frequency and the high-frequency components in the transmitted signal, there is a zero frequency, or dc component present.

If all of the frequency components are not properly amplified in the video amplifier section, a distorted image is produced. The distortion may appear as a lack of fine detail, a lack of image sharpness in the larger objects, or a lack of contrast.

In addition to frequency distortion, phase distortion must also be eliminated as much as possible. Phase distortion means that certain components (frequencies) that make up complex waveforms are not passed by the amplifier in the same length of time that other frequencies are passed. For example, in resistance-coupled amplifiers, the coupling capacitor and the grid resistor, acting together, cause a phase shift that varies with the frequency. Phase distortion may alter the background of the picture shown on the TV screen; portions that should be white maybe gray or even black. At the lower frequencies, excessive phase shift may cause larger objects to be blurred on the screen. At the higher frequencies, excessive phase shift causes the fine detail to be blurred.

The average resistance-coupled amplifier has a flat response of only a few thousand hertz; therefore, it is not suitable for amplifying video frequencies. In order to amplify the higher frequencies as much as the middle

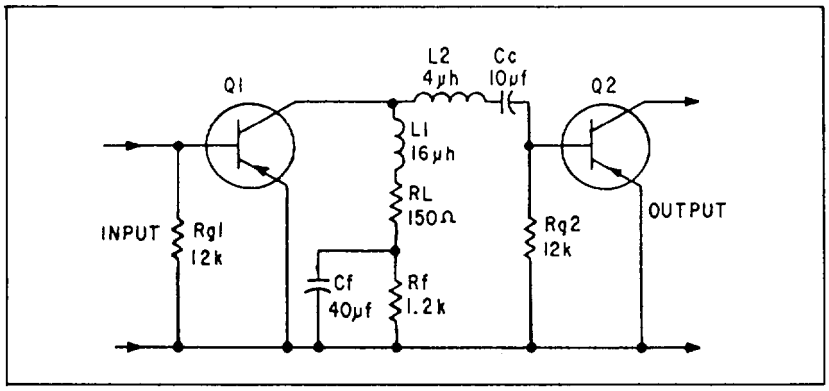


Figure 8-33.-Transistor video amplifier.

range of frequencies, it is necessary to use some form of high frequency compensation. This type of compensation commonly takes the form of shunt compensation, series compensation, or a combination of both. An example of the combination type of compensation is shown in figure 8-33.

In a resistance-coupled amplifier, the output capacitance of the first amplifier stage, the distributed capacitance of the wiring, and the input capacitance of the second amplifier stage tend to shunt the high frequencies of the first stage to ground. Therefore, there is less output voltage at these frequencies available to the second stage. The high frequencies are, therefore, not amplified as much as the middle range of frequencies.

The shunting effect is compensated for (or the frequency range extended) by the use of a small inductor (shown as L1 in figure 8-33) inserted in series with the load. The value of this inductor is chosen so that it will neutralize the distributed (output and input) capacitance of the circuit. That is, this inductor, together with the distributed capacitance, forms a parallel-resonant circuit that is resonant at a frequency where the response curve begins to fall appreciably. The frequency range is thereby extended. This type of compensation is called shunt compensation, and the coils are called peaking coils.

Series compensation is also used. In this case, a small inductor (shown as L2 in figure 8-33) is added in series with the coupling capacitor and forms a series-resonant circuit (resonant at a frequency where the response curve begins to drop) with the distributed capacitances. At resonance, increased current flows through these capacitors, and larger voltages are available at the input of Q2. To prevent a sharp peak in the response curve, a resistor is shunted across the series inductor.

To amplify the lower frequencies as much as the middle or high range of frequencies, it is necessary to use some form of low frequency compensation. At low frequencies, the reactance of the coupling capacitor is large, and much of the signal voltage is dropped there and not available at the base input. If a large coupling capacitor could be used, the stray capacitance and leakage current would be increased. Of course, a coupling capacitor introduces some phase shift at low frequencies.

It is possible to compensate for the loss in gain and the increase in phase shift at the lower frequencies by dividing the load resistor into two parts and bypassing one part with a capacitor. Look at figure 8-33. The load resistance is divided into two parts by using RL and Rf. One part is bypassed by using Cf. At the lower frequencies, the load includes both resistors; therefore, the output voltage is higher. At the higher frequencies, a portion of the load is effectively bypassed by the capacitor, and the output is proportionately lower.

DC RESTORERS

The dc restorer (or clamper) restores the dc component of a pulse waveform after this component has been removed by the passage of the waveform through the coupling capacitor in the video amplifier stage. It is necessary to reinsert the correct dc component at the input of the TV picture tube if the correct level of background illumination is to be maintained. Also, if the correct dc component is not reinserted, the blanking level will vary (instead of remaining constant as it should), and retrace lines will appear on the screen during the time the blanking voltage is insufficient to cut off the picture tube during retrace.

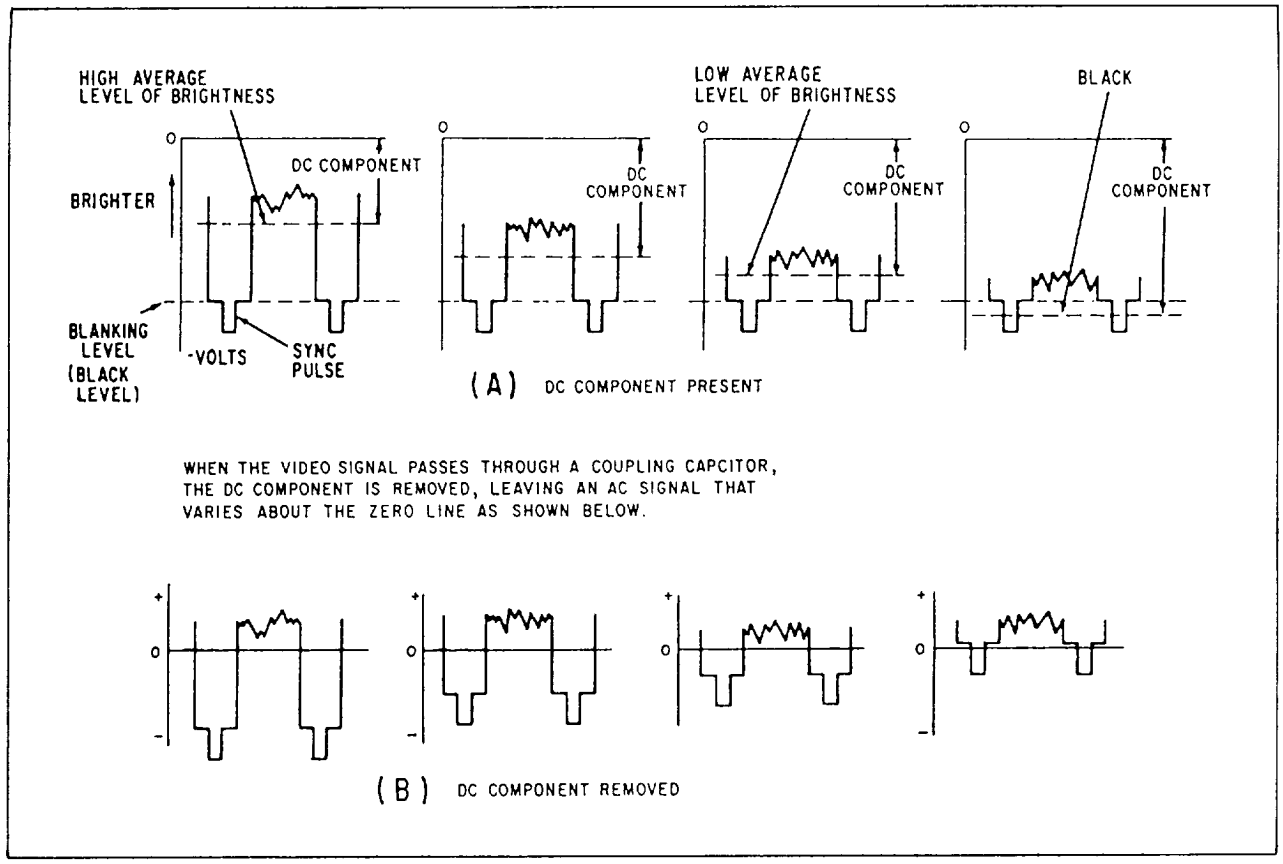


Figure 8-34.-Function of the dc component in the video signal.

The average brightness of one seamed line may differ widely from the average brightness of another scanned line, as shown in figure 8-34, view A. The average dc component depends on the average brightness of a scanned line. A low dc component in the negative direction means a high level of brightness exists during that line. A high dc component in the negative direction means a low level of brightness exists during that line. Therefore, the average dc component establishes the blanking level.

Figure 8-34, view B, illustrates what happens when the dc component is removed by the passage of the video signal through a coupling capacitor. Although the picture tube may be biased so that it is not driven in a positive direction beyond a certain value, nevertheless the blanking level varies and the retrace is often visible. The background brightness level also differs from that at the transmitter.

A simplified dc restorer circuit is shown in figure 8-35. The function of the circuit is to restore the dc component that was lost when the video signal passed through the coupling capacitor of the video amplifier. DC restoration is accomplished by adding enough dc voltage to the instantaneous ac signal to bring the blanking voltage to the cutoff point.

The dc diode restorer acts in the following manner. Without the diode, the input signal voltage appears as shown in figure 8-34, view B. During the negative portion of the cycle (when the blanking and synchronizing pulses are active), the diode (fig. 8-35) conducts because its cathode is negative and its plate is positive. Capacitor C charges rapidly through the diode

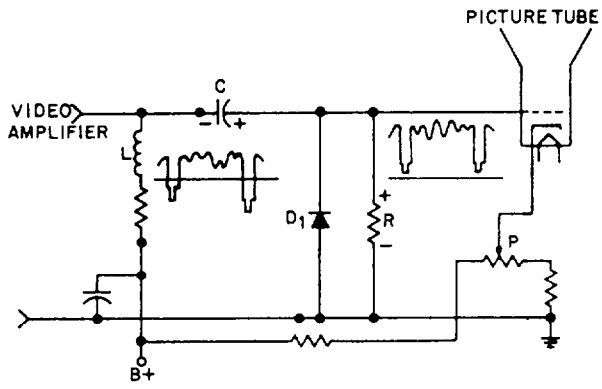


Figure 8-35.-Diode dc restorer.

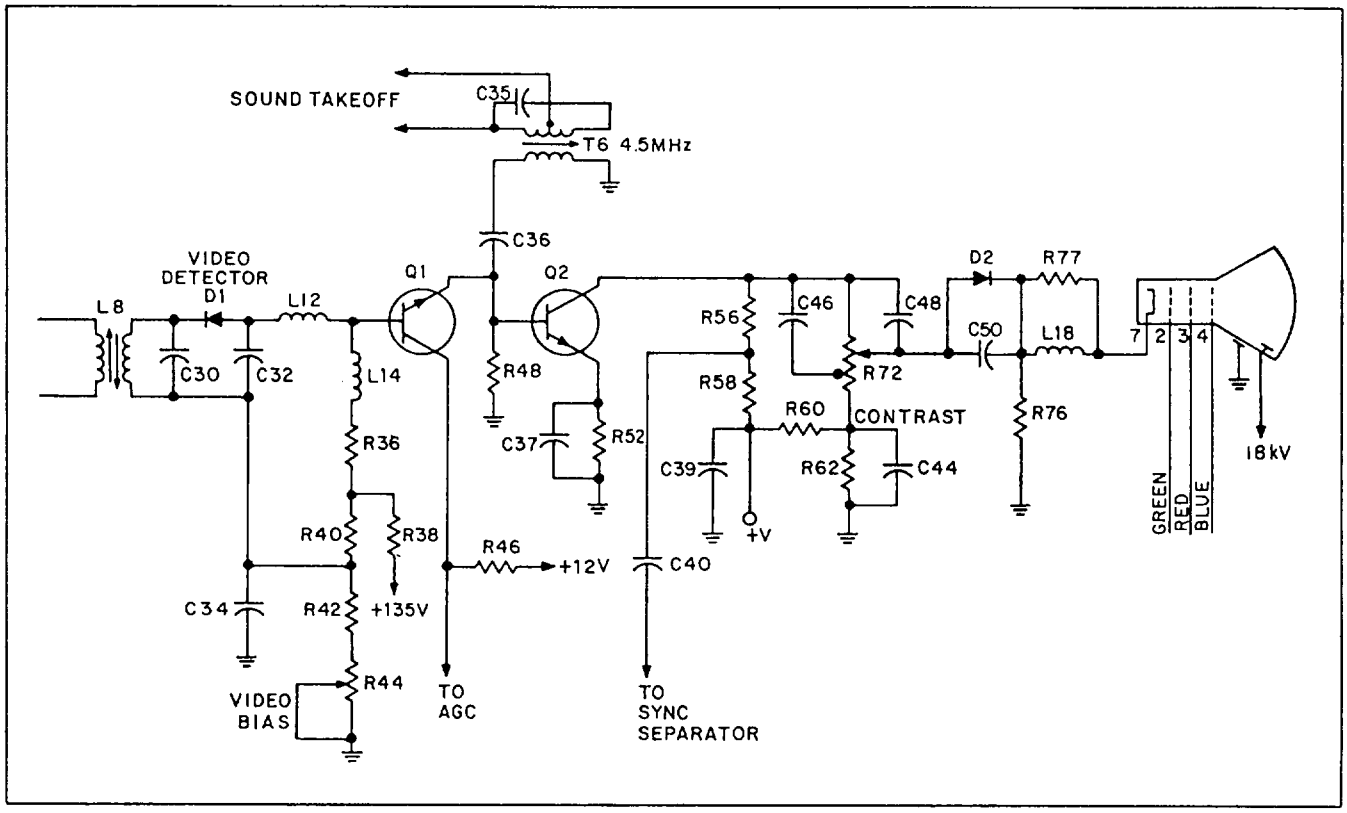


Figure 8-36.-Direct-coupled video amplifier stages.

and has the polarity indicated. The amount of the charge (voltage across C) depends on the strength of the input signal.

During the positive portion of the signal, shown above the 0 line in figure 8-34, view B, the diode cannot conduct, and C (as shown in figure 8-35) discharges slowly through R. A positive potential, which reduces the bias on the picture tube, is applied between grid and cathode during the scan interval between the blanking pulses. During this interval, the diode is effectively an open circuit, and the video signal appears across R in series with the dc voltage supplied by C. The greater the input voltage, the less the net bias remaining on the grid of the picture tube and the higher the average brightness. Thus, the condition existing in figure 8-34, view A, is reestablished.

When direct coupling is employed between all stages in the video amplifier section, including the coupling stage between the video detector and the first video amplifier and picture tube, dc restoration is not required. An example of this direct coupling is shown in figure 8-36. Detector D1 is direct coupled to first video amplifier Q1 through peaking coil L12. This is an emitter-follower stage, which develops a signal across R48. The base of second video amplifier Q2 receives its signal directly from the emitter of Q1. The second amplifier is a conventional (common emitter) amplifier stage direct coupled to the picture tube through the contrast control, diode D2, and peaking coil L18. Diode D2 is used to compensate for nonlinearity in transistor amplifier Q2 on high-amplitude signals.

TELEVISION SOUND SYSTEMS

Learning Objectives: *Describe the sound systems of a TV receiver to include the split-carrier sound system intercarrier-sound system, and integrated-circuit (IC) sound system.*

The sound system of a TV receiver is essentially the same as that of an FM receiver. An important difference, however, is the system used for IF amplification. There are two systems in use–the split-carrier sound system and the intercarrier-sound system.

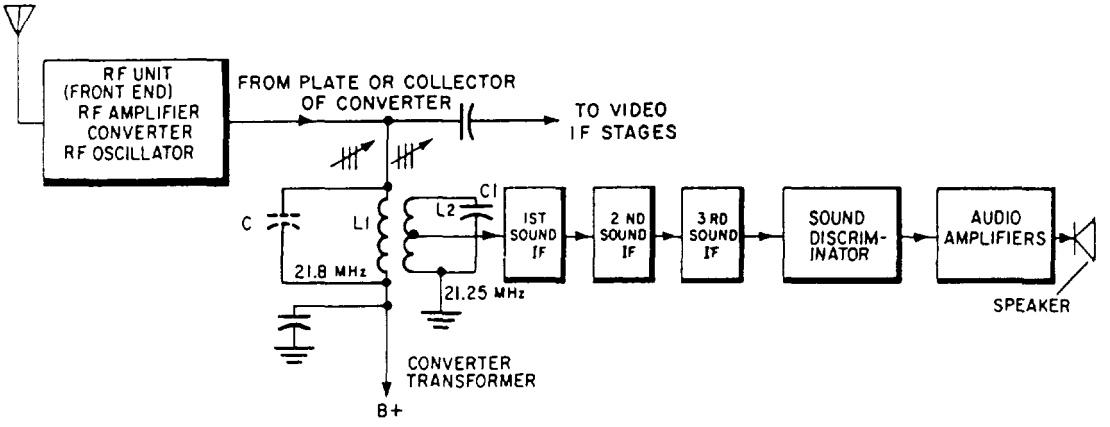


Figure 8-37.-Spilt-carrier sound system.

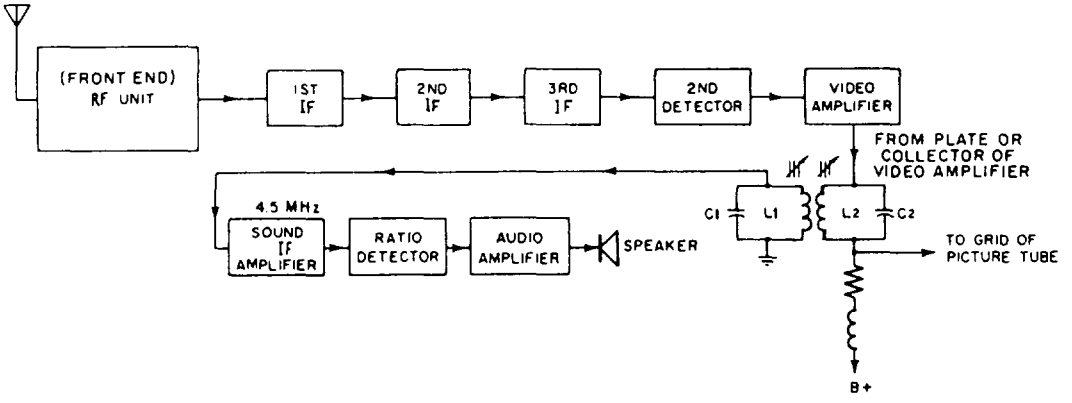


Figure 8-38.-Intercarrier sound system.

SPLIT-CARRIER SOUND SYSTEM

A block diagram of a split-carrier sound system is shown in figure 8-37. Although a large number of the newer TV receivers use intercarrier sound, some sets still employ the split-sound system. In the split-sound system, the sound IF is removed at the output of the converter (or from the first video IF) and amplified in a series of sound IF amplifiers.

INTERCARRIER-SOUND SYSTEM

A block diagram of an intercarrier-sound system is shown in figure 8-38. Its obvious advantage is that the sound is amplified along with the video, which means fewer audio stages are necessary.

The carrier IF in the output of the RF unit is the same as those in the output of the RF unit in the split-sound system. Because the IF stages in the intercarrier system must pass both the picture and the sound IF carrier, the bandpass must be wide enough to pass both of these frequencies.

INTEGRATED-CIRCUIT (IC) SOUND SYSTEM

An ideal IC for a TV sound section would have five terminals: one for ground, one for power, a 4.5 MHz input, a lead for the volume control, and a connection that goes directly to the speaker. Figure 8-39 shows a TV FM sound circuit, which contains an IC and 13 other components. The two coils are single-slug tuned and simple to align.

The power transistor and output transformer are required because ICs are usually small devices with low-power dissipation. Some additional gain is therefore required to change to the power and impedance requirements of a 4-ohm, 2-watt speaker.

The two IF amplifiers are not transformer or resonant-circuit coupled. Instead, adequate selectivity is obtained with just the resonance of L1 and C2. A gain of 2,000 at 4.5 MHz is typical. This large gain saturates the last IF amplifier and produces the square-wave signal required for the quadrature detector.

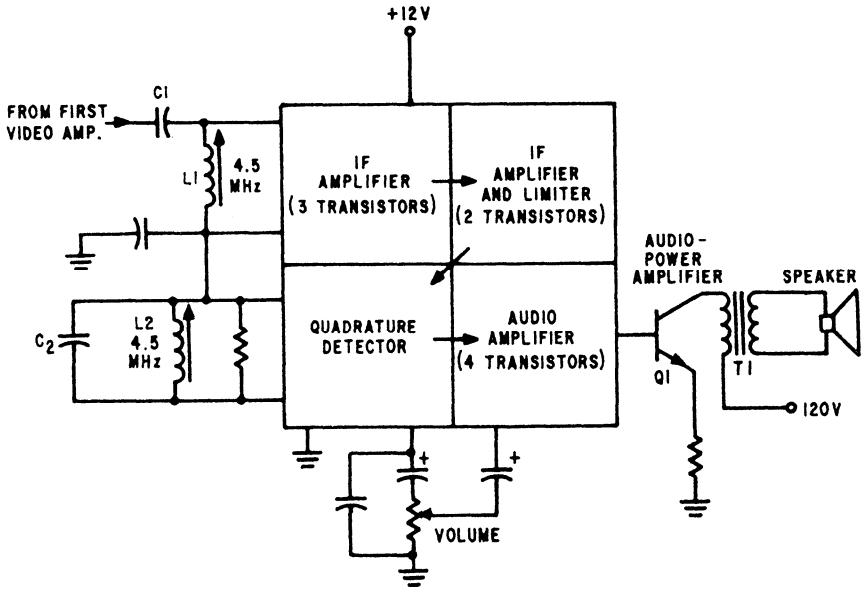


Figure 8-39.—IC sound system.

The quadrature detectors an FM detector circuit that acts as both a discriminator and a limiter in one stage. This type of FM detector is easily produced as part of an integrated circuit. Tuning merely requires making L2 and C2 resonant at 4.5 MHz.

The audio-voltage amplifier is a three-stage, four-transistor, de-coupled amplifier. Direct coupling allows a good bass response and also avoids the use of capacitors. Note that, except for the speaker transformer and the two capacitors associated with the volume control, this system would have a bass response down to dc.

PICTURE TUBES

Learning Objective: *Recall the purpose and function of a monochrome picture tube.*

A monochrome picture tube is a specialized form of the cathode-ray tube. An electron gun in the tube directs abeam of electrons toward a fluorescent material on the screen, which glows when struck by the electrons. Between the gun and the screen are deflection coils that deflect the beam horizontally and vertically to form a raster. The brightness of the screen at any point depends upon the number and velocity of electrons striking that point. The brightness of the picture is controlled by varying the grid-bias voltage with respect to the cathode voltage. This bias can be changed by varying either the cathode voltage or the grid voltage.

Color picture tubes operate on the same basic principle as monochrome picture tubes. The difference between the two systems is the types of phosphors that coat the screen. The different types of phosphors produce colors when struck by electron beams. Three basic or primary colors are used in combination to produce all the other desired colors. These primary colors are red, green, and blue. In a three-gun color picture tube (fig. 8-40), there is a separate gun for each of the color phosphors. The tube's screen consists of small, closely spaced phosphor dots of red, green, and blue. The dots are arranged so a red, green, and blue dot form a small triangle. The shadow mask provides a centering hole in the middle of the triangle of dots. The convergence electrode causes the three separate electron beams to meet and cross at the hole in the shadow mask.

Each electron gun is electrostatically focused by a common grid voltage. In other words, each gun has its own electrode, but all three are connected together

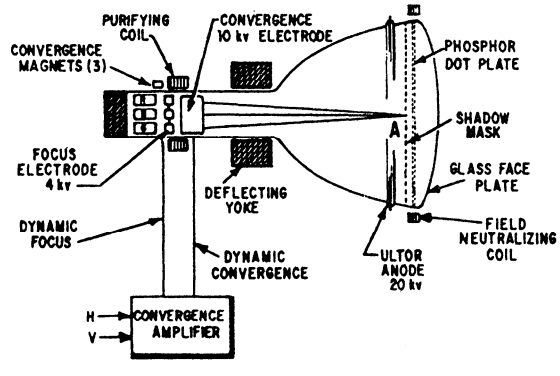


Figure 8-40.—Color picture tube.

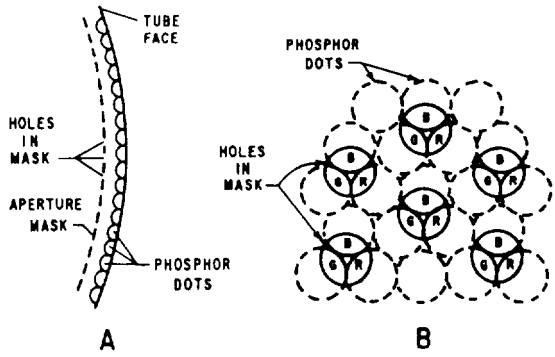


Figure 8-41-Color picture tube. (A) Side view; (B) front view.

requiring only one grid voltage. The three electron beams scan the screen as controlled by the deflection yoke, which is mounted externally around the neck of the tube. As the three beams scan the phosphor screen horizontally and vertically in the standard scanning pattern, the dot triads will light according to the video input signals (fig. 8-41).

The purifying coil produces a magnetic field within the tube, which aligns the electron beams parallel to the neck of the tube. Rotating the purifying coil adjusts the electron beams so they strike their respective color dots without striking neighboring dots. When this adjustment is made for the red dots, the other two electron beams are aligned as well.

The high voltage anode is a metallic ring around the tube. The field neutralizing coil aids color purity at the outer edges of the picture tube. A metal shield, called a mu-metal shield, is placed around the bell of the tube to prevent stray magnetic fields from affecting the electron beams.

COLOR CIRCUITS

Learning Objective: *Identify color TV circuits and the difference between color and monochrome receivers.*

A color TV receiver (fig. 8-42) contains many circuits that are markedly different from the circuits used in monochrome receivers. The differences are outlined in the following paragraphs.

The tuner and amplifier stages in color receivers are designed to pass a wider band of frequencies than do conventional monochrome receivers. Wideband characteristics in these stages are necessary to assure uniform amplification of the high-frequency color subcarrier sidebands that carry the chrominance (color) information.

Usually, the video amplifier stage in monochrome receivers consists of only one stage of amplification, while color receivers usually contain three stages of amplification. Additional stages of amplification are made necessary because the luminance signal in color receivers is used to drive the cathodes of all three electron guns in the CRT, as compared to the single-gun monochrome CRT.

A video delay line, usually located between the video output stage and the CRT, is used to delay the luminance signal for a fixed period of time so the luminance and chrominance information arrive at the CRT simultaneously. As shown in the block diagram (fig. 8-42), this fixed delay is necessary because the chrominance signals pass through additional stages before being applied to the control grids of the CRT. Were it not for the delay of luminance information, the two signals would not arrive in coincidence, and a distorted video presentation would result.

Because of the use of an aperture mask type of picture tube, the brightness of a color receiver is characteristically low. This means that higher voltage is necessary to maintain adequate brightness. The output voltage of the high voltage supply is nominally 20-25 kV as compared with 15-18 kV for monochrome receivers. All three electron guns must be sharply focused onto the screen to obtain good monochrome and color reproduction. The focus rectifier in color receivers provides a variable focus voltage (4 to 5 kV) that is applied to the electrostatic focus elements of the CRT. The load of the high-voltage rectifier must be held fairly constant in color receivers. Otherwise, severe blooming or shrinking of picture size will occur during reception of signals having a varying brightness level. The voltage regulator circuit provides a fairly constant anode voltage regardless of the brightness level of incoming signals.

The color demodulator section is the “heart” of the color TV receiver. In this section, the 3.58 MHz subcarrier sidebands are demodulated to produce color information signals. The color information signals are then applied to a matrix where color difference signals are produced by matrixing proportionate amounts of the demodulated signals. The color difference signals are, in turn, amplified and applied to the control grids of the CRT in the proper proportions to reproduce the televised scene.

The color convergence circuits provide a secondary control over the electron beam of each gun. Convergence of the three electron beams to exact locations on the face of the CRT is necessary to produce

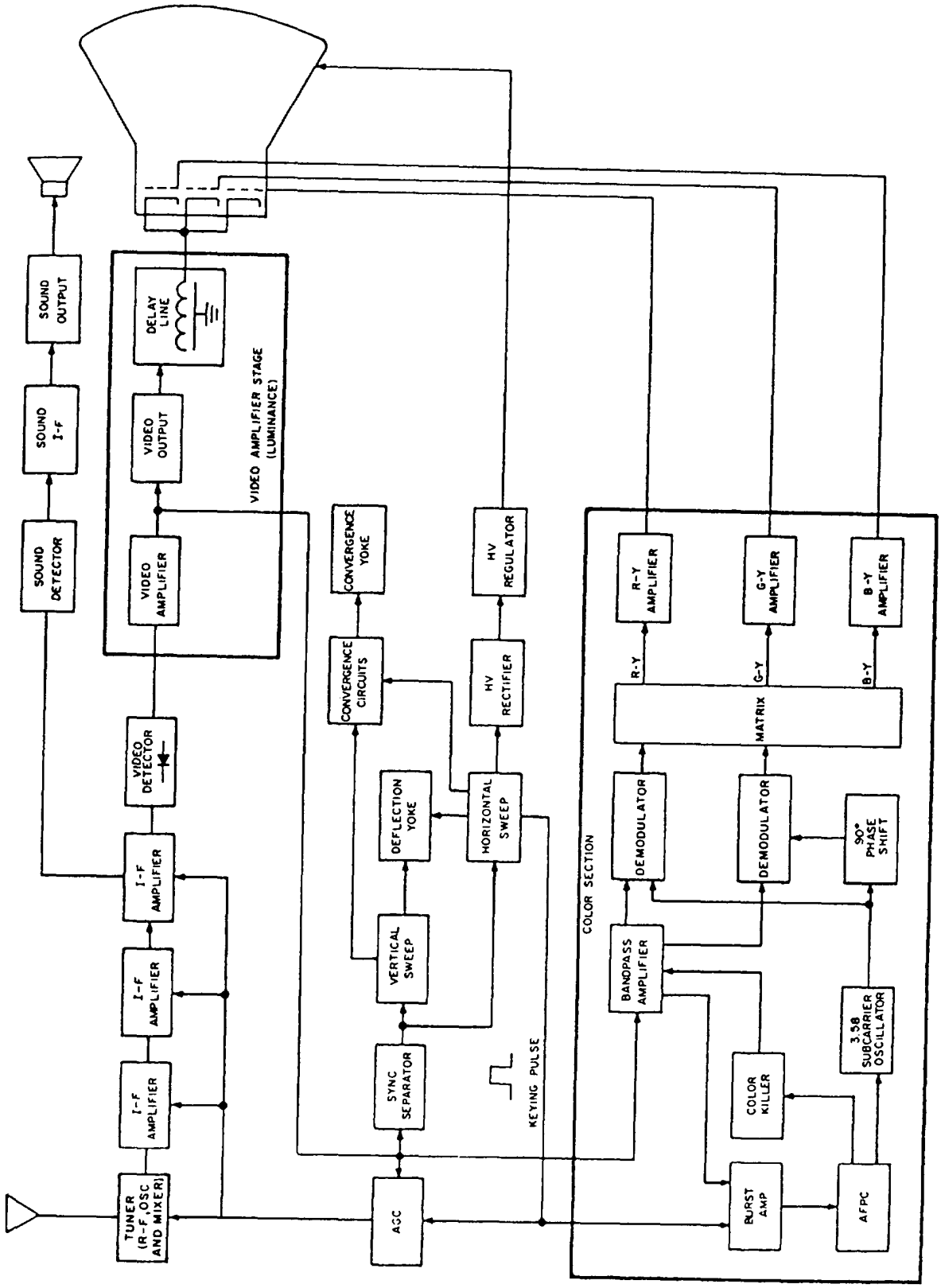


Figure 8-42.–Color television receiver, block diagram.

good monochrome and color images on the three-gun CRT.

Other differences in color receiver circuits, such as automatic color control, tuning indicators, and color reception indicators, serve to simplify the operation of front-panel controls.

BANDPASS AMPLIFIER

The chrominance 3.58 MHz subcarrier signal and the color-synchronizing burst signal are coupled from the first video amplifier to a bandpass amplifier (fig. 8-42). The bandpass amplifier is tuned to amplify the subcarrier frequency and reject the lower-frequency video signals. The 3.58 MHz subcarrier sideband signal must now be demodulated to recover the color information. Consequently, a locally generated 3.58 MHz oscillator signal must be mixed with the subcarrier sideband to produce the chrominance signals. The frequency and phase of the locally generated signal is critical. It must be nearly identical to that of the station transmitter in order to recover exact color information. Therefore, the local oscillator is constantly synchronized by the burst reference signal. At the output of the bandpass amplifier, the burst signal, which is composed of approximately 10 cycles of the 3.58 MHz signal generated by the transmitting station's oscillator (fig. 8-43), is selected by a burst separator. The burst separator in this instance is a time-gated amplifier that is gated, or "keyed" on, by a horizontal pulse from the flyback transformer during the horizontal retrace time. The burst signal is then applied to the automatic frequency and phase control (AFPC) circuits, which control the 3.58 MHz local oscillator.

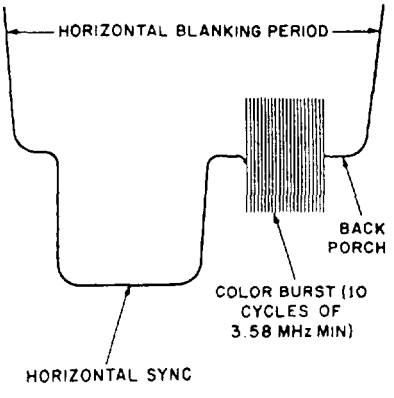


Figure 8-43.-Location of color burst on horizontal sync pulse.

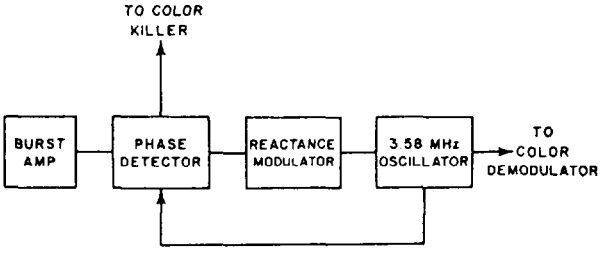


Figure 8-44.-AFPC circuit.

AUTOMATIC FREQUENCY AND PHASE CONTROL (AFPC) CIRCUIT

The burst signal is coupled to the AFPC circuit, as shown in figure 8-42. The reference circuit is usually one of two types: the stable oscillator-phase detection type or the frequency-phase detection supplemented type. Only the former will be discussed. The stable oscillator-phase detection type of AFPC, shown in figure 8-44, accepts the keyed signal from the burst gate and applies it to a phase detector. In the phase detector, the locally generated oscillator signal is compared with the incoming burst signal and an error or corrective voltage is developed. The error voltage is applied to a reactance modulator, which is connected in parallel across the 3.58 MHz oscillator tank. The reactance modulator acts as a variable capacitor, with its capacitance either increasing or decreasing, depending upon the polarity of the voltage applied. Thus, when the phase of the local oscillator changes appreciably, the voltage developed by the phase detector changes the capacitance of the reactance modulator, and the oscillator output signal shifts in phase to correspond to the burst signal.

The frequency of the local oscillator is primarily controlled by an input crystal tank circuit. However, when the local oscillator does drift off frequency, the phase detector error or correction voltage causes the oscillator to swing through its frequency range until the phase and frequency error voltages equal zero, and the oscillator locks in.

COLOR-KILLER CIRCUIT

The purpose of the color-killer circuit is to disable the color circuits when a monochrome signal is being received. Usually this is accomplished by biasing the bandpass amplifier below cutoff. No signal passes until the burst is applied to the color-killer circuit. In order to demonstrate the effectiveness of a color-killer circuit, assume that a signal of high noise characteristics is

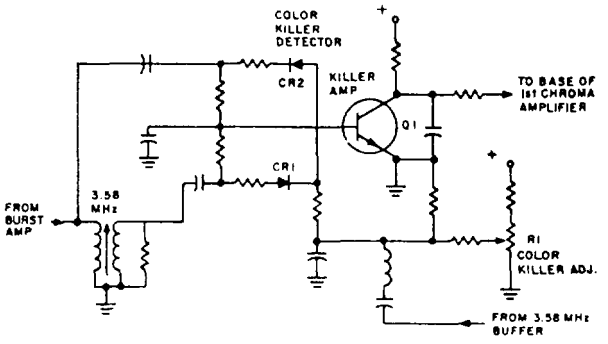


Figure 8-45.-Color-killer circuit.

present at the input of the bandpass amplifier. The transistor goes into conduction, which enables the color circuits and results in a colored noise presentation on the CRT. The color-killer circuit prevents this noise presentation on the CRT by sensing the absence of the color burst signal. If the burst is absent, the color-killer circuit conducts and applies bias to the bandpass amplifier, thereby cutting off the color circuits. When the color-killer circuit senses the presence of the burst signal, such as by sensing a negative voltage at the output of the phase detector shown in figure 8-44, the color-killer circuit ceases to conduct, and thereby permits the bandpass amplifier to admit the chrominance information.

In later model receivers, various circuits have been devised to eliminate this undesirable "color" during monochrome transmission. One simple method involves merely biasing off the bandpass amplifier by adjusting the familiar "color control" on the front of the receiver. However, a second and more sophisticated method of color-killer circuit operation is shown in figure 8-45. This diode phase detector circuit uses the 3.58 MHz burst signal and the 3.58 MHz signal from the crystal oscillator as inputs. It compares the phase of these two signals and generates a dc output proportional to their phase difference.

Note the location of the color-killer potentiometer (R1). It is adjusted so the Q1 is biased up to a point just slightly below cutoff. If the bursts are not present in the video signal, then a large phase error is detected. This causes a positive bias to add to the bias mentioned above, and Q1 turns on. This sends a negative bias to the chroma amplifiers, turning them completely off. If color bursts are present, the color-killer detector does not turn Q1 on, thereby allowing the chroma amplifiers to operate normally.

CHROMA DETECTORS

The purpose of the chroma detector circuit is to recover the chrominance components from the 3.58 MHz subcarrier sideband. These are transmitted by the station transmitter. The chrominance components are then used by the receiver circuits to reproduce a replica of the televised scene of the face of the CRT.

MATRIX AND COLOR DIFFERENCE AMPLIFIER CIRCUITS

Matrix circuits are designed to reassemble the chrominance signals to form the original camera output signals. The signals corresponding to the camera video signals are combined with the luminance signal to produce a replica of the televised scene on the face of the CRT. Matrix circuits take on many forms, and the particular circuit used in any one individual receiver depends upon the receiver design. At the transmitter, the luminance signal (E_Y) and chrominance signals (E_I and E_Q) were formed by combining proportionate parts of the red, green, and blue camera tube outputs. It would seem, then, that to reverse the process a similar matrix should be used at the receiver. At the receiver, however, the chrominance signals are demodulated to form color difference signals of R-Y and B-Y, rather than the transmitted chrominance signals of E_I and E_Q. (The reason is based on the economic advantages of using equiband circuits.) As a general rule, the luminance (E_Y) signal is combined with the color difference signals in the color CRT, however, this again is determined by receiver design.

SYNCHRONIZING CIRCUITS

The previous discussion on TV was focused primarily on delivering the video signal to the picture tube. Equally important, from the point of view of overall receiver operation, is the system by which the various circuits are synchronized with those at the transmitter and made to function together to produce the desired picture. Refer to figure 8-42 for an overview of the system. As previously stated, the blanking pulses and the horizontal and vertical synchronizing (sync) pulses are amplified in the various stages along with the video information. The sync separator, automatic gun control, and sweep circuits are virtually the same in color receivers as in monochrome receivers.

The detected composite video signal (containing synchronization pulses, blanking pulses, and video) is applied to the control grid or cathode (depending on phase) of the picture tube. The video information

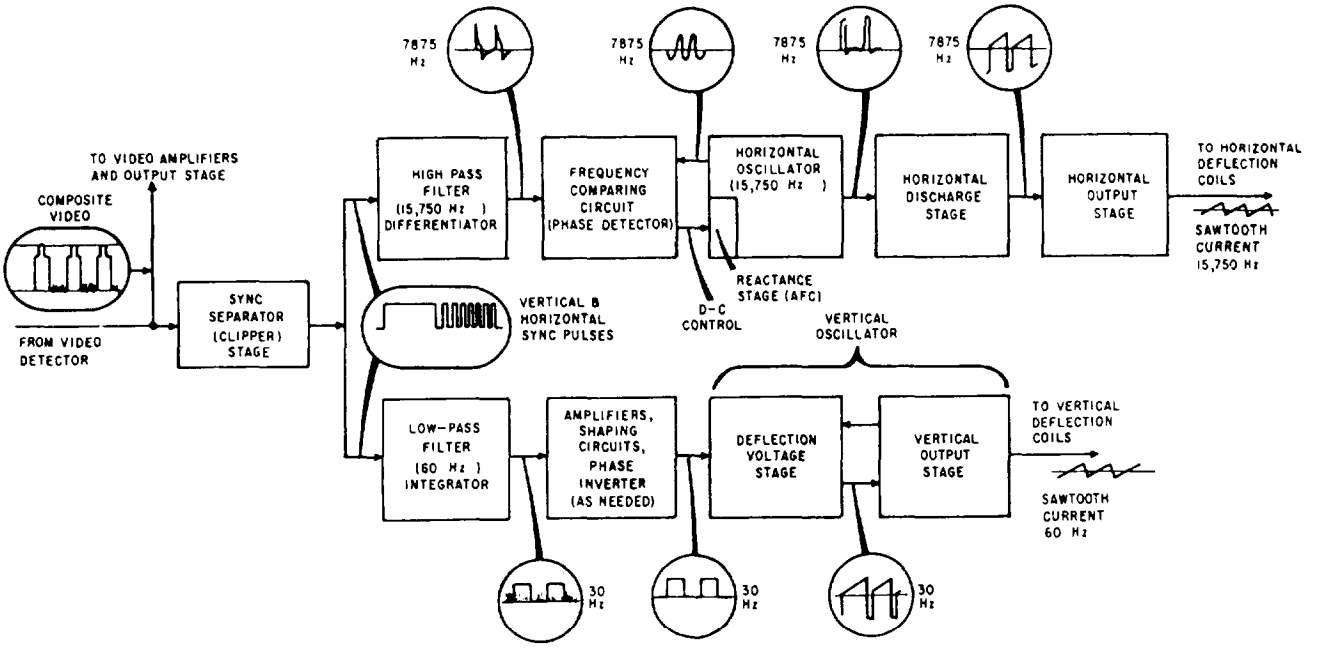


Figure 8-46.-Block diagram of synchronizing circuits.

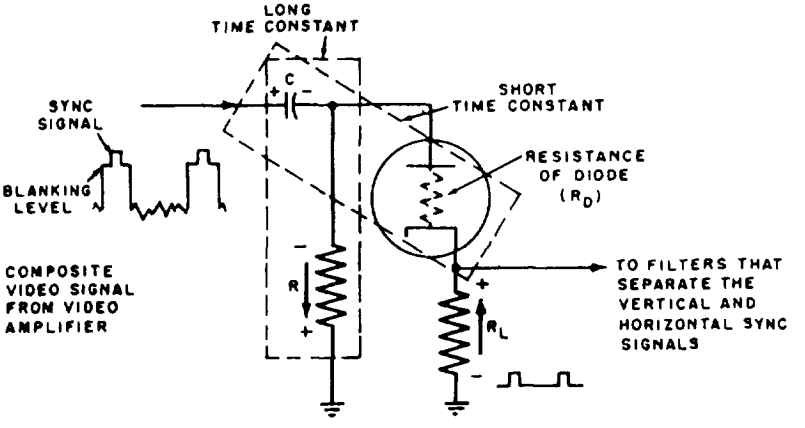


Figure8-47.-Diode sync separator.

intensity modulates the scanning electron beam, producing varying degrees of black through white information on the screen. The blanking pulses cut the picture tube off to prevent a visual indication of retrace. The sync pulses are present but have no effect since they are present during the time the picture tube is cutoff and only drive the tube beyond cutoff.

After detection and amplification, the composite video signal is also fed to a sync separation or clipper stage. The vertical and horizontal sync signals are removed from the composite video signal and filtered. Following this, the pulses are amplified and reshaped according to the needs of the synchronization and sweep systems. A block diagram of the synchronization circuits is shown in figure 8-46 with the associated pulse waveforms.

The horizontal sync signal fires the horizontal oscillator at exactly the right instant to maintain the proper synchronization between the horizontal sweep in the receiver picture tube and the horizontal sweep in the transmitter camera tube. The output of the horizontal oscillator is formed into a sawtooth waveform. It is then amplified and applied to the horizontal deflection coils.

The vertical sync signal fires the vertical oscillator at the right instant to maintain the proper synchronization between the vertical sweep in the receiver picture tube and the vertical sweep in the transmitter camera tube. As in the case of the horizontal

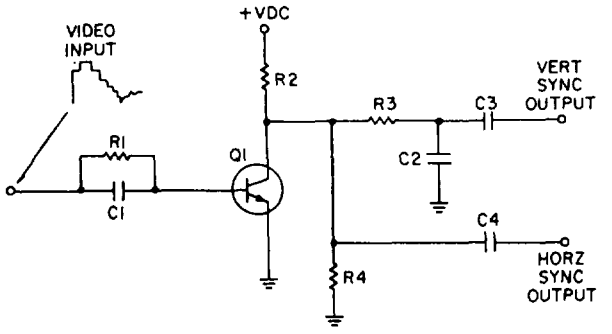


Figure 8-48.-Transistor sync separator.

oscillator output, the vertical oscillator output is formed into a sawtooth wave (modified into a trapezoidal form). It is then amplified and applied to the vertical deflection coil.

SYNC SEPARATORS

Sync separation, or sync clipping, may be accomplished by the use of circuits employing tubes or transistors. A simplified circuit of a diode sync separator is shown in figure 8-47. During the time the sync voltage is applied to the input, the diode plate is positive with respect to the cathode, and capacitor C is charged through the low resistance of the conducting diode. Between pulses, capacitor C discharges through R, and thus maintains a negative bias between plate and ground. This cuts off all signals up to the blanking level. The bias is maintained at approximately the blanking level, and only the sync signal causes pulses of current to flow through R1, across which the output is taken.

The output pulses consist of the horizontal sync pulses, the equalizing pulses, and the serrated vertical sync pulses. These pulses are fed to filter circuits that separate the vertical sync pulses from the horizontal sync pulses.

The simple diode separator is easily converted into a transistor sync separator. This is possible because the transistor-input terminals (base-emitter terminals) are actually a diode. The transistor merely adds gain to the basic diode sync separator. Figure 8-48 shows a simplified transistor sync separator.

The problems involved in removing (sync clipping) the sync signals from the composite signal and in amplifying, separating, and using those signals to control the horizontal and vertical oscillators are treated only in a general way in this section. There are many methods of solving these problems, and therefore a detailed treatment of each method is not possible in this training manual.

Because the repetition rate of the vertical sync pulses is 60 pulses per second and that of the horizontal sync pulses is 15,750 pulses per second, they can be separated by filters. One filter, the high-pass filter, is used to pass and shape the trigger voltages for the horizontal oscillator (multivibrator or blocking oscillator), as shown in figure 8-49. The circuit in this figure has a short time constant with respect to the period (width) of the horizontal pulse. The output signal is developed across R.

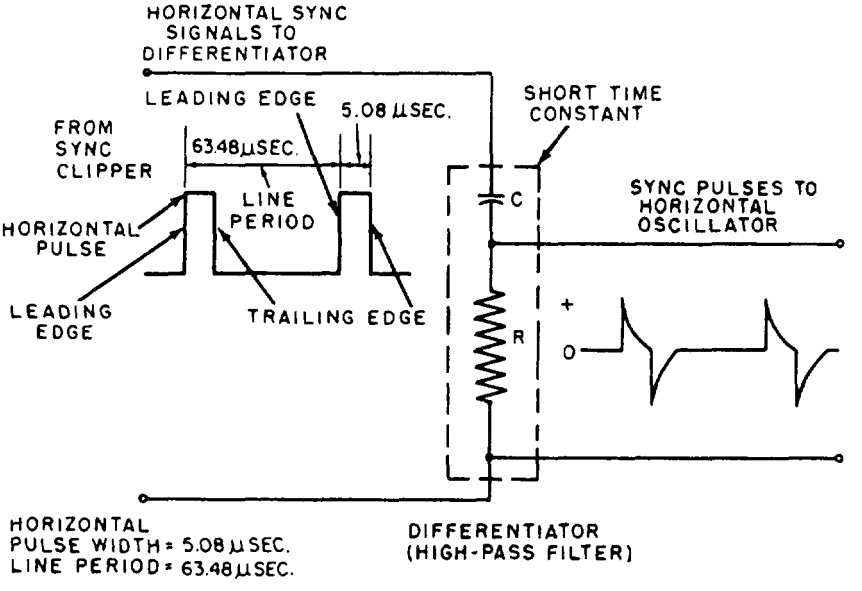


Figure 8-49.-High-pass filter for horizontal sync.

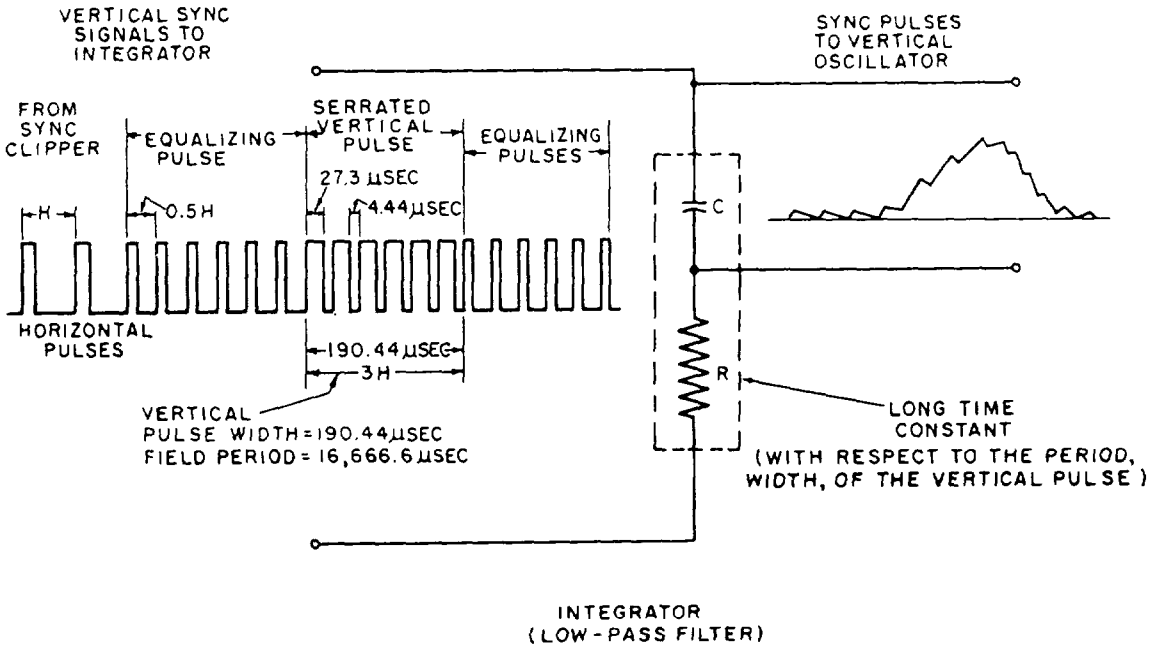


Figure 8-50.-Low-pass filter for sync pulses.

The leading edge of the square-wave input pulse causes a rapid charge of C through R. The trailing edge causes an equally rapid discharge of C through R. The flow of charge and discharge currents through R causes the sharp spikes of output voltage, as shown in figure 8-49. Only one spike in each pair (for example, the positive- spike) is needed to trigger the horizontal oscillator. The other spike of the pair occurs at a time when the oscillator is insensitive to triggering pulses.

The low-pass filter used to pass and shape the trigger voltages for the vertical oscillator (a multivibrator or blocking oscillator) is shown in figure 8-50. The RC time constant is long with respect to the width of each serration in the vertical pulse. Because of the long time constant in the integrator circuit, C does not have time to discharge during the interval between serrations. However, the RC time constant is short compared to the period of the combined vertical serrated pulses. Thus, C charges up to the peak value during the time the vertical serrated pulse is applied (190.44 sec in the figure) and discharges to zero before the next horizontal pulse arrives.

The circuit is relatively insensitive to the longer, low-repetition-rate pulses that control the vertical oscillator. The reason is clear when you note that the time constant (with respect to the width of each horizontal pulse) of the differentiator circuit is short and the output is taken across the resistor.

Because of the long time constant (with respect to the width of each horizontal pulse) of the integrator circuit, the horizontal pulses have very little effect, and the equalizing pulses have even less effect. The only pulse that produces a useful output is the serrated vertical pulse. Sixty of these pulses occur each second, 30 for each of the two fields. The vertical pulses are serrated to provide the triggering action for the horizontal oscillator during the vertical retrace period.

SWEEP CIRCUITS

After being separated and shaped, the horizontal and vertical sync pulses are applied to the vertical sweep oscillators, respectively, so they maybe triggered at the correct instant to synchronize the receiver with the transmitter. Both the vertical and horizontal sweep oscillators, when fed into the correct circuits, produce current sawtooth waveforms.

The sawtooth waveforms produced by the horizontal sweep oscillator are amplified and applied to the picture tube in a manner that will cause the electron beam to be deflected (swept) horizontally across the face of the tube. Likewise, the waveforms produced by the vertical sweep oscillator cause the electron beam to be deflected from the top to the bottom of the picture tube. Multivibrators and blocking oscillators are two types of resistance-capacitance oscillators commonly used in the sweep circuits (vertical and horizontal) of TV receivers.

In electromagnetic deflection systems, the driving force in the picture tube is a magnetic field. To develop such a field, a sawtooth deflection current is required. Further, because of the inductive action in the output state of a tube-type vertical-output amplifier, a

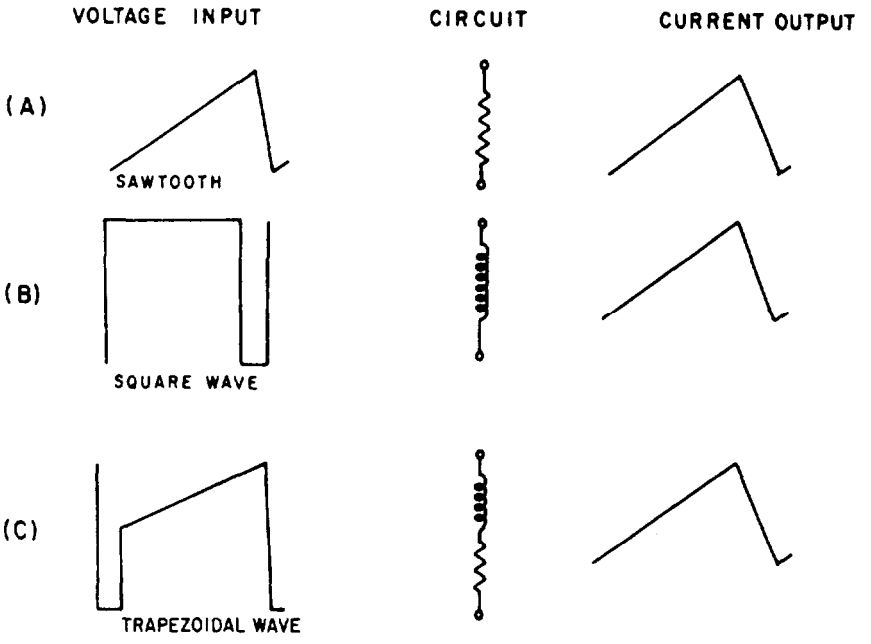


Figure 8-51.-Trapezoidal waveform.

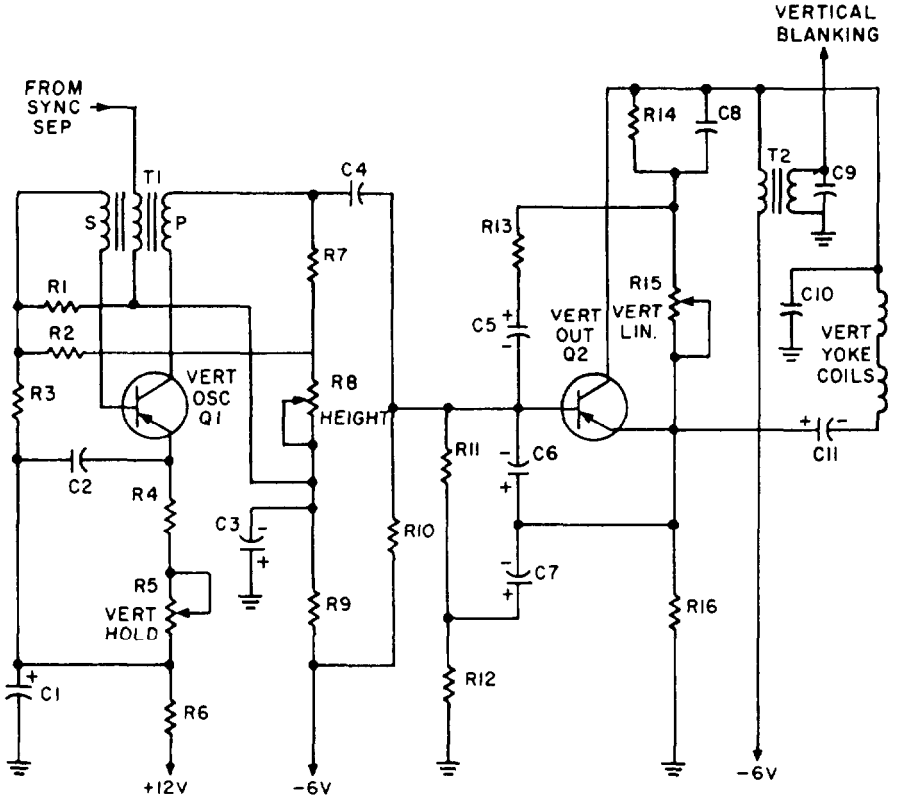


Figure 8-52.-Vertical sweep section.

trapezoidal waveform is required to produce this sawtooth of current. This is illustrated in figure 8-51. To produce a sawtooth of current through an inductive circuit, a square wave of voltage must be applied (view B). To produce a sawtooth of current through a resistive circuit, a sawtooth of voltage must be applied (view A). To produce a sawtooth of current through an inductive-resistive circuit, both a sawtooth and square wave (trapezoidal wave) must be applied (view C).

Vertical Sweep

The vertical sweep circuit (fig. 8-52) produces a current that moves the electron beam of the picture tube

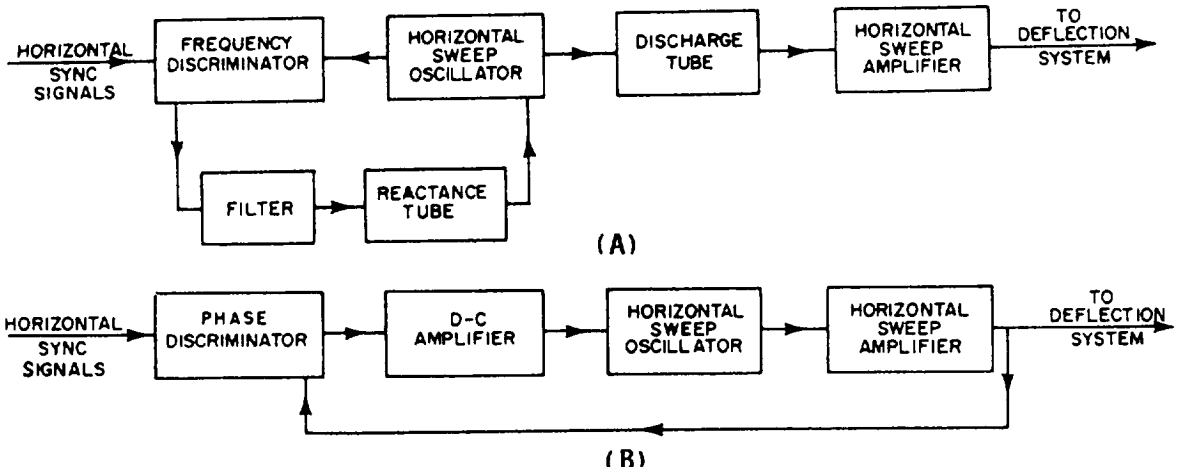


Figure 8-53.-Systems of horizontal sync control.

from the top to the bottom. You should note that when transistors are used for the vertical and horizontal output amplifiers, the input wave to the output transistor is generally not trapezoidal but is close to a sawtooth.

The vertical oscillator (Q1) shown in figure 8-52 is a conventional blocking oscillator. The signal from this oscillator is fed to the vertical output stage (Q2) and on to the series-connected vertical yoke coils. Oscillator frequency is determined by the RC time of C2, R4, and R5. The height control (R8) varies the collector bias voltage, thus changing the amplitude of the output signal. Without compensation, this could change oscillator frequency since it affects the charging capacitor by changing the amount of charge. Hence, a compensating network is included to prevent this from occurring. R1 is connected from the negative six-volt supply to the secondary of winding T1, shunting both R8 and R2. The effect of this network is to change the forward bias on the oscillator so that the frequency is shifted in the opposite direction to compensate for any shift caused by the height adjustment.

The oscillator signal is coupled to the base of the output stage through C4. Forward bias for Q2 is provided by the voltage divider network (R10, R11, and R12) in the base circuit. Both emitter and collector currents normally flow. The positive pulse from the vertical oscillator opposes this forward bias and abruptly reduces the emitter-collector current. This period of abrupt current drop is the vertical retrace time and lasts for the duration of the oscillator pulse. C5, C6, and C8 combine to form the sawtooth wave for the vertical trace period. R15, in series with C8, forms a waveshaping feedback circuit between the emitter and collector of Q2. C5 and R13 perform the same function between the output and the base.

Vertical output transformer T2 is a voltage step-up device with the secondary providing sufficient voltage for vertical retrace blanking. C9 is connected from the secondary to ground to prevent any horizontal pulses from distorting the vertical output. The vertical yoke coils, which are actually the collector ac load, are returned to the emitter of Q2 through C11 to isolate the yoke from dc. C10, connected from the collector to ground, provides further faltering.

Automatic Frequency Control (AFC)

The use of incoming sync pulses to trigger and control the vertical and horizontal sweep oscillators is the simplest, most economical, and most direct method of controlling the motion of the electron beam in a TV. This simple, direct system would be satisfactory if it were not for the presence of noise pulses that may cause the oscillators to fire at the wrong time. When the vertical oscillator fires at the wrong time, the picture is not properly synchronized vertically. The picture bounces, or moves in jumps upward or downward across the screen. When the horizontal oscillator fires at the wrong time, the picture is not properly synchronized horizontally. The picture tears or becomes streaked, giving the appearance that the picture is jumbled.

Although noise pulses may affect the operation of troth the vertical and the horizontal oscillators, a far worse effect is felt by the horizontal oscillator. The long time constant of the vertical filter makes it insensitive to the short bursts of noise energy, and the effect on the vertical oscillator is not generally objectionable. In closed circuit TV, noise from electrical machinery can be just as destructive as atmospheric noise in a conventional home TV system.

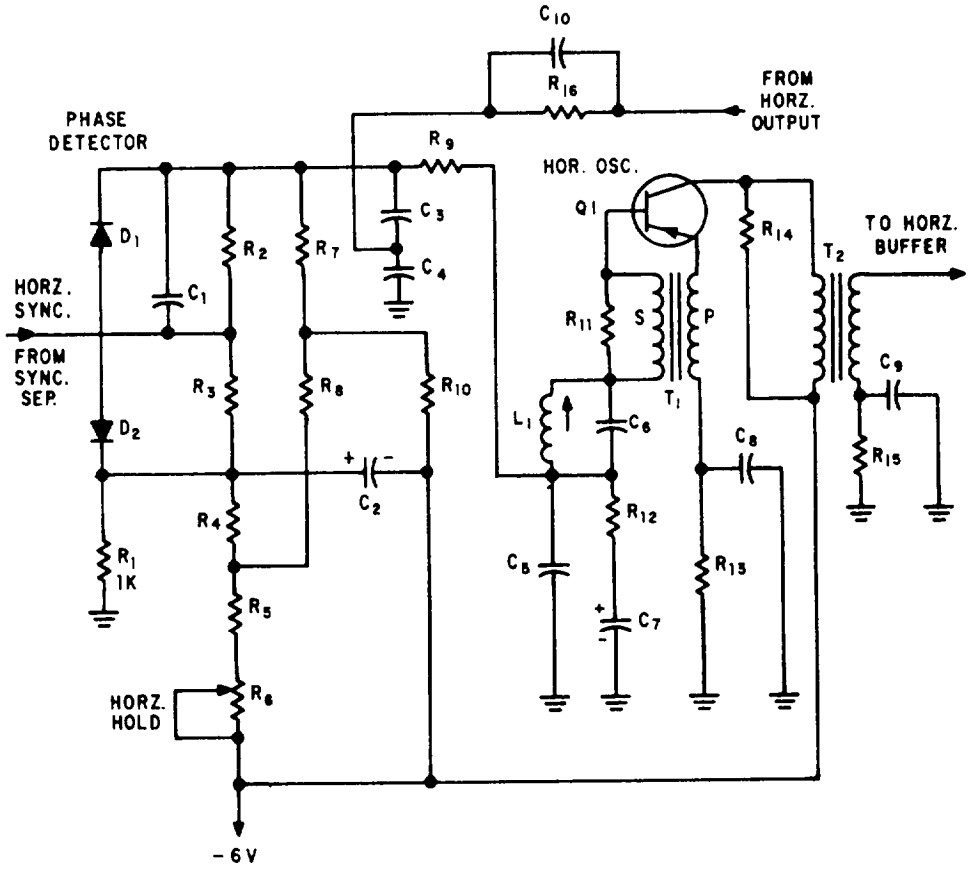


Figure 8-54.-Horizontal sweep (Part A).

The short time constant of the filter that feeds the sync pulses to the horizontal oscillator permits the passage of short bursts of noise energy. Consequently, it is necessary to employ a control circuit that effectively isolates the horizontal oscillator from the effects of noise pulses, and at the same time permits the sync pulses to assume control.

Two systems that isolate the horizontal sweep oscillator from the effects of noise bursts are shown in the block diagrams of figure 8-53. In the system shown in view A, two signals are applied to the frequency discriminator. They are horizontal sync signals and horizontal sweep oscillator signals. The frequency discriminator compares the frequency (or phase) of these signals and produces an output dc voltage that depends on the difference between the frequencies (or phase) of the two signals. The output voltage, normally varying at a relatively slow rate, is fed via a low-pass filter to the grid of the reactance tube. This tube functions in such a manner that its output changes the frequency of the horizontal sweep oscillator to maintain its frequency exactly the same as that of the incoming horizontal sync pulses.

The second method of isolating the horizontal sweep oscillator from noise bursts is shown in figure 8-53, view B. As in the previous system, a frequency (or phase) discriminators used. It compares the sync-signal input from the filter or sync amplifier with the input feedback from the horizontal sweep amplifier and produces a dc output that is proportional to that difference. The reamer in which this dc output is amplified and used causes the frequency of the horizontal sweep oscillator to lock in step with the incoming sync signals.

Horizontal Sweep

In general, the same basic types of deflection oscillators found in the vertical deflection system are also used in the horizontal system. These are the blocking oscillator and the multivibrator types that employ both tubes and transistors.

Figure 8-54 is a simplified schematic diagram of the phase detector and oscillator portion of the horizontal sweep section of a transistorized TV receiver. The phase detector samples a feedback signal from the horizontal output stage and compares its phase with the incoming

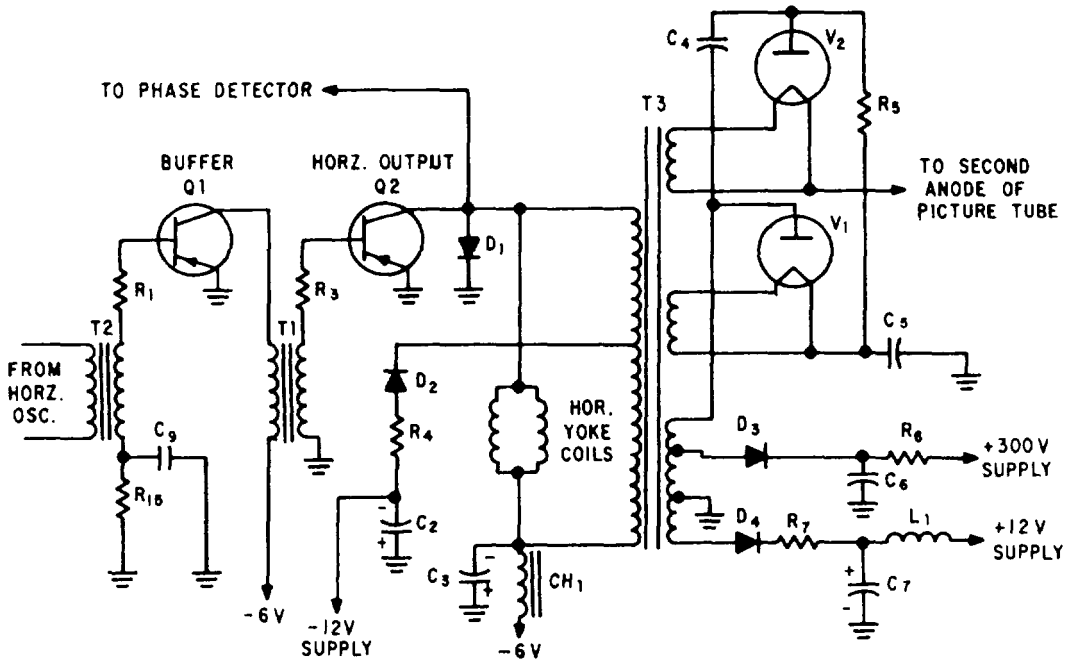


Figure 8-55.-Horizontal sweep (Part B) and high voltage.

horizontal sync pulses. Any phase difference results in an appropriate dc voltage from the phase detector being applied to the oscillator. This signal then causes the oscillator to change its frequency.

When the pulses are in phase, the outputs from D_1 and D_2 are equal and opposite, thus no dc control voltage is produced for the base of the oscillator. When the oscillator frequency is low or lagging the sync, D_1 conducts more heavily than D_2, producing a positive control voltage at the output. This positive voltage, applied to the base of Q_1, increases the frequency of oscillations. When the frequency of the oscillator is high or leading the sync, D_2 conducts more heavily, producing a negative control voltage that lowers the frequency of the oscillator.

The horizontal oscillator, Q1, is a blocking oscillator that operates similar to the one in the vertical section. Transformer T_1 is the blocking-oscillator transformer, and L_1 is a stabilizer coil.

Stabilizing coil L_1 is in series with the secondary of T_1 and is shunted by C_6. When an abrupt current change through the coil occurs due to oscillator action, C_6 discharges through the coil. The resulting magnetic field around the coil induces a voltage that recharges the capacitor, but to a lower level than previously charged. Thus, the oscillations are damped out. The timing of this ringing action is adjustable to the point where it helps fire the oscillator at a constant rate, thus providing a degree of stability.

Horizontal hold control R_6 establishes the bias on the base of Q_1 to bring the oscillator within control range of the phase detector. Since the setting of R_6 can affect the balance of the phase detector, isolating and compensating resistors R_4, R_7, R_8, and R_{10} are incorporated to minimize this effect. The horizontal output pulse that is fed back to the phase detector is shaped into a sawtooth wave by C_4, C_{10}, and R_{16}.

The balance of the horizontal sweep circuits (buffer, horizontal output, and high voltage) are shown in figure 8-55. The positive output pulse of the oscillator is inverted by interstage transformer T_2 and applied to the base of Q_1 as a negative pulse. The buffer is normally cut off since there is no forward bias from the base to emitter.

The negative pulse at the base of the buffer amplifier drives the stage into conduction. RC network C_9 and R_{15}, however, develops a base bias from the signal current, which permits conduction only on the peaks of the pulses. Since a common emitter circuit is used, a positive pulse appears at the collector. The positive signal from the buffer is transformer-coupled through T1 to the base of the horizontal output transistor Q2.

The output transistor is a special power-type transistor capable of handling the relatively high-power requirements of the stage. At the collector, a positive

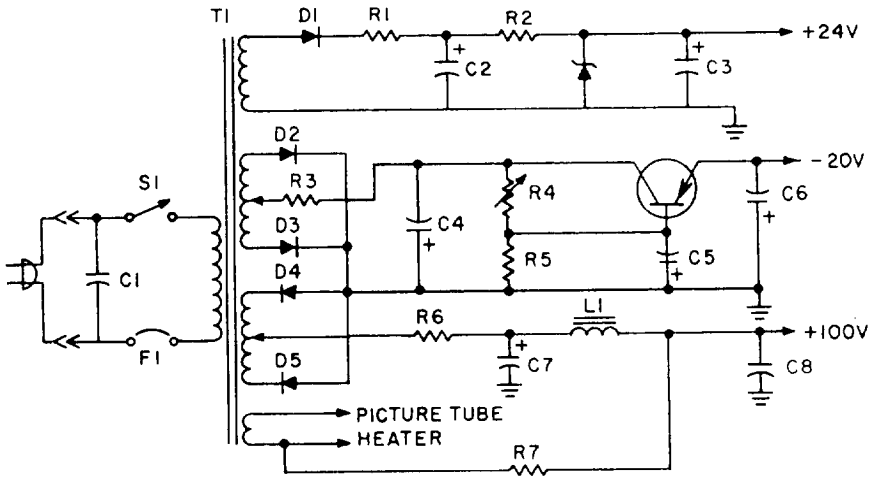


Figure 8-56.-Black and white, solid-state power supply.

horizontal sweep pulse appears and is directly coupled to the horizontal yoke coils connected in parallel. The horizontal pulse causes a relatively linear sawtooth current in the yoke coils. Yoke damping action is provided by diode D_1, which dampens any self-resonant oscillations of the yoke.

A tap on the primary of horizontal output transformer T3 is used to develop a -12 volt supply through diode D_2. Two taps on the secondary of the same transformer provide +12- and +300-volt power supplies through their respective diode rectifiers and filters.

When the horizontal output transistor is conducting, collector current is drawn through the primary of T3, and the collector voltage declines from -6 volts in a positive direction, producing a positive horizontal sweep pulse. The heavy current flow through the transformer primary causes high peak-voltage pulses to be developed across the high-impedance secondary winding. These pulses are applied to the plate of the first high-voltage rectifier, V_1. The rectifier current charges C_5 to approximately 5,000 volts. This voltage is applied to the plate of the second rectifier, V_2, through R_5. Capacitor C_4 also couples the high-voltage pulse from the transformer secondary to the plate of V_2, thus adding to the high dc voltage applied through R_5. As a result, approximately 10,000 volts appear at the filament of V_2. This is the high voltage that is applied to the second anode of the picture tube (kinescope). Voltages up to 30 kV are found in color TV high-voltage systems.

TV POWER SUPPLIES

Learning Objective: *Identify the circuitry of low-voltage color and black and white power supplies.*

Because of the high-voltage requirements of the CRT, TV receivers employ two power supplies–high voltage and low voltage. The high-voltage section was discussed in the previous section. A typical transistorized low-voltage supply will now be discussed.

New solid-state TV receivers use both low-voltage and high-voltage transistors. This requires a large assortment of voltages. Since transistors are low-impedance devices, they work best when driven by voltages from low-output-impedance supplies.

A power supply that provides multiple low-impedance output voltages is shown in figure 8-56. The -20 and +24 output voltages drive most of the low-level stages in the set. A Zener regulator is used for the +24 volt supply, and an active filter is employed in the -20 volt supply. The +100 volt output services the output stages of the following sections: video, vertical, horizontal, and audio. Note the wiring arrangement of the two full-wave rectifiers. This is done in some receivers so that part of the diode can be physically connected to ground. The ground is then used as a heat sink, eliminating the need for an insulator between ground and the rectifier body.

Generally speaking, color TV power supplies are slightly larger and more complex than those found in black and white sets. This is due in part to the power

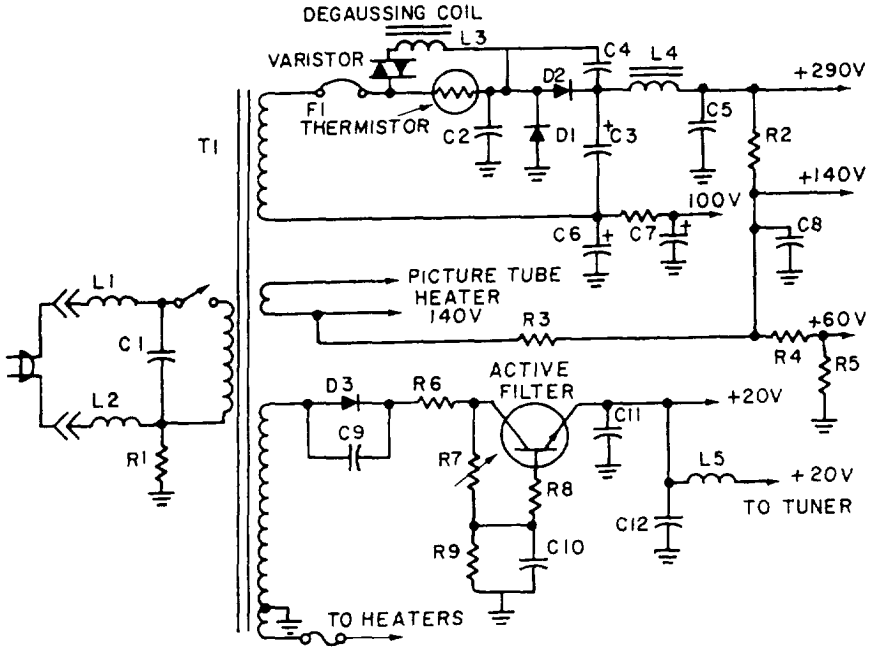


Figure 8-57.-Color, solid-state power supply.

required by the extra circuits in the color sets and the automatic degaussers in the B+ line.

Figure 8-57 shows a typical power supply used in large-screen color TV receivers. This type of power supply, with its numerous output voltages, feeds both transistors and tubes. Most of the transistors receive their power from the +20 volts developed by the active filter. After a little extra faltering, this same +20 volts is sent to the tuner. R7 is used to adjust this output. All of the tube heater filaments, except the picture tube, receive their power from the same output winding that provides the +20 volt outputs. The picture tube heater is on a separate winding that is biased to +140 volts to prevent cathode to heater shorts.

The +290-volt B+ supply has a degaussing circuit in series with it. A thermistor and varistor control the current in the degaussing coil. As the circuit warms up, the thermistor decreases in resistance and the varistor increases in resistance. The net result is a surge of current through the degaussing coil for a few seconds before the set warms up. Very little current then flows through the coil during normal receiver operation. The +290-volt supply uses a half-wave voltage doubler. Capacitors C2 and C4 are across the diodes for transient protection. This feature gives the diodes a much longer life.

CHAPTER 9

COMPUTERS AND PROGRAMMING

Chapter Objective: *Understand the operating principles of computers to include programming fundamentals as well as an overview of the ATLAS programming language.*

As late as the middle 1970's, the phrase "kick the tires and light the fires" was the main theme in launching an aircraft sortie. This meant that as long as there was an airframe with nothing falling off, an engine that would start and achieve takeoff speed, and air in the tires, the aircraft would be launched so pilots could get flight time. In the last 10 to 12 years, the term mission has become the prime objective of the aircraft. This is not meant to belittle the importance of the engines and airframes. Obviously, they are important, but aircraft and pilots are designated to perform certain missions. The performance of these missions is dependent upon the status of the various avionics packages. If one or more of these packages are degraded or does not work at all, the aircraft is considered to be partial mission capable or not mission capable. This lack of mission capability has thrust many an avionics work center supervisor into the spotlight. If you are one of the supervisors who has been there, then you know how *pleasant* the maintenance chief is to you. Then you realize that aircraft maintenance is not a game.

As we head toward the 21st century, newer and more sophisticated aircraft are being designed and built. The avionics systems are becoming more complex, thus allowing the aircraft to perform more complex missions. The increased complexity forces the solutions to problems in microseconds. The only system capable of performing these solutions is the computer. In turn, each associated avionics system will act as a sensor that feeds continuously updated information to the computer. The computer assimilates the data and sends out information where it is needed. Future aircraft may even include sending data to and from the engines, flight controls, fuel systems, and hydraulic systems. The F/A-18 does part of this now.

Because computers are used so extensively in Navy aircraft, the avionics supervisor must have a basic understanding and working knowledge of computers.

COMPUTER MAKEUP

Learning Objective: *Identify electronic components of a computer.*

The electronic components of a computer (cathode-ray tubes, transistors, microchips, printed circuit cards, and soon) are commonly called *hardware. Software,* on the other hand, is a term that is applied to a set of computer programs, procedures, and possibly associated documentation concerned with the operation of a data processing system. Software includes compilers, assemblers, executional routines, and input/output libraries. The advances in computer software provide the industry with the greatest realm of application possibilities. The problem of attempting to communicate with a computer has led to the development of symbolic languages that approach human language. The fact that a person can tell a computer what to do, just as one directs the actions of another person, has been made possible by the advances in software.

Programming in a universal language has led to the development and refinement of a number of computer languages. Many of these languages are for a special area or purpose; for example, FORTRAN (FORmula TRANslator) for business and scientific programs, COBOL (COmmon Business Oriented Language) for business, and Jovial for large scale, computer-based, command and control systems. PL/1 (programming language/one) is a language for real-time systems. Each of the languages fulfills a specific need for a specific problem but lacks the universal ideal application.

Software is also used to overcome design deficiencies in computers. Programming around design deficiencies is a common practice in the computer industry. Software is, in fact, often used to determine design feasibility. The practice of designing a computer with a computer is a common practice of design engineers.

Perhaps the best software application has been in the area of real-time processing. Real-time processing is a situation in which data is submitted to the computer and an immediate response is required. The capability of a computer to perform real-time processing could determine the success or failure of an aircraft's mission.

COMPUTER APPLICATIONS

Learning Objective: *Identify computer applications.*

Computer applications fall into a variety of broad categories. Information retrieval is one such application, or in a narrower sense indexing or cataloging. Information is stored under a variety of key words or index headings. By calling up one of these headings, a listing of all or part of the information will be output by the computer. Another application is simulation. This involves simulating the operation of a new computer using an older computer model. In this way, design deficiencies can be identified without going through the time-consuming and expensive process of building the newer unit. Real-time control of a production process is another application. For example, the petroleum and chemical industries put this process to good use. The computer can detect minute changes in the production process and initiate immediate corrective action.

The advent of personal (home) computers has greatly expanded the computer-use horizon, from the routine upkeep of a checkbook balance to the more complex functions of financial planning, home security, and computer video games. The application of the computer and its functions is virtually endless. For this reason, there are some people who believe that the computer will soon control everything and everyone. This is not necessarily the case, however, because computers can do only what their creators have intended them to do. The computer enables man to do more than he has been able to do before. For example, computations that required years to calculate by human methods can now be accomplished in a matter of moments by modern computers. This has become particularly evident in our space program. The ability to put a man on the moon and send *Voyager I* and *Voyager II* on their journeys would have been impossible without the use of computers. Fears over job losses are, for the most part, needless. While some jobs may be eliminated, new ones are created. Thus, a worker may have to learn a new skill (a laborer may have to be retrained as a computer programmer or operator). Rather than destroying jobs, the computer creates opportunities where none existed before.

TYPES OF COMPUTERS

Learning Objective: *Identify the types of computers and the analytical processes used by each type.*

In general, there are two basic types of computers: *analog* and *digital.*

ANALOG COMPUTERS

The term *analog,* as applied to computers, pertains to representation by means of continuously variable physical quantities. For example, an analog computer can be a device that solves problems by setting up electrical circuits that represent the physical equivalents of certain phenomena and by making measurements of them. These electrical circuits vary with changes in the phenomena. However, the analog computer is by no means restricted to electrical circuits as equivalents. The physical equivalents may be gear trains, gases, fluids, and so on. A digital computer, on the other hand, is a device that solves problems by manipulating the numerical equivalents of phenomena in accordance with mathematical and logical processes. These numerical equivalents may be expressed as binary numbers, octal numbers, decimal equivalents, and so forth. In an electronic digital computer, the numerical equivalents are generally expressed as binary numbers: 1s and 0s. Values of voltage and current are used to represent the 1s and 0s of the binary numbers.

Analog computers, because of their nature, have some inherent limitations. The use of physical equivalents limits their versatility. They are limited to performing only the tasks for which they were designed or, in certain instances, closely related tasks.

DIGITAL COMPUTERS

The versatility of digital computers is based on the fact that they use numerical equivalents not only to represent the data to be processed, but also the instructions for processing the data, In other words, digital computers are generally provided with a wide variety of instructions. They are designed to respond in certain ways to the numerical equivalent of these instructions. Programming is merely a matter of modifying and/or arranging these instructions so that the computers will respond in a predictable manner to a given situation. While much more versatile than analog

systems, digital systems are still limited to the variety of tasks they can perform by such factors as follows:

- The design of their central processors
- The variety of input/output devices used
- The programmer's capability to develop a numerical method for representing and solving the problem

There are two basic types of digital computers: special-purpose and general-purpose.

Special-Purpose Digital Computers

Special-purpose digital computers are designed to follow a specific set of instruction sequences that are fixed at the time they are manufactured. To change the operation of this type of computer, the actual construction of the machine has to be altered.

General-Purpose Digital Computers

General-purpose digital computers follow instruction sequences that are read into and stored in memory prior to the calculation performance. This type of computer operation can be altered by inputting a different set of instructions. Since the operation of general-purpose digital computers can be changed with relative ease, as compared to the special-purpose computers, they provide a far greater usage flexibility.

DIGITAL COMPUTER OPERATION

Learning Objective: *Recognize operating principles of a digital computer.*

Each major section of the digital computer is comprised of various electrical circuits, such as flip-flops (bistable devices), amplifiers, AND circuits, OR circuits, and passive memory elements. These elements are, in turn, organized into registers (a series of electronic devices for temporary storage of a binary word), counters (a series of electronic devices that progress through a specific binary sequence), and gates (AND or OR functions to set a flip-flop or generate a times condition signal). The computer manipulates binary numbers (1s and 0s) representing numerical values or conditions. Devices to retain these binary figures comprise the majority of the computer registers, and each register has a distinct purpose or function. Many operations require that the binary word or data be transferred from one register to another, and, possibly, several different words may be transferred simultaneously.

Gates are used to control the transfer of data words from one register to another. These gates consist of diode and resistor networks. The gate circuit generates a signal to transfer the contents of one register to another at a particular time if certain conditions are met, such as if the instruction being executed is an add, and if one of the numbers being added is a negative number. If these conditions are met, a command signal is generated; if they are not met, the signal is not generated.

Several gates in the computer are active only during specific instructions, such as divide, and, then possibly, only during that instruction. On the other hand, some gates generate signals that are common to several instructions. In the design of a computer, each instruction that the computer is to perform is very methodically analyzed, and, for each signal required, a gate is designated to generate the signal.

The size of the registers determines the general size of the computer. Not all registers in the computer have the same word length. Some are determined by the accuracy required, while others are determined by the instruction word, number of addresses in the memory, and various other parameters.

DIGITAL DATA PROCESSOR

Learning Objective: *Referring to various schematic and block diagrams, recognize the components of a digital data processor and the function(s) of each.*

Figure 9-1 is a functional block diagram of a digital data processing set. Of the processes that take place

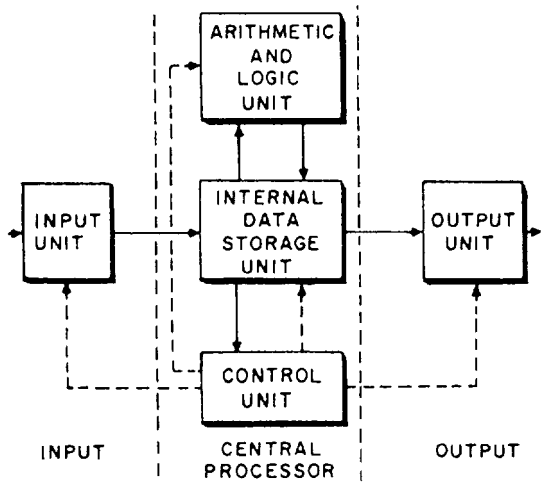


Figure 9-1.-Representative digital data processor block diagram.

within a computer, the manipulation of data is the most important. This data manipulation takes place within the central processor (CP) (area encompassed by vertically dashed lines). The CP is composed of three basic units:

- Control unit–This unit directs the overall operation of the computer in accordance with a prescribed plan.

- Arithmetic-logic unit–This unit performs the actual processing.

- Internal data storage unit–This unit stores the data to be processed and the prescribed plan (program).

CONTROL UNIT

In a typical digital computer, the control section includes the instruction register, the P register, the general register(s), and SC register. An explanation of these registers follows.

Instruction register. This register holds the instruction code during execution. The size of the register is dependent upon the instruction word and makeup of the computer. The instruction code usually has more than one part or field.

P register. The P register contains the address of the next sequential instruction to be executed. The contents of the P register are automatically advanced by one by the P + 1 adder.

General register. This register stores the quantity used for address modification. In addition, it usually has the properties of automatic increment or decrement. Most computers have more than one general register.

SC register. The SC register consists of one or two registers to hold a shift count. Its size is dependent on the maximum number of places that a word can be shifted.

An easy way to comprehend the operation of the control unit is to compare it to a telephone exchange. The act of dialing a phone number energizes certain switches and control lines in a telephone exchange. In a similar manner, each program instruction, when executed, causes the control section to energize certain "switches" and "control lines." This enables the computer to perform the function or operation indicated by the instruction.

A computer program can be stored in the internal circuits of the computer or it may be read instruction-by-instruction from external media. The internally stored-program type of computer (generally referred to as a "stored-program" computer) is the most practical type to use when speed and fully automatic operation is desired.

In addition to the command that tells the computer what to do, the control unit also dictates how and when each specific operation is to be performed. It is also active in initiating circuits that locate information stored in the computer and in moving this information to the point where the actual manipulation or modification is to be accomplished.

In the stored-program computer, the control unit reads an instruction from the memory section (as instructed by the program). The information read into the control unit from memory is in the form of varying voltage levels that make up a "binary word," and represents a specific operation that is to be performed. The location of the data to be operated on is generally a part of the instruction and energizes circuitry that causes the specified operation (add, subtract, compare, and so on) to be executed. Subsequently, the control unit "reads" the next instruction or jumps as directed to find the next instruction to execute.

Computer instructions are broken down into four general categories. These categories are transfer, arithmetic, logic, and control.

Transfer commands transfer data from one location to another. One of the instructions is usually an address in memory, and the other is either a register or an input/output device.

Arithmetic instructions combine two pieces of data to forma single piece of data, using one of the arithmetic operations. In some computer types, one of the pieces of data is in a location specified by an address contained in an instruction, and the other is already in a register (usually the accumulator). The results are usually left in the accumulator.

Logic instructions make the digital computer much more than an expensive "adding machine." The use of logic instructions enables the programmer to construct a program capable of a number of tasks. These instructions enable a computer used for inventory maintenance to follow one set of procedures if an inventory item count is too large and another if the count is too small. The choice of which set of procedures to use is made by the control unit under the influence of the logic instructions. Logic instructions provide the computer with the ability to make decisions based on the results of previously generated data.

Control instructions send commands to devices not under direct control of the control unit, such as input and output units. The address portion of the control instruction does not specify a location in memory, but is usually a coded group specifying an action required of a particular piece of equipment.

In a single-address computer, where each instruction refers to only one address or operandi, the instructions are normally taken from the memory in sequential order. If one instruction comes from a certain location, such as X, the next instruction is usually taken from location X + 1. However, the execution of a logic instruction may produce a result that dictates that the next instruction is to be taken from an address as specified in a portion of the logic instruction. For example, the logic instructions may initiate certain operations in the computer to determine if the content of a given register in the arithmetic section is negative. If the answer is yes, the location of the next instruction is specified in an address section of the logic instructions. If the answer is no, the next instruction would be taken from the next sequential location in the memory.

Every computer provides circuitry for a variety of logic instructions, thus providing the capability of selecting alternate instruction sequences if certain desirable or undesirable conditions exist. The ability to "branch" at key points is the special feature of the computer that makes it able to perform such diverse tasks as missile control, accounting, and tactical air plotting.

ARITHMETIC-LOGIC UNIT

The arithmetic-logic unit (ALU) is the section in which arithmetic and logic operations are performed on the input or stored data. The operations performed in this unit include adding, subtracting, multiplying, dividing, counting, shifting, complementing, and comparing.

Generally, information delivered to the control unit represents instructions, while information routed to the arithmetic unit represents data. Frequently, it is necessary to modify an instruction. This instruction may have been used in one form in one step of the program but must be altered for a subsequent step. In such cases, the instruction is delivered to the arithmetic unit, where it is altered by addition to or subtraction from another number in the accumulator. The resultant modified instruction is again stored in the memory unit for use later in the program.

All arithmetic operations can be reduced to any one of four arithmetic processes: addition, subtraction, multiplication, or division. In most computers, multiplication involves a series of additions; and division, a series of subtractions.

The arithmetic unit contains several registers-units that can store one "word" of computer data. This group of registers generally include D, X, and Q registers (so named for identification purposes only), and a unit called an accumulator (A register). During an arithmetic process, the D, X, and Q registers temporarily hold or store the numbers being used in the operation, called operandi. The accumulator stores the result of the operation. The control unit instructs the arithmetic unit to perform the specified arithmetic operation (as requested in the instruction). It then transfers the necessary information into the D, X, and Q registers from memory and controls the storage of the results in the accumulator or in some specific location in memory.

The arithmetic unit also makes comparisons and produces yes or no or go-no-go outputs as a result. The computer can be programmed so that a yes or go result causes the computer to perform the next step in the program, while a no or no-go instruction may cause the computer to jump several programmed steps. A computer can also be programmed so that a no result at a certain point in the program will cause the computer to stop and await instructions from a keyboard or other input device.

INTERNAL DATA STORAGE UNIT

In some digital computers, the internal data storage unit, or memory section, is constructed of small magnetic cores, each capable of representing an on (1) or off (0) condition. A system of these cores arranged in a matrix can store any computer word that is represented in binary form.

All computers must contain facilities to store computer words or instructions (which are intelligible to the computer) until they are needed in the performance of the computer calculations. Before the stored-program type of computer can begin to operate on its input data, it is first necessary to store, in memory, a sequence of instructions and all figures, numbers, and any other data that is to be used in the calculations. The process by which these instructions and data are read into the computer is called loading.

The first step in loading instructions and data into a computer is to manually place enough instructions into memory by using the console or keyboard so that these

instructions can be used to bring in more instructions as desired. In this reamer, a few instructions are used to "bootstrap" more instructions. Some computers use an auxiliary (wired) memory, which permanently stores the bootstrap program, thereby making manual loading unnecessary. These instructions may be stored in "chips" (referred to as "read only memories" or ROMs).

The memory (or storage) section of a computer is essentially an electronically operated file cabinet. It is actually a large number (generally between 1 and 40,000) of storage locations; each location is referred to as a storage address or register. Every computer word that is read into the computer during the loading process is stored or filed in a specific storage address, and is almost instantly accessible.

The types of memory storage devices used most frequently in present-day computer technology are magnetic cores, semiconductor memories, thin film, magnetic drums, magnetic tapes, and magnetic disks.

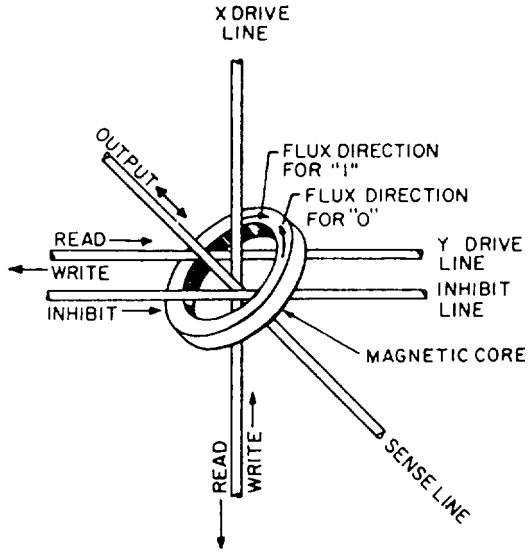


Figure 9-2.-Magnetic core showing X, Y, inhibit, and sense lines.

Magnetic Cores

One of the methods for storing internal data in a computer is realized by using magnetic cores. Cores are generally constructed by either of two methods. The first type of core, called a tape-wound core, is fabricated by wrapping a tape of magnetic material around a nonmagnetic toroidal form. A *toroid* is a term used to describe a doughnut-shaped solid object. The second type of core is called a ferrite core, and it is made by molding finely ground ferrite into a toroidal form. The ferrite used in this application is a ceramic iron oxide possessing magnetic properties. The ferrite particles are then heat-fused or "sintered" by the application of heat and pressure.

In magnetic core memories, each data bit is stored in the magnetic field of a small, ring-shaped magnetic core (fig. 9-2). Magnetic cores generally have four wires running through them. Two wires are used for READ selection. (These same two wires are used for WRITE by reversing the direction of current flow.) An inhibit wire prevents writing a 1 when a 0 is to be written. The sense wire picks up the signal voltage generated by the shifting of core from 1 to 0 in a READ cycle.

Since a single core stores only one bit of a word, a large number of cores are required to handle all the bits in every word to be stored. These cores are arranged in "arrays" (in rows and columns) to assign memory address locations and quickly write data and locate data for read-out purposes. The technique used most frequently for writing and reading data in magnetic core arrays is known as the coincident-current technique.

In computer memory applications, the ferrite core is magnetized by a flux field produced when a current flows in a wire (drive line) that is threaded through the core. The core retains a large amount of this flux when the current is removed. Flux lines can be established clockwise or counterclockwise around the core, depending upon the direction of the magnetizing current. A current in one direction establishes a magnetization in a core in a given direction. Reversing the direction of the current flow reverses the direction of the flux field and the core magnetization. These two unique states represent 0 and 1, respectively.

Semiconductor Memories

Semiconductor memories are used in many modem computers. Most of the semiconductor memories are of the MOS LSI type, which may be static or dynamic. MOS means metal oxide semiconductor, and LSI means large scale integration.

Thin Film

Thin film memory consists of Permalloy, a ferromagnetic material, deposited (under controlled conditions in a vacuum chamber) on a supporting material (substrate) of thin glass. When all air has been removed from the chamber, a shutter arrangement is opened, and vapors from molten Permalloy pass through

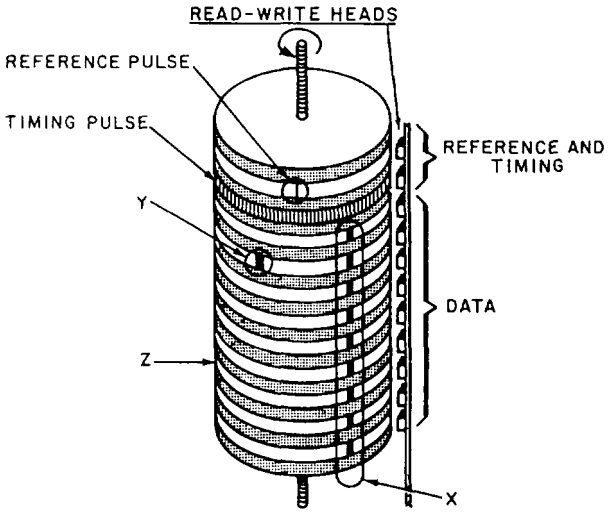


Figure 9-3.-Magnetic drum.

a mask and are deposited on the supporting material (substrate). The pattern thus formed is determined by the shape of the mask. The thickness of each spot (magnetized area) is controlled by the amount of time the shutter is open.

A magnetic field is applied parallel to the surface of the substrate during deposition. The film spots become easier to magnetize in the direction parallel to that in which the magnetic field was applied during the deposition process. This direction is known as the preferred direction; likewise, the axis of this magnetism is called the preferred axis.

Magnetic Drums

The magnetic drum storage device is a cylinder that rotates at a constant velocity. Information is written on or read from the drum when its magnetic surface passes under magnetic heads, which are similar to the magnetic heads found on commercial tape recorders.

Magnetic drums provide a relatively inexpensive method of storing large amounts of data. A magnetic drum (fig. 9-3) is made from either a hollow cylinder (thus the name drum) or a solid cylinder. The cylinder may consist entirely of a magnetic alloy, or it may have such an alloy plated upon its surface. Many drums are made by spraying on magnetite, an iron oxide. The surface is then coated with a thin coat of lacquer and buffed.

Representative drums have diameters ranging from 12.7 to 50.8 centimeters (about 5 to 20 inches, respectively). The surface of the drum is divided into tracks or channels that encircle the drum. A number of READ and WRITE heads (at least one for each track) are used for recording and reading. The drum is rotated so that the heads are near, but not touching, the drum surface at all times.

As the drum rotates, the tracks are continuously passing under their respective head. Each track is subdivided into cells, each of which can store one binary bit. All the cells that are positioned under the heads of a multitrack drum at the same time are called a slot. With some drums, each head “reads” or” writes” one bit of a word. Thus, when a word is written into or read from a slot, each track contains one bit of that word. The number of heads used depends on the size of the word that the computer is designed to handle.

One of the tracks provides timing signals for the drum rotation. The timing track determines the location of each set of storage cells around the chum. Each timing signal denotes a unit of time of the drum rotation. For example, if the timing track is 80 inches long and timing signals are recorded at 120 pulses per inch, there are 9,600 locations for bit storage on the track. If the drum has 32 tracks in addition to the timing track the drum has the capacity to store a total of 307,200 bits.

Some drums use two or even three timing tracks. These timing tracks are used for synchronization purposes and are sometimes called control or clock tracks. The timing pulses establish the time scale to which all circuits through the computer are synchronized.

The retrieval of data from a rotating drum can be a rather involved process, as can be realized by drawing a comparison to the core memory of a computer. When core memory is used, all the data is stored in the cores in a static condition. The data can be located at a given place at any instant and easily read from that location in serial or parallel form to represent the same data that was stored in that location.

Transfer of the data from constantly rotating magnetic drums, on the other hand, is complicated. Timing pulses are not used to synchronize the drum speed (which may vary slightly from time to time). Thus, some method must be used to ensure that data read into the drum memory in a given bit position will be read from the memory with the same time reference. The probability of an incompatible time relationship between the drum speed and synchronizing (clock) pulses makes it necessary to establish some means of compensating for variations in drum speeds.

In practice, the drum contains a control point and a number of sectors in a specific format. The control point

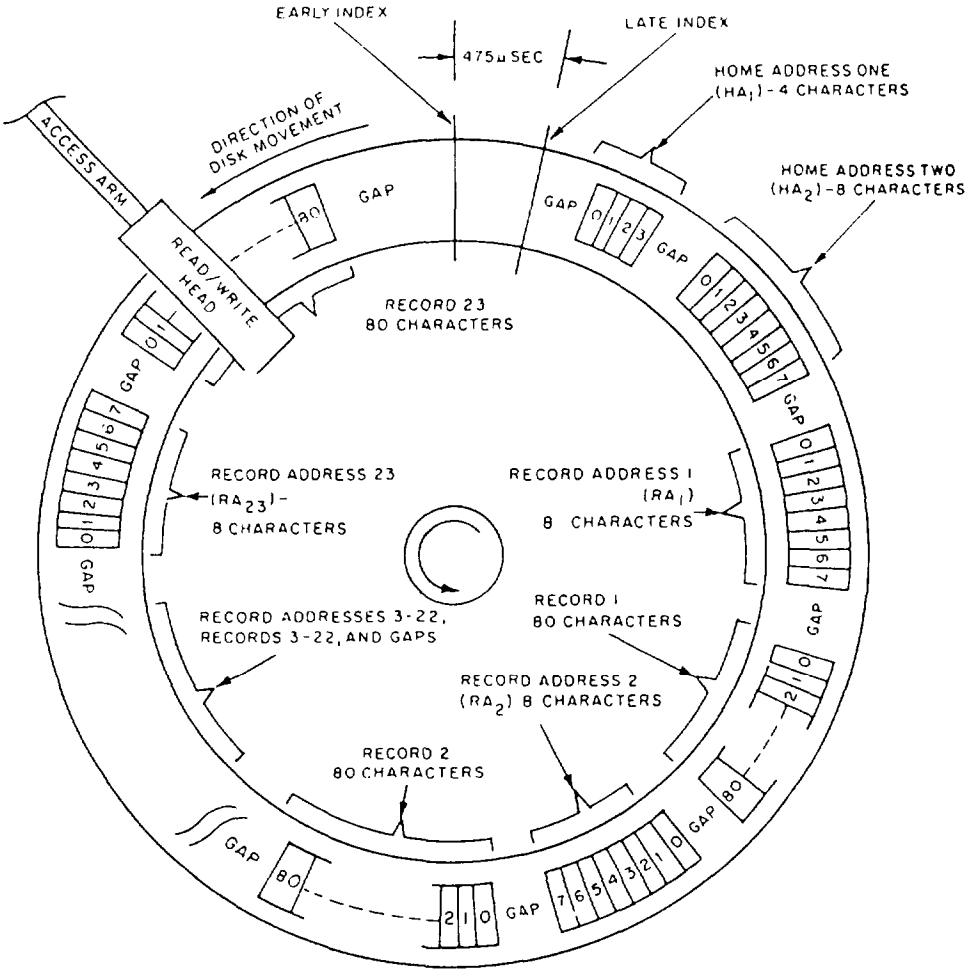


Figure 9-4.-Circular data track.

is a magnetic mark that specifies a starting location on the drum. All data stored on the drum is referenced to this indexing point or "reference pulse," as shown in figure 9-3.

Magnetic Tape

Magnetic tape is widely used as a storage medium for large amounts of data, or it may be used as a main storage backup. However, it is normally not used as an internal (main) storage medium because of its long access time. This is readily realized if you consider that needed information is widely (and sometimes randomly) distributed along the tape. Thus, the two main advantages of using magnetic tape are its large storage capacity and low cost.

Magnetic Disks

The magnetic disk is a convenient medium for semipermanent storage of mass volumes of production programs. For many applications, disks are superior to magnetic tape for rapid acquisition and storage of mass volumes of systems programs and data.

Magnetic disks resemble phonograph records that have been coated with iron oxide. The disks (or platters) are arranged in stacks in much the same way as a record stack in a jukebox. All the disks are continuously revolving and spaced apart so that a record head driven by an access mechanism can be positioned between the disks.

Data is recorded at a certain address on a specified disk. When readout of a particular bit of data is desired, the recording head is automatically positioned and the data is read serially from the surface of the selected disk.

The basic unit of information on the disk is called a character. By design, each character contains a given number of bits (for fixed-word applications). One or more of these characters in a group form a record. A circular data track (fig. 9-4) consists of one or more records, associated record addressees, gaps, and data

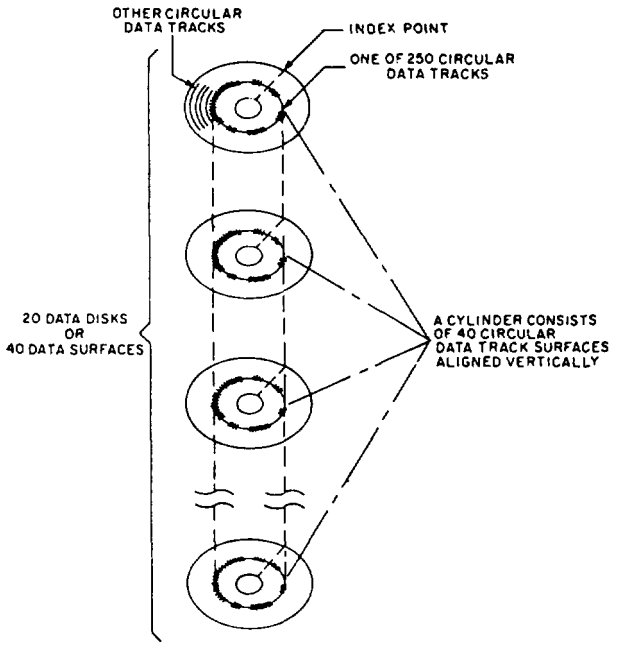


Figure 9-5.-Data storage disk assembly.

track identification. A number of data tracks aligned on vertically arranged disks (fig. 9-5) form a cylinder of information. A magnetic disk file system may contain one or more bands (modules). Each module contains a specified number of disks with their associated cylinders and data tracks. The flow chart in figure 9-6 illustrates the procedures necessary to retrieve or store information.

INPUT/OUTPUT (I/O) SECTION

Learning Objective: *Describe how digital computers communicate with external peripheral devices.*

The I/O section is that portion of the digital computer through which the CPU communicates with the external peripheral devices. In a useful computer function, data is read into the computer, processed, and then transferred to the output. The peripheral units handle the data input and output display functions. The I/O section controls the transfer of data between the computer and the peripherals.

The I/O section is the interface between the computer and any external devices. An interface is an assembly of electronic circuits that make the computer compatible with the peripheral units. This compatibility permits the computer and peripheral units to "speak" to and interpret one another intelligently. The compatibility involves logic levels, timing or speed, and control.

When digital data is transmitted between two units, the binary voltage or current levels must be compatible. Logic-level conversion is often required to properly interface different types of logic circuits. For example, logic-level shifting is often required to properly interface bipolar and MOS circuits. The speed of the data transmission must also be compatible. Some type of temporary storage between the two units may be required as a buffer to match the high-speed CPU to a low-speed peripheral unit.

Control is another function of the interface. There are status lines that tell when the computer or peripheral unit is ready or busy, and strobe lines that actually initiate the data transfers. This process is often referred to as "handshaking."

The type of information exchanged between the I/O unit and the peripheral devices includes data, addressing, and control signals. Since multiple I/O units can usually be connected to a computer, some coding scheme is required to select the desired unit. This is

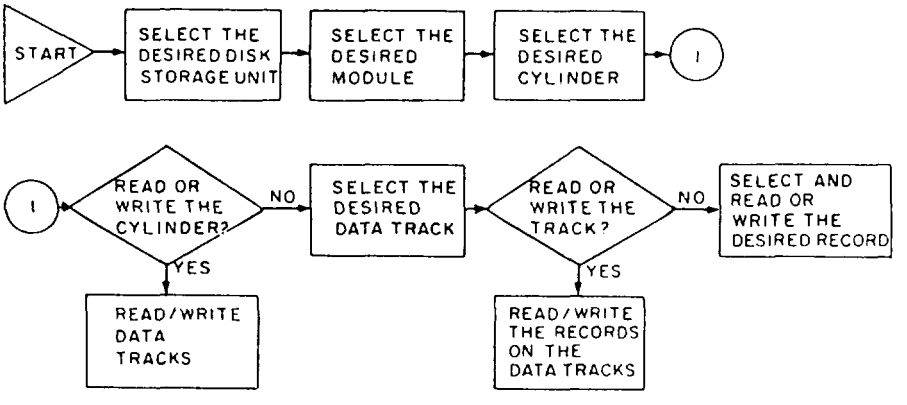


Figure 9-6.–Flow chart for storage and retrieval of data from disks.

usually done with a binary word used as an address. The address is transmitted to all the peripheral devices. The unit recognizing the address is connected to the I/O section. Data can then be transmitted to or from the device over the interconnecting data lines. The actual data transfers are controlled by control signals between the two devices.

Programmed data transfers that take place as the result of executing an I/O instruction usually cause data to be transferred between the peripheral unit and the accumulator register in the CPU. Other CPU registers may also be used, depending upon the computer architecture and the instruction. In some computers, peripheral units are treated as memory locations. The peripheral units are addressed as storage locations, and all memory reference instructions can be used in performing I/O operations. No special I/O instructions are used.

PARALLEL VERSUS SERIAL DATA TRANSMISSIONS

There are two methods of transmitting digital data: parallel and serial. In parallel data transmission, all bits of the binary data are transmitted simultaneously. For example, to transmit 8-bit binary numbers in parallel from one unit to another, eight interconnecting data transmission lines are required Each bit requires its own separate data path. All bits of a word are transmitted at the same time. Therefore, a significant amount of data can be moved in a given period of time. The disadvantage of the parallel method is the large number of interconnecting cables between the two units. For large binary words, cabling becomes complex and expensive. This is particularly true if the distance between the two units is great. Long multiwire cables are not only expensive, but also require special interfacing to minimize noise and distortion problems.

Serial data transmission is the process of transmitting binary words a bit at a time. Since the bits time-share the transmission medium, only one interconnecting lead is required.

While serial data transmission is much simpler and less expensive because of the use of a single interconnecting line, it is a very slow method of data transmission. Nevertheless, serial data transmission is useful in systems where high speed is not a requirement. Serial data transmission techniques are widely used to transmit data between a computer and its peripheral units. While the computer operates at very high speeds, most peripheral units are slow because of their electromechanical nature. Slower serial data transmission is more compatible with such devices, Since the speed of serial transmission is more than adequate in such units, the advantages of low cost and simplicity of the single interconnecting line can be realized.

PARALLEL DATA TRANSMISSION

In a parallel data transmission system, each bit of the binary word to be transmitted must have its own data path. There are a variety of ways to implement this data path. The two basic classifications of transmission line circuits are single-ended and balanced. Single-ended transmission systems use a single wire data path for each bit. When combined with a ground or return reference, the electrical circuit between the sending circuit and the receiving circuit is complete. In a balanced-transmission line system, two conductor cables are used to send the data. The data on the dual-transmission line is complementary. The dual-transmission lines also use a ground return reference. While a single-ended transmission line is simpler and less expensive, it is subject to more noise problems than the dual- or balanced-transmission line system.

SERIAL DATA TRANSMISSION

The simplest, most economical, and easiest to use method of transferring digital information from one point to another is serial data transmission. In a serial system, the digital data is sent one bit at a time; therefore, only a single pair of transmission wires is required. The serial transmission of data is far slower than parallel transmission. However, in most computer systems, the low-speed penalty is no disadvantage. Data rates achievable in serial data systems are sufficiently high to make them very practical.

Serial data transmission is preferred because it is inexpensive. It is especially beneficial in transferring data over long distances. For long distances, you can see that multiple parallel lines are far more expensive than a single cable.

Low-speed serial data transmission also offers another benefit; that is, the data rates are slow enough to permit the transmission of data over telephone lines. In this case the digital data is converted into a pair of audio tones representing binary 1s and 0s. These can be transmitted very economically for long distances over standard telephone lines. In addition, low-cost tape recorders can be used to record the serial data. This provides a low-cost means of mass data storage and

retrieval. Standard audio cassette recorders are widely used in this application.

Serial data transmission also permits transmission of data by radio. A radio communications path represents only a single interconnecting link similar to a transmission line pair. Therefore, for data to be transmitted by radio, it must be in serial format. Serial digital data is used to modulate a radio carrier in various ways.

In digital computer systems, you will find that both serial and parallel data transmission methods are used. Parallel methods are used where high speed and short distances prevail. Serial data transmission is used where low cost, simplicity, low speed, long distances, and minimum cost are necessary.

INPUT/OUTPUT (I/O) DEVICES

I/O devices are similar in operation but perform opposite functions. It is through the use of these devices that the computer can communicate with devices external to the computer itself (peripheral devices).

Input Devices

Input data may be in any one of three forms:

- Manual inputs from a man-machine interface (MMI) such as a keyboard or console
- Analog and/or digital inputs from instruments or sensors
- Inputs from a source on or in which data has previously been stored in a form intelligible to the computer

Computers can process hundreds of thousands of computer words per second. Thus, a study of the first method (manual input) reflects the inability of human-operated keyboards or keypunches to supply data at a speed that matches the speed of digital computers. A high average speed for keyboard operation is two or three characters per second, which, when coded to form computer words, reduces the data input rate to the computer to less than a word per second. Since the computer can read several thousand times this amount of information per second, it is clear that manual inputs should be minimized to make more efficient use of computer time.

Instruments used as input sensors are capable of supplying several thousand samples regarding pressure, temperature, speed, and other measurements per second. This is equivalent to 10,000 to 20,000 bits or binary digits per second. Digital computers that use these devices must be equipped with analog-digital (A/D) converters (assuming the input is in an analog format) to convert physical change to specific increments.

Input data that has previously been recorded on punched cards, perforated tapes, magnetic tapes, or magnetic drums and disks in a form understood by the program may also be entered into the computer. This method is much faster than entering data manually from a keyboard. The most commonly used input devices in this category are magnetic tape readers and paper tape (perforated tape) readers.

Output Devices

Output information is also made available in three forms:

- Displayed information such as codes or symbols presented on a monitor screen, which are used by the operator to answer questions or make decisions.
- Control signals, that is, information that operates a control device such as a lever, aileron, or actuator.
- Recordings, which is information stored in a machine language or human language on tapes or printed media.

Devices that store or read output information include magnetic tapes, punched cards, punched paper tapes, monitors, electric typewriters, and high-speed printers.

One of the main features of computers is their ability to process large amounts of data quickly. In most cases, the processing speed far exceeds the ability of input devices to supply information. One common limitation of most input devices is that each involves some mechanical operation; for example, the movement of a tape drive or card feeder. Because a mechanical movement of some part of these devices cannot take place fast enough to match electronic speeds with the computer, these input devices limit the speed of operation of the associated computer. This is particularly evident in cases where successive operations are dependent upon the receipt of new data from the input medium.

Several methods of speeding up mechanical operations have been devised, all of which are designed to move a smaller mass a shorter distance and with greater driving force. Many of these designs have been

directed toward increasing the drive speed of magnetic tapes. For example, present-day tape drives can pass up to 400 inches of tape per second over a tape-reading head. Card readers can read up to 2,000 cards per minute. Present-day disk systems operate at speeds up to 3,600 RPM.

The comparative rates of data for these systems are as follows:

- Card systems–2,700 characters per second
- Tape systems–350,000 characters per second
- Disk systems–15,000,000 characters per second

Another method of entering data into a computer, which we have not previously mentioned, is to link two or more computers together and program them to communicate with each other. This is perhaps the fastest method of entering or extracting data from a computer. An example of this is Data Link.

Regardless of the type of input/output (I/O) device being used, the purpose of the I/O section is to provide the computer with access to these devices or sensors.

The I/O section of a computer provides the necessary lines of communication and generates such signals as are necessary for the computer to establish communications with and, where necessary, to control the operation of the I/O devices. The I/O section, once it has been initiated by the control section, usually operates independently of the control section except when it must time-share memory with the control section.

PROGRAMMING FUNDAMENTALS

Learning Objective: *Recognize concepts and procedures used in construction of a computer program.*

Computer programming is the process of planning a solution to a problem. You can derive a general outline for calculating total resistance of a parallel resistance circuit by using the following steps.

1. Take the reciprocal of the resistance in ohms of all resistors in a circuit.

2. Calculate the sum of the values from step 1.

3. Compute the reciprocal of the value from step 2.

The process of preparing a program from this explanation is not difficult. One basic characteristic of the computer must be considered-it cannot think. It can only follow certain commands, which must be correctly expressed and must cover all possibilities. If a program is to be useful, it must be broken down into specifically defined operations or steps. These operations or steps, along with other data, must be communicated to the computer in a language that it can understand.

NOTE: The instructions are read sequentially unless otherwise stated.

Generally, the steps that a computer follows in the execution of a program are as follows:

1. Locates parameters (constants) and such data as necessary for problem solution.

2. Transfers the parameter and data to the point of manipulation.

3. Performs the manipulation according to certain rules of logic.

4. Stores the results of manipulation in a specific location.

5. Provides the user with a useful output.

Even with a simple program, such as the resistance program, each step must be broken down into a series of machine operations. These instructions, along with the parameters and data necessary for problem solution, must be translated into a language or code that the computer can understand.

Programming is a complex problem that may involve writing a large number of instructions. It may also involve keeping track of a great many memory cells that are used for instruction and data storage, which is time-consuming and can lead to errors.

To reduce time and the possibility of errors for complex program preparation, the compiler has been developed. The compiler is a program that takes certain commands and then writes, in a form the machine understands, the instructions necessary for a computer to execute these commands. Compilers can bring many instructions into the final program when called upon or signaled by a single source statement. The compiler is problem oriented because the operations produced are those needed to work the problem as set out by the problem statement. Compilers are built at various levels or degrees of complexity. The simplest form of compiler takes one mnemonic phrase and writes one machine instruction. A mnemonic code is an abbreviated term describing something to assist the human memory. For example, to shift the contents of the A-register right nine places, the mnemonic code RSH.A9 is used. This causes

the compiler to write an instruction that shifts the contents of the A-register right 11$_8$ places. A compiler written on this level is commonly called an assembler. Note the advantages: (1) no opportunity to use the wrong function code; and (2) no necessity to convert the shift count to OCTAL.

SUBROUTINES

As a program grows larger, certain functions must be repeated. If the instructions necessary to perform each of these repeated functions are grouped together to form subroutines, these subroutines may then be referenced by a relatively few instructions in the main program. This eliminates repeating certain groups of instructions throughout the program.

EXECUTIVE ROUTINES

The instructions that control access to the various subroutines are called the executive routine of the main program. Depending upon the complexity of the program, there may also be executive subroutines within the executive routines.

Housekeeping is a term used frequently with subroutines. At the time of entry into a subroutine, the contents of the various addressable registers mayor may not be of value. An addressable register is defined as any register whose contents can be altered under program control. The programmer must take steps to preserve the contents of these registers unless they are of no value. This process is termed *housekeeping.*

JUMP AND RETURN JUMP INSTRUCTIONS

The jump and return jump instructions are used in the construction of executive routines. These instructions provide the computer with the ability to leave the sequential execution of the main program or executive routine, execute any of the subroutines stored in its memory, and then return to the execution of the main program.

Execution of a return jump instruction causes the address of the next instruction to be executed in the main program to be stored (usually in the entry cell of the subroutine). It then causes the instruction of the second cell of the subroutine to be executed. The last instruction to be executed will usually be a straight jump to the address contained in the entry cell. Since a jump instruction specifies the address of the next instruction to be executed, the computer is provided with a means of returning to the main program once the subroutine has been executed.

PROGRAM CONSTRUCTION

The process of writing a program is broken down into six basic steps:

1. Statement. A statement forms a clear, comprehensive statement of the problem.

2. Analysis. Analysis consists of laying out the problem in a form that will lend itself to arithmetical and/or logical analysis, determining what logical decisions must be made, and if data manipulation is required.

3. Flow diagram. A flow diagram, or chart, is an expansion of steps in which special symbols are used to represent the various operations to be performed and the sequence in which they are to fall.

4. Encoding. The process of converting the operations listed in the flow chart into language the computer will use; either machine instructions, words, or compiler statements.

5. Debugging. This is the process of locating errors in the program. Various techniques are available for this purpose. A program maybe written to include some aids for itself, or a separate debugging program maybe run to test the operation of a malfunctioning program. For a simple program, a trial solution may be done on paper, and the computed results compared with those actually obtained at each step.

6. Documentation. Documentation is very important because later changes may be warranted in a program or it may be desirable to use subroutines from another program. Proper documentation will ensure that this can be accomplished. Documentation should include the following:

- Program title
- Problem statement
- Programmer's name
- Date
- Memory area used and/or number of cells used
- Registers used
- I/O devices required
- Flow diagram(s)

- Hard copy (program listings, especially a listing of the coded instructions)
- Program tapes

FLOW CHARTING

The programmer constructs a program "map" in determining a solution to a problem. This map is commonly called a flow chart or flow diagram and serves a multitude of important functions. The flow chart "maps" the logical steps required decisions to be reached, and paths to be followed as a result of the decisions. When properly annotated it defines input/output requirements, address allocations, data accuracy considerations, and register usage. A flow chart is valuable to a programmer when "debugging" a program and when making future changes.

Flow charting can be constructed at various levels of complexity. A high-level flow chart consists of a few symbols and presents a broad overview of the problem. A low-level flow chart may approach a one-to-one correspondence between flow chart symbols and program instructions. Usually, there will be several flow charts for a program area. These may be compared to the prints found in a maintenance manual. Maintenance manual prints include a block diagram to show the relationship of the major units (high level), functional block diagrams showing the major circuits in a unit (intermediate level), and the schematics of the circuits (low level). Flow charts should beat such a level that they will implement all the uses previously discussed

MAINTENANCE PROGRAMS

As we have previously stated, a routine or program is a series of instructions that control the operations of a computer. Each instruction is used to cause some action that is part of the overall task the computer must perform. Therefore, an instruction maybe considered as the basic building block of a computer program.

An overall check of a computer can be done by the use of a maintenance program. The maintenance program provides a thorough and rapid method for the detection of failures in a specific portion of a computer. This type of overall maintenance check is flexible and efficient. The programs may use the same type of tape, memory, computing, and external storage media as operational programs. The maintenance program can be altered when the computer or auxiliary components are changed. The program can also be constantly improved. Generally, no extra test equipment is required since the computer circuits are used to perform the test. Testing by means of maintenance programs results in the computer circuits being used in a more comprehensive manner than during normal program execution. When a program has been checked and accepted as a good maintenance tool, it is not subject to deterioration. In contrast, test equipment may be checked and accepted only to become unreliable shortly after being placed in use.

Maintenance programs are divided into three main classes: reliability, diagnostic, and utility programs. Maintenance programs that are used to detect the existence of errors are called reliability programs. Reliability programs should be arranged to check as many computer circuits as possible.

Maintenance programs that are used to locate the circuits in which computer malfunctions originate are called diagnostic programs. An effective diagnostic program should locate the source of trouble as closely as possible. Actually, in many cases, reliability programs have diagnostic features, and diagnostic programs have reliability features. For convenience, a program is called either a reliability or diagnostic program, depending on its intended emphasis. In general, programs that check rather than diagnose are shorter and simpler.

PERIPHERAL AVIONICS SYSTEMS

Learning Objective: *Identify peripheral avionics systems and describe their interaction with the computer.*

The aircraft computer is considered the most important avionics system in achieving the mission of the aircraft. However, the success of the computer depends upon its external sensors or other avionics systems. The quality of data fed to the computer determines the quality of data fed out of the computer. The following avionics systems provide inputs to and receive outputs from the computer: navigation, radar, ordnance/weapons, and data link. These are only a few of the major aircraft avionics systems that interface with the airborne computer; each system is discussed briefly below.

NAVIGATION

Navigation systems are designed to tell pilots where they are, where they have been, and where they are going. The TACAN/DME system provides known station reference points, while an inertial navigation system provides continuous updating of such

information as latitude and longitude. This information is fed to the computer where it is compared, updated, and sent out to other systems.

SEARCH/TRACK RADAR

A search radar system is designed to give visual indications of what is around the aircraft. Some of the present-day aircraft have a 150-mile or greater range. Depending upon the size and/or speed of the radar indications, a computer can determine whether the target is stationary or moving, a landmass or an aircraft, friendly or unfriendly, and many other items of information. If a target is determined to be unfriendly, a tracking radar can be used to tell the pilot what to do to eliminate the unfriendly target.

ORDNANCE/WEAPONS

The design characteristics and ballistics of the many types of ordnance, weapons, and missiles require the use of a computer to store the information. The airborne computer aids the pilot by telling him/her when to release the weapons. The computer greatly increases the pilot's chances of destroying designated targets.

DATA LINK

Combat aircraft have to have the most up-to-date information available to successfully complete combat missions. On an aircraft carrier, the combat information center, CIC, is normally in constant contact with an airborne CIC, usually an E-2 or P-3 aircraft. These two CICs will *crosstalk* by use of the data link system. Basically, data link involves a series of transmitted pulses that represent information. The pulsed information is sent to the computers of all combat aircraft to enhance their chances of success.

ATLAS SYSTEM

Learning Objective: *Identify and understand the ATLAS program to include a test program.*

ATLAS (Abbreviated Test Language for All Systems) was developed by Aeronautical Radio Incorporated as a structured language to provide greater uniformity between test specifications as provided by automatic test equipment (ATE) suppliers. ATLAS defines the requirements of the test in terms of the unit under test (UUT). ATLAS is designed to be easily read and understood by both people and computers. The Navy plans to use ATLAS programs in many of the automatic test equipments, such as VAST, to maintain Navy avionics equipment.

The ATE is provided with an ATLAS compiler/interpreter that translates ATLAS program statements into an intermediate language, which allows high-speed interpretation and reduces the amount of memory required. Characteristics of this interpreter permit the source language to be reconstructed on-line so that program changes can be inserted easily.

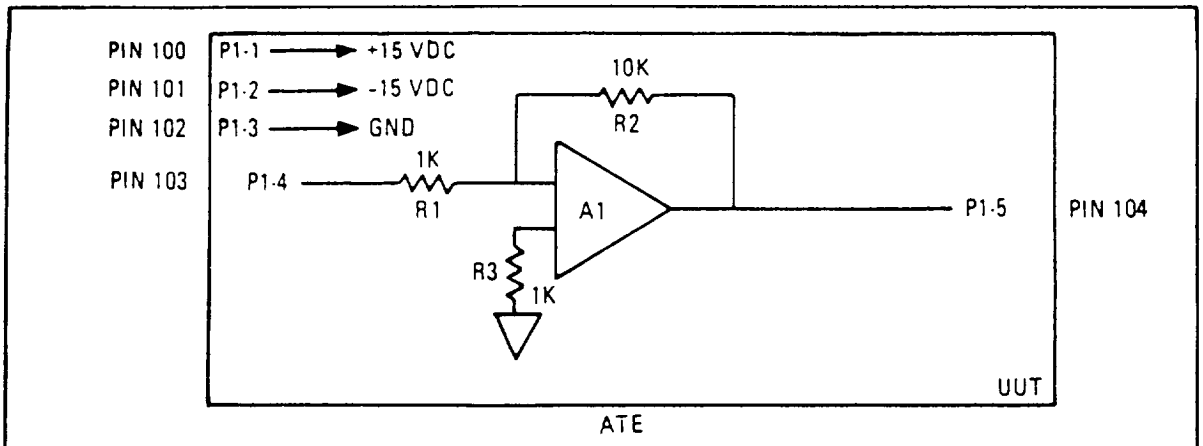
To use ATLAS with ATE, a compiler or interpreter is required to analyze the ATLAS statements and perform the requested actions. A compiler is a program that translates the ATLAS statements into machine language instructions suitable for execution by a computer. A compiler has the advantage of fast execution and reduced computer memory size, but it has the disadvantage of tedious program preparation because of the difficulty in introducing changes. An interpreter is a program that translates and executes ATLAS programs on a statement-by-statement basis. The original ATLAS statements are retained with the interpreter so that programs may be modified on-line. This means that the programmer may generate and verify a program using a keyboard and display conversationally to correct mistakes as they are found. The disadvantages of an interpreter are slower execution and a need for a relatively large computer memory.

SAMPLE TEST PROGRAM

Figure 9-7, view A, shows a schematic diagram for a simple electronic UUT. Figure 9-7, view B, shows a corresponding ATLAS functional test program for the UUT. The UUT is a simple inverting operational amplifier circuit having a nominal gain of 10 and requiring ±15 volts dc power. After a few of the program characteristics are mentioned, a statement-by-statement examination is given.

Program Characteristics

The test program starts when dc power is applied to the UUT. The amplifier input is then grounded, and a check is made at the amplifier output to determine that the output signal is less than 1 millivolt. Any signal in excess of this would have to be due to amplifier oscillations or offset voltage. The program then begins a loop that checks the amplifier gain at 10 different frequencies, ranging from 1 kHz to 10 kHz. The testis terminated if a failure is detected at any point. Test results are displayed at the end.



A

```
 ØØØØØ5 BEGIN, ATLAS PROGRAM $
C0000ıØ   ATLAS DEMONSTRATION TEST PROGRAM $
C     3Ø   DEFINE CONNECTIONS BETWEEN UUT&ATE AND ANTICIPATED RESPONSES $
      4Ø DEFINE, CONNECTIONS, PIN ıØØ= P1-1, PIN ıØ1= P1-2, PIN ıØ2= P1-3,
         PIN ıØ3= P1-4, PIN ıØ4= P1-5 $
      5Ø DECLARE, 'GOOD OUTPUT' (ıØ) $
      6Ø FILL, 'GOOD OUTPUT'(1), .9, .9, .9, .9, .9, .9, .9, .9, .8 $
C000ıØØ   HOOKUP UUT POWER TO ATE POWER SUPPLIES $
      Ø5 APPLY, EARTH, CNX HI P1-3 $
      1Ø APPLY, DC SIGNAL, VOLTAGE 15V, CNX HI P1-1 $
      2Ø APPLY, DC SIGNAL, VOLTAGE -15V, CNX HI P1-2 $
C000200  TEST FOR OSCILLATIONS OR OFFSET $
      1Ø APPLY, EARTH, CNX HI P1-4 $
      2Ø VERIFY, (VOLTAGE-TRMS), AC SIGNAL, LT  1MV, CNX HI P1-5 LO P1-3 $
      3Ø GO TO, STEP  35Ø IF NOGO $
      4Ø REMOVE, EARTH, CNX HI P1-4 $
C     5Ø   TEST GAIN AT 1Ø DIFFERENT FREQUENCIES  $
      6Ø LOOP, THRU STEP  29Ø WITH 'F'= 1 TO 1Ø $
      7Ø APPLY, AC SIGNAL, VOLTAGE  .1V, FREQ 'F'KHZ, CNX HI P1-4 $
      8Ø VERIFY, (VOLTAGE), AC SIGNAL, UL 2-'GOOD OUTPUT' ('F') LL
         'GOOD OUTPUT'('F'), CNX HI P1-5 LO P1-3 $
      9Ø GO TO, STEP  3ØØ IF NOGO $
      94 DISPLAY, RESULT, "TEST COMPLETE, UNIT GOOD" $
      98 REMOVE, ALL $
      99 FINISH $
C000300   FAILURE MESSAGES $
      1Ø DISPLAY, RESULT, "UNIT FAILED GAIN TEST AT"; 'F'; "KHZ" $
      2Ø REMOVE,  ALL $
      3Ø FINISH $
      5Ø DISPLAY, RESULT, "UNIT FAILED OFFSET/OSCILLATION TEST" $
      6Ø REMOVE, ALL $
      7Ø FINISH $
 ØØØ9ØØ TERMINATE, ATLAS PROGRAM $
```

B

Figure 9-7.-Example of a UUT schematic and a corresponding ATLAS test program.

Note that the program uses only capital letters. These are used because the input/output devices (such as the keyboard, CRT display,or printer) use only capital letters.

Note that each statement of the program begins with a number. These numbers are called statement numbers and serve to identify the statement. A program is made up of statements, most of which are instructions to the computer. Statement numbers specify the order in which the statements are to be performed by the computer. Program statements may be typed in any order. Before the program is run, the computer sorts and edits the

program, putting the statements into the order specified by their statement numbers.

Note that the format used in printing step numbers does not print the most significant four digits of the step number unless they differ from those of the preceding step. This implies that the second occurrence of step 10 is really statement 110, and the third occurrence of step 10 is really statement 210. This distinction is extremely important in retrieving program steps to be modified.

After its statement number, each statement starts with an English word. This word is the statement verb. The verbs for statements in ATLAS provide for testing UUTs, performing calculations, controlling the flow of the program, and communicating with the station operator.

Program Examination

The following is a step-by-step examination of the sample test program shown in figure 9-7. Statements 10 and 30 are simply comment statements that have been included to make it easier for a reader to determine the function of the program. Statement 40 defines the connections between the UUT and the automatic test equipment (ATE). Each pin on the ATE has been assigned an integer number, and each pin on the UUT is identified by its normal name presented as a string of characters. This statement tells the test program that P1-1 on the UUT is connected to pin 100 on the ATE; P1-2 on the UUT is connected to pin 101 on the ATE; and so forth.

Statement 50 is used to define the anticipated response in an array variable form. It defines the array name as 'GOOD OUTPUT' and specifies that it has 10 elements. Statement 60 specifies that the first nine of the elements have a value of .9, and the tenth element has a value of .8. These values are then used for limit checking in statement 280 as 10 values of gain expected for 10 frequencies.

Statement 100 is a comment statement explaining the function performed by steps 105, 110, and 120. Statement 105 grounds pin P1-3. Statement 110 applies 15 volts dc to pin P1-1 on the UUT with the 15 volts return connected to ground. Statement number 120 applies minus 15 volts dc to pin P1-2 on the UUT with the return connected to ground.

Statement 200 begins the test for oscillations or offset. Statement 210 applies signal ground to pin P1-4 on the UUT. Statement 200 ascertains that the root-mean-square voltage level on pin P1-5 of the UUT is less than 1 millivolt. Statement 230 directs the program to statement 350 if the test performed in statement 220 results in a NOGO condition. Statement 240 removes the ground on pin P1-4 of the UUT at the completion of the test.

Statement 250 begins a test of ac signal gain with frequencies ranging from 1 kHz to 10 kHz. Statement 260 initializes a program loop extending through statement 290, and specifies that the loop is to be performed with variable 'F' equal to 1 during the first iteration, 2 during the second iteration, 3 during the third iteration, . . . through 10 during the tenth iteration. Statement 270 causes the ATE to apply an ac signal having a voltage of 0.1 volts and a frequency of 'F' kHz to pin P1-4 on the UUT and to apply return to pin P1-8. Statement 280 directs the ATE to measure the rms voltage level on pin P1-5 of the UUT and to ascertain whether or not it is greater than the value of the array variable 'GOOD OUTPUT' ('F'). Statement 290 directs the program to proceed to statement 300 if the preceding test resulted in a NOGO condition. Since statement 290 is within the range of the loops setup by statement 260, control will return to statement 270 unless 'F' is greater than or equal to 10.

If the loop has been completed, program execution continues with statement 294, which displays the message TEST COMPLETE, UNIT GOOD on the ATE CRT. Statement 298 then removes all connections to the UUT, and statement 299 returns control to the test station operator.

In the event that any of the tests resulted in a failure, the program would branch as directed in statements 230 or 290 to the appropriate error message statement. The branch to statement 310 results in a displaying of UNIT FAILED GAIN TEST AT, the value of 'F', kHz on the CRT. Statement 320 removes all connections to the UUT, and statement 330 returns control to the test station operator. The branch to statement 350 results in a display of the message UNIT FAILED OFFSET/OSCILLATION TEST on the CRT. Statement 360 removes all connections to the UUT, and statement 370 returns control to the test station operator.

Each of the statements is important to the proper operation of the test. For example, what would have happened if statements 110 and 120 had been omitted? The UUT would have had no power applied so that it would pass the test for oscillations or offset, but it would fail the gain test. Failing to include statement 240 would have resulted in an error when step 270 was executed since this would represent an attempt to apply an ac signal to a pin on the UUT that previously had been

grounded. This error would be detected by the ATE, and would result in an error message being displayed and the test being terminated. If statement 294 was omitted, the operator would have no assurance that the test program completed all the tests.

The particular choice of statement numbers is arbitrary, as long as the statements are numbered in the order in which they are to be executed in stepping through the program. The statements could have been numbered 1, 2, 3, . . . , 31, although consecutive numbering is not recommended. Gaps are left between statement numbers so that it will be possible later to insert additional statements to make corrections to the test program. If it becomes necessary to add two statements between those numbered 210 and 220, they can be given any two numbers between 210 and 220; for example, 214 and 216. In the editing and sorting process, the computer will put them in their proper place in the program. If it is desirable to segregate independent tests, it can be done easily by starting each test with a statement ending in 00. This makes it easier to examine the program and determine which steps are used in any single test. A comment statement describing the test to be performed is considered good form.

ATLAS STATEMENT COMPONENTS

ATLAS statements contain various components, including characters, numbers, labels, variables, and arrays. These components are discussed below.

Authorized Characters

The following are the only authorized characters for use in ATLAS statements.

Uppercase letters A through Z

Numbers 0 through 9

Miscellaneous symbols:

Asterisk	*	Greater than	>
Comma	,	Not Equal (pound sign)	#
Period	.	Currency Symbol	$
Equal	=	Slash	/
Plus	+	Right Parenthesis	(
Minus	-	Left Parenthesis	)
Hyphen	-	Apostrophe	'
At symbol	@	Double Quote	"
Less Than	<	Blank Space	
Semicolon	;	Percent	%

When it is necessary to identify an alpha character literally as lowercase, such as in a manufacturer's pin designation, a slash(/) is written preceding each affected character. The minus sign and hyphen are represented by the same character and will be interpreted according to the context of the statement. In this document, the apostrophe (') is frequently referred to as a single quote mark in connection with its usage with labels.

Numeric Representation of Constants

The ATLAS interpreter uses two basic types of internal numeric representations for numbers. Numbers are classified as being either real or bit pattern. Real numbers are usable for all types of arithmetic computations. Bit pattern numbers are used for stimulating and checking the responses of digital UUTs. The rules for expressing numbers in ATLAS program are listed below.

1. Real numbers may be written as decimal numbers, with or without a decimal point, or they may be expressed in exponential notation for convenience in representing very large or very small numbers. Real numbers in decimal form maybe written with a plus or minus followed by the string of characters 0 through 9, with the decimal point placed in the conventional reamer. Numbers written without the prefix plus or minus are accepted as positive. (Examples: -5268, 28.00, +55.) Real numbers may also be written in the exponential form ±n.m E ±p where n, m, and p are numeric strings of up to six digits. (Examples: 1E6, 3E + 8, 1.72E - 19.) If m is zero, the decimal point maybe omitted. Real number accuracy is limited to six significant digits with a range of ±10 to the 38th power.

2. Bit patterns may be expressed in binary, octal, hexadecimal, or character string forms. A sign preceding the number has no meaning for bit patterns. Conversion from the decimal number form and several bit pattern representations are shown in table 9-1.

3. Binary numbers are written with the letter B followed by the appropriate string of the numbers 0 and 1 enclosed in single quotes. (Example: B'0110010'.) Up to 528 binary bits are permitted.

4. Octal numbers are written with the letter O followed by the appropriate string of numbers 0 through 7 enclosed in single quotes. (Example: O'701340'.) Up to 176 digits are permitted.

Table 9-1.-Number Conversions

DECIMAL	BINARY	OCTAL	HEXADECIMAL
0	B'0000'	O'00'	X'0'
1	0001	01	1
2	0010	02	2
3	0011		3
4	0100	04	4
5	0101	05	5
6	0110	06	6
7	0111	07	7
8	1000	10	8
9	1001	11	9
10	1010	12	A
11	1011	13	B
12	1100	14	C
13	1101	15	D
14	1110	16	E
15	1111	17	F

5. Hexadecimal numbers are written with the letter X followed by the character string using 0 through 9 and A through F, which represents the hexadecimal value enclosed in single quotes. (Example: X'53A5D2'.) Up to 132 hexadecimal digits are permitted.

6. The character string configuration is written with the letter C followed by any string of characters enclosed in single quotes except the apostrophe and currency symbol characters, which may NOT be used. A character string is limited to 66 characters maximum. (Example: C'2A/='.)

Labels

Unique character strings enclosed in single quotes are called labels. They are required in certain statements to provide unique identifiers within the program. A label may contain up to 62 characters. Blanks are significant in a label. Labels may be selected at the writer's convenience, and if properly selected, they will improve the readability of the program. Labels maybe formed by any combination of the authorized characters except the apostrophe ('), currency symbol ($), at symbol (@), and double quote (").

Variables

Computer memory locations are made available to the ATLAS test programmer for storage of measured values, anticipated values, and intermediate results of arithmetic computations. These locations are designated as variables, and are identified by variable names, which may consist of up to 62 alphanumeric characters. These names are enclosed in single quotes and maybe assigned uniquely by the programmer. Each variable may be used by the programmer to store real, octal, binary, hexadecimal, or character string forms of data. Variable locations that are to receive more than 32 bits of data must be declared larger in a DECLARE statement. Nondeclared variables larger than 32 bits have the most significant bits truncated to 32 bits.

Two forms of variables are provided: simple variables and array variables. A simple variable is a single memory location designated by a single name such as 'NAME', whereas an array variable is a group of memory locations assigned a common name. All variables are set undefined at the start of execution of an ATLAS test program. The variable must be defined

or assigned a value somewhere within the test program. Use of an undefined variable will result in a terminating error message when the program is executed. The value of a program variable is assigned by the program as it is executed, and it may change as new values are obtained. The value of a program constant is assigned by the test programmer. It remains fixed as a number during the program execution.

The variable, known as 'MEASUREMENT' (abbreviated 'ME'), is a permanent, preassigned identifier with special significance to the ATLAS interpreter. This variable will have its value set to the value of the measured characteristic by all READ, MEASURE, MONITOR, and VERIFY statements. The value contained in 'MEASUREMENT' may then be used in any subsequent statements. 'MEASUREMENTS' may be used anywhere in the program where a real valued variable may appear. It may be assigned a value through a CALCULATE statement, although the user must be aware that its value will be reassigned by subsequent execution of any statement that returns a measured value. The value of 'MEASUREMENT' may be retained beyond the time it would otherwise be lost through execution of another sensor statement by assigning its value to another variable through the use of a CALCULATE statement. The inclusion in a statement of a RESULT modifier followed by a variable name causes the value of 'MEASUREMENT' to be stored as the value of the variable upon completion of statement execution. This eliminates the need for a separate CALCULATE statement to perform this assignment. An example would be as follows;

```
30 MEASURE, (RES), IMPEDANCE, RESULT
   'IN-Z', CNX HI J1-40 LO J1-45 $
```

Arrays

In addition to the simple variables used by ATLAS, there are variables that can be used to designate the elements of an array. These are used where a subscript would ordinarily be used; for example, the coefficients of a polynomial (A_0, A_1, A_2, . . .) or to designate the elements of a vector.

The array variables used in ATLAS consist of the array name enclosed in single quotes followed by the subscript enclosed in parentheses. The subscript may of course be a formula. Thus, the programmer might write 'A'(0), 'A'(l), 'A'(2), and so forth, for the coefficients of the polynomial mentioned above. Individual elements of the array may be accessed by referring to the array name followed by a value enclosed in parentheses, such as 'A'(4). The value in parentheses may range from 0 to the maximum number of elements that the array may contain. The number of elements in the array must be specified at the start of the ATLAS test procedure through a DECLARE statement.

The array 'A'(0), 'A'(1), 'A'(2),. . . 'A'(10) can be entered into the program by writing the following:

```
000010 DECLARE 'A'(10)$
```

This statement identifies the variable 'A' as an array having 11 elements, which maybe referred to as 'A'(0), 'A'(1), . . . 'A'(10). A subsequent FILL statement may then be used to load the array, such as the following:

```
000020 FILL 'A'(0),2,4,6,8,10,9,8,7,6,5,5 $
```

This statement loads the array with the values shown so that 'A'(0) = 2, 'A'(1) = 4, . . . and 'A'(10) = 5. If the array is not to be preloaded with data, the FILL statement can be omitted. Note that the variable 'A' may no longer be referred to as a simple, unsubscripted variable.

Arithmetic Calculations

The ATLAS interpreter provides a powerful facility for performing arithmetic calculations and storing the results in variable locations through use of the CALCULATE verb. A single CALCULATE statement may be used to set a number of variables equal to a single value, a number of variables equal to a number of values, or any combination of these. Some examples of permissible CALCULATE statements are as follows:

```
000010 CALCULATE, 'A' = 'B' = 'C' = 3$

    20 CALCULATE, 'A' = 4, 'B' = 7, 'C' 7 = 19$

    30 CALCULATE, 'A' = 'B' = 4, 'C' = 17, 'O' =
                  5*(17.2 + (TAN('P')))**4$

    40 CALCULATE, 'E' = 'A', 'D' = 'E' *4$
```

In a CALCULATE statement, the value on the right side of the equal sign is stored in the memory location specified by the variable on the left side of the equal sign. This variable then retains that value until it is changed by execution of some subsequent statement. In statements involving multiple-equivalence operations separated by commas or semicolons, the operations are performed in left to right order. Parentheses are used to indicate the order of reduction of the arithmetic operations and establish priorities for the calculation. If there is any question about the priority, add more parentheses to eliminate possible ambiguities. The order of priorities is summarized in the following rules:

Table 9-2.-ATLAS Arithmetic Functions

SYMBOL	EXAMPLE	MEANING
+	'A' + 'B'	Addition (Add 'A' to 'B')
-	'A' - 'B'	Subtraction (Subtract 'B' from 'A')
*	'A * 'B'	Multiplication (Multiple 'B' by 'A')
/	'A'/'B'	Division (Divide 'A' by 'B')
**	'A' * * 2	Exponentiation (find $'A'^2$)

Table 9-3.-ATLAS Language Mathematical Functions

FUNCTIONS	INTERPRETATION
SIN('X')	FIND THE SINE OF 'X' (IN DEGREES)
COS('X')	FIND THE COSINE OF 'X' (IN DEGREES)
TAN('X')	FIND THE TANGENT OF 'X' (IN DEGREES)
ATAN('X')	FIND THE ARC TANGENT OF 'X' (RESULT IN DEGREES)
EXP('X')	FIND $E^{'X'}$

1. The formula inside the parentheses is computed before the parenthesized quantity is used in further computations.

2. In the absence of parentheses in a formula involving addition, multiplication, and exponentiation, the computer first performs the exponentiation, then performs the multiplication, and then addition comes last. Division has the same priority as multiplication, and subtraction the same as addition.

3. In the absence of parentheses in a formula involving operations of the same priority, the operations are performed from left to right.

The computer performs arithmetic calculations by evaluating formulas that are supplied in the program. These formulas are very similar to those used in standard mathematical calculations. Five arithmetic operations can be used to write a formula. These are listed in table 9-2.

Logical operators may be used to operate on bit string numbers in a CALCULATE statement. The logical operators available are AND, OR, EXOR (exclusive OR), and NOT. For example:

```
CALCULATE, 'D' = 'A' AND 'B', 'C' (3) = 'D'(1)
OR 'D' (2), 'F'('X') = NOT 'G' ('X'), 'Z' = 'M'
EXOR 'N' $
```

Three rules pertaining to a CALCULATE statement are given below.

1. Mixed mode calculations (calculations involving both real and bit string numbers) are not allowed.

2. CALCULATE is the system default verb and is assumed if no verb is present in the statement.

3. UUT pin numbers may be assigned or reassigned to ATE pins with a CALCULATE statement of the following form:

```
000100 CALCULATE, @PI-A@ = 5 $
```

In the above example, UUT pin PI-A has been reassigned to ATE pin 5. Any subsequent statements involving PI-A in the CNX list will operate on ATE pin 5. The UUT pin identifier in the CALCULATE statement is always bracketed with two @ symbols. Pins defined in CALCULATE statements should not be included in DEFINE, CONNECTIONS statements.

Mathematical Functions

In addition to the five arithmetic operations and logical operators, the computer can evaluate a number of mathematical functions. These functions are given the special names shown in table 9-3. (NOTE: For a complete listing of ATLAS language mathematical

```
C000050  THESE ARE SAMPLE ATLAS STATEMENT $
000100   APPLY, DC SIGNAL, VOLTAGE 5V, CNX HI J1-2 LO J1-3 $
    10   APPLY, AC SIGNAL, VOLTAGE 9V, FREQ 1 KHZ, CNX HI J2-3 I.O J2-4 $
    30   APPLY , SHORT, CNX HI J1-5 J1-6 JI-7 $
000200   MEASURE, (VOLTAGE), AC SIGNAL, CNX HI J1-2 LO J1-3 $
    10   MEASURE, (FREQ), AC SIGNAL, CNX HI J1-2 LO J1-3 $
    20   VERIFY, (VOLTAGE), DC SIGNAL, GT 19V, GO-TO-STEP 700, CNX HI J1-19 LO J1-20 $
    30   MEASURE, (RES), IMPEDANCE, RESULT 'IN-Z', CNX HI J1-40 LO JI-45 $
000300   REMOVE, SHORT, CNX HI J1-5 J1-6 J1-7 $
    10   REMOVE, ALL $
```

Figure 9-8.-Example of test-oriented ATLAS statements.

functions, refer to the HATS Technical Manual, NAVAIR 17-15CAJ-48.3.)

ATLAS TEST PROGRAM STATEMENT CONSTRUCTION

An ATLAS program contains one statement for each action to be performed. The actions are specified by ATLAS verbs, such as BEGIN, APPLY, MEASURE, DISPLAY, CALCULATE, REPEAT, and TERMINATE. If the verb corresponds to a UUT stimulus or measurement, nouns and modifiers are used to define the signal. AC SIGNAL and DC SIGNAL are examples of ATLAS nouns. Modifiers are used to specify parameters or qualities of a signal. VOLTAGE, DISTORTION, FREQ, and RISE-TIME are typical noun modifiers. Each stimulus or measurement statement also contains a connection field specifying the points on the UUT to which a signal is to be applied or measured.

An ATLAS test program consists of a series of valid ATLAS statements. These statements are ordered by statement number, and execution proceeds in the sequence of ascending statement numbers. Each statement has three fixed fields, which are followed by a set of statement peculiar fields. The first three fields of every statement are the flag, statement number, and verb. The flag and statement number fields may be null. The remaining fields are variable in composition, length, and location, depending upon the type of statement being written. The general construction of ATLAS statements, including fields, is given below.

ATLAS Statements

Test-oriented statements in ATLAS have the following general construction:

fstatno <verb> (<measured characteristic>), <nown>, <statement characteristics> <cnx field>$

Examples of ATLAS statements are listed in figure 9-8. As can be seen in the figure, each statement starts out with a flag field, f, which is followed by the statement number field, statno. A verb following the statement number identifies the function performed by the statement.

There is a 1022-character limit to the length of an ATLAS statement. Completing the statement on one line is not necessary. By inserting spaces between statement elements, the writer can begin a new line at any point in the statement, with subsequent lines indented if desired. Extra spaces cannot be inserted in the middle of the ATLAS language syntax elements, as defined in table 9-4. The fixed fields (Flag, Statement Number, and Verb) must be written on the first line, as required by the format.

ATLAS Statement Fields

These fields include flag field, statement number field, verb field, measured characteristic field, noun field, statement characteristic field, and CNX field.

Table 9-4.ATLAS Statement Elements

VERBS

APPLY (AP)
ASSIGN
BEGIN
CALCULATE
COMMON (C0)
COMPARE (CP)
CONNECT (CT)
DECLARE (DEC)
DEFINE (DEF)
DELAY
DELETE
DISCONNECT (DT)
DISPLAY (DI)
DO
ED
END
EOF
EXECUTE
EXTERNAL
FETCH (F)
FILL
FINISH (FI)
GO TO (GOTO)
INDICATE (IN)
LIST (L)
LOOP
MEASURE (ME)
MONITOR (MO)
PERFORM (P)
PLIST
PRINT (PR)
PTAPE
READ
REMOVE (RE)
REPEAT (REP)
REPLACE
RUN (R)
SAVE
SCRATCH (SCX)
SETUP (SE)
STEP AT (ST)
TERMINATE (TE)
TLIST
TRAP
VERIFY (VE)
WAIT FOR (WF)

NOWNS

AC SIGNAL (AC)
ALL
DC DIGNAL (DC)
DIGITAL TEST (DIG)
EARTH
EVENTS
IMPEDANCE
PULSED DC
RAMP SIGNAL
SHORT (SH)
SQUARE WAVE SIGNAL (SWS)
SYNCHRO
TIME INTERVAL
TRIANGULAR WAVE SIGNAL (TWS)

FLAG MODIFIERS

AC-COUPLED
+CLOCK
-CLOCK
CONTINUOUS (CONT)
HIGH-ACCURACY (HIAC)
HIGH-FREQUENCY (HIFQ)
IF-GO
REF-POS-SLOPE (RPS)
REF-NEG-SLOPE (RNS)
RESP-COMP (RSC)
RESP-SAVE (RSS)
SINGLE
+SLOPE
-SLOPE
STIM-ONLY (STO)
STIM-RESP-COMP (STRC)
STIM-RESP-SAVE (STRS)
+SYNC
-SYNC
TYPE-PARALLEL (TYP)
TYPE-SERIAL-LSB-FIRST (TYSL)
TYPE-SERIAL-MSB-FIRST (TYSM)
REPETIVE

LIST MODIFIERS

ERROR (ERR)
ERROR-INDEX (ERRI)
GO-TO-STEP (GTS)
MASK-ONE (MAO)
REF
RESP
STIM

MODIFIERS

ANGLE
APERTURE
AVERAGE
BIT-RATE (BIR)
CAL
CLOCK-DELAY
COUNT
CURRENT (CU)
DC-OFFSET (DCO)
DELAY
DISTORTION (DIST)
FALL-TIME (FT)
FAULT-COUNT (FC)
FILTER
FREQ (FR)
LEVEL
MASK-ONE (MAO)
MAX-TIME (MT)
NAME
NEG-SLOPE
NOISE
NOM
OVERSHOOT
PERIOD
POS-SLOPE
PRF
PULSE-WIDTH (PW)
RATE
REF-LEVEL (RLV)

MODIFIERS–CONTINUED

REF-VOLTAGE-MAX (RVM)
RES
RESP-DELAY (RSD)
RESULT
RISE-TIME (RT)
SAMPLES
SYNC-LEVEL
TIME
UNDERSHOOT
VOLTAGE (VO)
VOLTAGE-AV (VV)
VOLTAGE-COMP (VOP)
VOLTAGE-MAX (VX)
VOLTAGE-ONE (VONE)
VOLTAGE-P (VP)
VOLTAGE-PP (VPP)
VOLTAGE-TRMS (VT)
VOLTAGE-ZERO (VZER)
WORD-LENGTH (WORL)
WORD-RATE (WOR)

LASAR

DATA-FILE (DF)
FAULT-MESSAGE (FTM)
FAULT-SET (FS)
IF-FAULT-SET (IFS)
LASAR-CHECK (LC)
LASAR-TEST (LAT)
MISMATCH (MM)

MISC

AND
CONNECTIONS (CON)
ELEMENTS
EXOR
GO-NOGO
MANUAL-DATA
MESSAGE (MES)
NOT
OR
PIN
PROCEDURE
QUERY (Q)

DEBUG FLAGS

LOC	
0277	OUTPUT DEVICE 0 = NONE, 1 = TTY, 2 = CRT, 3 = T.P., 4 = D.P., 5 = VOGUE
0300	NEGATIVE FOR '/SHORT TEST' DATA
0301	ADDRESSES
0302	-1 FOR EXC'S OAR'S
0303	FIRST SUB ADDR.
0304	LAST SUB ADDR.
0307	-1 FOR NOM. CAL.

CNX DESIGNATIONS

A
B
C
CNX
CNX-RESP
CNX-STIM
COMPL
HI
LO
N
REF-A
REF-B
REF-C
REF-HI
REF-LO
REF-N
SYNC
TRUE
X
Y
Z

DIMENSIONAL UNITS

A
BITS
DEG
GHZ
HZ
KHZ
KOHM
KPPS
KV
MA
MHZ
MIN
MOHM
MRAD
MSEC
MV
NSEC
OHM
PPS
PSEC
RAD
SEC
UA
USEC
UV
V

FLAG FIELD.– The first field of each statement is the flag field. It is one character in length and is located in the first column of the first row of each statement. The field must be used for a B, C, or E entry or left blank.

B Entry.– B in the flag field indicates the destination statement of a GO TO statement from elsewhere in the procedure. Characters following the statement number and preceding the $ are commentary and are ignored during translation.

C Entry.– C in the flag field indicates that information in that statement is a comment to be ignored during translation. Characters following the statement number and preceding the $ are treated as commentary and are ignored during translation.

E Entry.– E in the flag field indicates entry points where the test conditions arc completely stated and are not dependent on previous tests in any way. Characters following the statement number and preceding the $ are treated as commentary and are ignored during translation.

STATEMENT NUMBER FIELD.– The second field is the statement number. It is six digits long and provides a reference designator for each program statement. The first four digits of the number are called "test numbers," and the remaining two digits are called "step numbers." Each succeeding program step is assigned a higher number than the preceding one, but it is neither necessary nor advisable to use the next higher number. If the test number of a statement is the same as the previous statement, the first four characters of the statement number may be left blank. If no statement number is provided, the interpreter will execute the statement immediately upon successful completion of the syntax check. The statement will not be included in the memory resident program. The largest statement number is 32767.

VERB FIELD.– Thc third field is the verb field. Its length is variable depending on the entry to be made. Every ATLAS statement must have an entry in this field, except that if no entry is present, CALCULATE is assumed. The statement number and verb fields are separated by one space. A verb may include a mandatory space as an integral character of the word, such as in WAIT FOR. The verbs that can he used on the ATE are listed in table 9-4.

MEASURED CHARACTERISTIC FIELD.– A measured characteristic field is included in sensor-type statements to specify which of the available characteristics is to he evaluated by the sensor function. The field is positioned between the verb and noun in the statement and consists of one of the modifier mnemonics from the list associated with the noun. The characteristic to be measured is always enclosed in parentheses. Typical statements are as follows:

```
000200 MEASURE, (VOLTAGE), AC SIGNAL,
       CNX HI J1-2 LO J1-3 $

    10 MEASURE, (FREQ), AC SIGNAL,
       CNX HI J1-2 LO J1-3 $

000650 VERIFY, (VOLTAGE), DC SIGNAL, UL
       10V LL 8V, CNX HI J1-1 LO J1-2 $

000690 MEASURE, (VOLTAGE), DC SIGNAL,
       CNX HI J1-4 LO J1-6 $
```

Notice that evaluation data such as upper and lower limits may be needed in statements that contain this field.

NOUN FIELD.– The noun is a word or words that define(s) the electrical or physical element upon which the verb is to act. The available electrical or physical elements are dependent upon the station hardware configuration. The nouns that can be used on an ATE are listed in table 9-4.

STATEMENT CHARACTERISTIC FIELD.– The statement characteristic field follows immediately after the noun field. The information in this field identifies additional modifiers required by the interpreter to further define a signal that is to be applied or a measurement that is to be taken. Entries in the statement characteristic field consist of the name of a modifier such as FREQUENCY or VOLTAGE followed by a value that specifies the magnitude of the modifier, Some types of modifiers (such as +SLOPE) do not require a value to be specified. The value of the modifier may be indicated by a 'label', which refers to a variable established by another statement in the program. A dimensional unit, such as Hz or kHz, may be appended to the modifier if desired. It has the effect of multiplying the value of the quantities by the required factor to convert the value into basic units, as all modifiers are handled internally in basic units (such as volts, hertz, and ohms). Modifiers may be entered in any order. Lists of modifiers and dimensional units that are legal are provided with the ATE hardware.

CNX FIELD.– The last field of a test program identifies the UUT pins to which a stimulus or load is to be applied or from which a measurement is to be taken. The field is separated from preceding fields by a comma

and the word CNX. The field contains a series of assignments of ATE functions to UUT pins. Each ATE function, such as a stimulus output or measurement input, has associated designators, which must be used within the CNX field. These are listed in table 9-4. Examples are HI, LO, X, Y, Z, N. An assignment of a function is accomplished by entering one of these designators followed by one or more UUT pin designators. The UUT pin designators used are generally identical to those used by the SRA manufacturer, such as J1-2 or P2-AA. The UUT pin designator may be as long as 62 characters using an arbitrary sequence of ATLAS characters excluding blanks, apostrophes, currency symbols, commas, @ symbols, and quotation marks. To prevent ambiguities in the statement, UUT pin designators may not be identical to ATE function designators, such as HI, LO, A, B, and C. The slash character (/) indicates that the next character, which must be alphabetic, is lowercase; for example, J/A-2 = Ja-2.

Example: 001000 APPLY, DC SIGNAL, VOLTAGE 5V, CNX HI J1-2 LO J1-3 $

The physical connection between the UUT pins and the station interface connector pins is defined during program execution by a DEFINE CONNECTIONS statement in the preamble to the program or in a CALCULATE statement.

Other ATLAS Statement Characteristics and Requirements

These include terminators, separators, and blank space requirements.

**STATEMENT TERMINATORS ($).–** The last character of every ATLAS statement must be the currency symbol ($).

FIELD SEPARATORS.– Since the fields following the verb are variable, a separator is used to identify the start of each new major field. A comma preceded and/or followed by optional spaces is designated for this purpose.

BLANK SPACE REQUIREMENTS.– Generally, blank spaces are not significant in the language, and their presence or absence in a statement will not affect the correctness of the ATLAS program. However, blanks are required to separate UUT pin designators within a CNX field. It is permissible to include any number of blank characters between any two language elements. A language element is any word, character, or group of words that is defined to have significance. Any number is a language element. Within language elements, the characters are fixed as defined. WAIT FOR, for instance, can only be written as an eight-character word, including the one blank space. LT may only be written as two characters with no space, but any number of spaces may follow WAIT FOR or LT.

ATLAS PROGRAM CONSTRUCTION

A complete ATLAS test program is a series of ATLAS language statements divided into two distinct sections referred to as the preamble and the procedure. The preamble section contains all the DEFINE, DECLARE, and COMMON statements and procedure definitions. The procedure section then uses the elements defined in the preamble section in performing its assigned tasks. Note that there is no statement that separates the preamble and procedure sections.

Preamble Section

The preamble section of an Atlas program precedes the procedure section, and it contains the BEGIN statement, all of the COMMON, DECLARE, and DEFINE statements, and the procedure (subroutine) definitions. FILL statements may also be included in the preamble. Preamble statements do not cause any tests to be executed, but the information described in the preamble may be repeatedly referenced in the procedure section.

BEGIN STATEMENT.– This statement designates the first statement of a complete ATLAS program. For example:

00100 BEGIN, ATLAS PROGRAM $

The BEGIN statement does not cause any action and is used only to designate the beginning of a program. It is not mandatory, but is considered good programming practice.

DEFINE, 'PROCEDURE NAME', PROCEDURE, STATEMENT.– This type of DEFINE statement assigns a name to one or more complete ATLAS statements, which form an executable subroutine. The defined series of statements is then executed when the 'procedure name' is cited in a PERFORM statement in the procedure section of the program. If any input data or results of the subroutine execution must be passed between the programs, a

parameter list must be provided. The word RESULT is used to identify the parameters that will return data from the subroutine. Parameter lists are not required to name a subroutine, but there is no facility for adding to a parameter list when the subroutine is performed. If used, parameters declared in the DEFINE statement will refer to the corresponding actual parameter in the perform statement. The following are sample statements:

```
000020  DEFINE, 'GENERATOR START',
        PROCEDURE $

000030  DEFINE, 'RELAY TEST', PROCE-
        DURE ('INPUT', 'VOLTAGE')
        RESULT ('GOOD')  $
```

DEFINE, CONNECTIONS, STATEMENT.– This type of DEFINE statement defines the physical connections between the ATE and the UUT at the interface adapter. Each entry in the pin definition list identifies a UUT pin using the name supplied by the equipment manufacturer and specifies the pin to which the UUT pin is mated or connected. For example: If UUT pin J1-4 is connected to ATE pin 17, write DEFINE, CONNECTIONS, PIN 17 = J1-4. A typical statement is as follows:

```
000010  DEFINE, CONNECTIONS, PIN 5 =
        J1-1, PIN 12 = J1-2, PIN 19=J1-3 $
```

All ATE interface pins and many ATE internal points are assigned pin codes from 0 to 32767. All UUT pins are identified with the pin designation supplied by the UUT manufacturer with these limitations:

1. A pin name maybe represented by an arbitrary sequence of ATLAS characters excluding blanks, apostrophes, currency symbols, @ symbols, commas, and quotation marks. The slash character (/) indicates that the next character is lowercase. For example, J/A-2 =Ja-2.

2. No more than 62 characters, including any slashes (/), are allowed in a pin name.

3. A pin name may not duplicate an ATE function designator such as HI or LO.

Each UUT pin used in the ATLAS program is defined in a DEFINE, CONNECTIONS, OR CALCULATE statement. Use of undefined pins will result in a terminating error, as the interpreter is unable to determine the signal routing required for the statement.

DECLARE STATEMENT.– This statement specifies the memory allocation to be used in storing array variables and those simple variables that must contain more than 32 bits of information. The first element of each array is element zero. The last element of the array is designated by a number in parenthesis in the DECLARE statement. A typical DECLARE statement is as follows:

```
000020  DECLARE, 'A'(10), 'B'(5), 'C', 61
        BITS $
```

This statement declares that array 'A' consists of 11 elements, and array 'B' consists of 6 elements. It further specifies that all elements in arrays 'A' and 'B' may contain up to 61 bits of information; additionally, the simple variable 'C' may contain up to 61 bits of information. Any number of arrays or simple variables may be declared in a single DECLARE statement as long as they must all contain the same number of bits of information. If the number of bits is not explicitly stated in the DECLARE statement, a default of 32 bits is assumed. It is unnecessary to declare simple variables for which 32 bits or less of storage will suffice. It is wasteful of computer storage to declare arrays larger than necessary or to specify more bits than are actually required by the variable.

COMMON STATEMENT.– This statement designates variables that are to be used in common by several ATLAS programs. The format is identical to the format used in DECLARE statements except that storage is allocated for variables specified in COMMON statements in a special area of memory. Array and variable size declarations specified in COMMON statements are processed exactly as they are in DECLARE statements. Arrays of variables that are specified in COMMON statements are processed exactly as they are in DECLARE statements. Arrays of variables that are specified in COMMON statements should not be specified in DECLARE statements. All COMMON statements must precede all DECLARE statements in any given ATLAS program. Specifying variables as common variables permits their values to be transferred between ATLAS program overlays. The names of the variables have no significance, and values are passed strictly in the order in which they appear in the COMMON statement. As a rule, however, all related program segments will have identical COMMON statements.

An example of a parameter passing through COMMON is shown in the below programs.

```
000010 COMMON, 'A' 'B'(3), 'C' $

20   FILL, 'B'(1), 1,2,3 $

30   CALCULATE, 'A' = 5, 'C' = 7 $

40   EXECUTE, 'NXT-PROG' $

50   FINISH $

PROGRAM 'NXT-PROG'

000010 COMMON, 'A', 'B'(3), 'D' $

20   PRINT, RESULT, 'A' 'B'(1), 'B'(2),
     'B'(3), 'D' $

30   FINISH $
```

The printed result appears as follows:

```
5 1 2 3 7
```

Note that in this case, the executing program defines the three variables 'A', 'B', and 'C' in common. 'B' is an array defined in statement 20. Statement 30 sets 'A' equal to 5, and 'C' equal to 7. Statement 40 executes the program named 'NXT-PROG'.

A listing of program 'NXT-PROG' shows three variables, 'A', 'B', 'D', in common. 'B' is defined as a list of three elements whose values are specified. The program prints the value of the three variables it receives (actually five values because of the array) and terminates. From the printed result, it can be seen that the program received the correct variable values from the EXECUTING program. The use of variable name 'D' in the second program shows that the names of the variables have no significance. Such redefinition of names through COMMON statements is not recommended because it maybe confusing to a person attempting to understand the program.

FILL STATEMENT.– This statement can be used to load an array with data specified in the FILL statement. FILL statements may be included in either the preamble or procedure section of the ATLAS program. FILL statements located in the preamble section of the ATLAS program will be executed before execution of the program is started. Any FILL statement encountered in the normal flow of the program will also be executed at the time it is encountered. This allows array elements to be modified through use of FILL statements during program execution.

Each fill statement may be used to load the elements of a single array only. The FILL statement specifies the initial array element that is to be loaded. Data values following the first value go into subsequent array elements. Data values may be of type real, binary, octal, hexadecimal, or character string. Some typical FILL statements are as follows:

```
000010 FILL 'A'(1), 5, 6.3, B'101', 7 $
000020 FILL 'B'(7), 3, 2, 1 $
```

In these examples, statement 10 fills elements 1 through 4 or array 'A' with the four data values specified in the statement; statement 20 tills elements 7 through 9 of array 'B' with the three data values specified in the statement.

Procedure Section

The procedure section is the main body of an ATLAS test program. It consists of a series of statements that are executed in sequence. All statement types are legal in the procedure section except those using the verbs DEFINE, COMMON, and DECLARE. Statements in the procedure section may refer to lists, UUT pins, or procedures, which are defined in the preamble section. Each statement describes a portion of the required test that must be completed before the next statement.

As stated above, execution of an ATLAS test program normally proceeds by executing statements in strict numerical order. However, this sequence may be altered through execution of a statement containing a GO TO verb or a GO-TO-STEP modifier. The sequence may also be changed by the programming technique known as a loop. (Loops will be discussed in detail later.) Such statements are used to branch the test program to some statement other than the one next in sequence. The program branch may be taken on either an unconditional or a conditional basis, depending on the structure of the statement.

UNCONDITIONAL BRANCHING.– In the following sample statement, line 10 is an example of the unconditional branch, the simplest type of GO TO statement. The test sequence is diverted as indicated every time the statement is executed. For example:

```
000010 GO TO, STEP 210 $
```

Conditional Branch on Arithmetic Comparison.– The following examples illustrate the

use of an arithmetic comparison as a condition on a GO TO statement. Formulas of any complexity may appear on either side of the arithmetic comparison symbol. Only the symbols = and # may be used for bit string comparisons. The branch will be taken only if the specified condition is true. In statement 110, the branch will be taken only if' 'B' is greater than or equal to 93. For example:

```
000100 GO TO, STEP 330 IF 'A' > 27 $
    10 GO TO, STEP 340 IF 'B' > = 93 $
    20 GO TO, STEP 350 IF 'C' < = 18 $
    30 GO TO, STEP 360 IF 'D' < = 84 $
    40 GO TO, STEP 370 IF 'E' = 16 $
    50 GO TO, STEP 380 IF 'F' # 14 $
    60 GO TO, STEP 390 IF 'G' = B'101 $
    70 GO TO, STEP 400 IF 'H' # B'101' $
```

Conditional Branch After COMPARE or VERIFY Verb.– A field for an evaluation characteristic is included in COMPARE and VERIFY statements. Limits for the value of a labeled variable or a measured signal are expressed in this field. For example:

```
fstatno  COMPARE, 'label', <evaluation characteristic> $

000620   COMPARE, 'RESULT', UL 9V LL 3V $

000690   VERIFY (VOLTAGE), DC SIGNAL, UL 10V LL 8V, CNX HI J1-1 LO J1-2 $
```

The specification of limits in the COMPARE or VERIFY statement allows either single-ended comparisons (greater than, less than, equal, not equal, greater than or equal, less than or equal) or double-ended comparisons that specify upper and lower bounds. The following abbreviations are used for limit specification:

UL - upper limit	GT - greater than
LL - lower limit	LT - less than
LE - less than or equal to	EQ - equal to
GE - greater than or equal to	NE - not equal to

As the words imply, algebraic magnitude determines which limit is identified as UL and which is LL. (Example: UL -4V LL -6V.)

The relationship between the value and the limits is preserved in a set of software flags, which may be examined by subsequent statements that contain a GO TO verb and an IF TEST such as the following:

```
 20 GO TO, STEP 250 IF HI $
 30 GO TO, STEP 260 IF LO $
 40 GO TO, STEP 270 IF GO $
 50 GO TO, STEP 280 IF NO GO $
 60 GO TO, STEP 290 IF LE $
 70 GO TO, STEP 300 IF GE $
 80 GO TO, STEP 310 IF NE $
 90 GO TO, STEP 320 IF EQ $
100 GO TO, STEP 250 IF HI, STEP 210 IF LO $
```

The software flags are HIGH, LOW, EQUAL, GO, and NOGO. GO TO statements may specify several conditions to be evaluated as shown in step 100. In this step, the branch is made to statement 250 if the condition is high, to step 210 if the condition is low, or to the step following 100 if neither of these is true. A single GO TO statement may contain any number of destination steps and conditions. Evaluation proceeds from left to right, and the first condition, which is identified as true, will have its branch selected. Execution will proceed with the next step in sequence if all conditions are evaluated as false.

Condition Branch With GO-TO-STEP or IF-GO Modifier.– The inclusion of a GO-TO-STEP modifier in the statement, as shown in the example below, will cause transfer to the statement number specified if the system NOGO is set upon completion of statement execution. This eliminates the need for a separate COMPARE statement as it allows a measurement to be taken, a comparison to be made, and a NOGO branch to be taken in a single statement, This eliminates the need for most GO TO . . . IF-NOGO statements in the test program.

```
20 VERIFY, (VOLTAGE), DC SIGNAL, GT 19V, GO-TO-STEP 700, CNX HI J1-19 LO J1-20 $
```

The inclusion of an IF-GO modifier in the ATLAS interpreter provides a companion to GO-TO-STEP to allow the branch to be taken on the GO condition rather than the NOGO condition. The GO-TO-STEP modifier must be used with the IF-GO modifier.

For example:

```
20  VERIFY, (VOLTAGE), DC SIGNAL, GT
    19V, GO-TO-STEP 700, IF-GO, CNX HI
    J1-19 LO J1-20 $
```

LOOPS.– When a program is being written in which one or more portions are performed not just once but a number of times, the portion to be repeated may be written just once through the use of a programming device known as a loop.

The use of loops is illustrated by a comparison of the following programs used to print a table of the first 100 positive integers together with the square root of each. Without a loop, the program would be 101 lines long as follows:

```
000010  PRINT, RESULT, 1, SQRT (1) $
000020  PRINT, RESULT, 2, SQRT (2) $
000030  PRINT, RESULT, 3, SQRT (3) $
                  . . .
000990  PRINT, RESULT, 99, SQRT (99) $
001000  PRINT, RESULT, 100, SQRT (100) $
001010  FINISH $
```

The following program uses a loop to obtain the same table with six instructions instead of 101.

```
000010  CALCULATE, 'X' = 1 $
000020  PRINT, RESULT, 'X', SQRT ('X') $
000030  CALCULATE, 'X' = 'X'+1 $
000040  COMPARE, 'X', LE 100 $
000050  GO TO, STEP 20 IF GO $
000060  FINISH $
```

Step 10 gives the value of 1 to 'X' and "initializes" the loop. In step 20, both 1 and its square root are printed In step 30, the value of 'X' is increased by 1. Step 40 compares 'X' to 100 and sets the condition flags. Step 50 directs the program back to step 20 if the GO condition flag was set by step 40. This process is repeated until the loop has been traversed 100 times. After 100 and its square root have been printed, the value of 'X' becomes 101. The comparison in step 40 results in setting the NOGO flag, and step 50 will not direct the program back to step 20. This causes the next step, step 60, to be executed. Execution of step 60 terminates the program. All loops contain four characteristics: initialization (step 10), the body (step 20), modification (step 30), and an exit test (steps 40 and 50). Because loops are so important and because hops of the type just illustrated arise so often, ATLAS provides LOOP and REPEAT verbs to specify a loop even more simply. ATLAS also provides loops within loops.

LOOP Verb.– This verb is illustrated in the following example:

```
000010  LOOP, THRU STEP 20 with 'X' = 1
        TO 100 $
000020  PRINT, RESULT, 'X', SQRT ('X') $
000030  FINISH $
```

only three steps are required to perform the entire program. In line 10, 'X' is set equal to 1 and a test is set up. After this, each time step 20 has been completed, 'X' is incremented by 1, and a test is made to determine whether the loop should be repeated or the next statement should be executed. The loop will be repeated as long as the incrementation results in 'X' being less than or equal to 100. Thus, step 10 replaces steps 10, 40, and 50 in the preceding example.

Note that the value of 'X' is increased by 1 each time through the loop. If an increase of 5 were wanted, it could be specified by writing:

```
000010  LOOP, THRU STEP 20 WITH 'X' = 1
        TO 100 BY 5 $
```

The computer would assign 1 to 'X' on the first time through the loop, 6 to 'X' on the second time through, 11 on the third time, and 96 on the last time. Another step of 5 would take 'X' beyond 100, so the program would proceed to step 60 after printing 96 and its square root. The increment value following the BY may be positive or negative. The same table printed in reverse order could be obtained by writing step 10 as follows:

```
000010  LOOP, THRU STEP 20 WITH 'X' =
        100 TO 1 BY -1 $
```

For a positive incrementation value, the loop continues as long as the control variable is algebraically less than or equal to the final value. For a negative incrementation value, the loop continues as long as the control variable is greater than or equal to the final value. In the absence of a BY value, an incrementation of +1 is assumed. More complicated loop statements are allowed. The initial value, the final value, and the incrementation value may all be formulas of any complexity. If the initial value is greater than the final value (less than for a negative incrementation value), then the body of the loop will not be performed at all, and the computer will automatically pass to the step following the end of the loop.

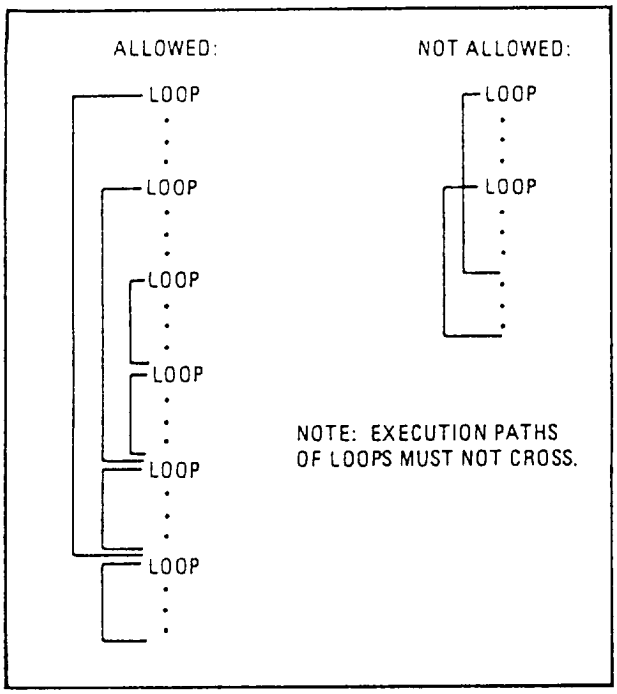


Figure 9-9.-Nested loops.

Nested Loops.– It is often useful to have loops within loops. These are known as nested loops, and can be expressed by loop statements written in any order. However, the loop statements must actually be nested and must not cross. Figure 9-9 illustrates the relationship of nested loops.

Any number of loops may be nested to any level, and transfers out of the bounds of loops are permitted. If such a transfer occurs on nested loops, only the loop whose range was transferred out of will be disabled. Consider the example shown in figure 9-10. Both loops A and B are initialized and execution proceeds until the GO TO C statement is executed. Because C is outside the range of loop B, loop B will be disabled at this time. This means that subsequent transfers back into the range of loop B, which do not pass through the LOOP statement, will not result in reactivation of loop B. The transfer does not, however, affect loop A, and this execution will continue normally. If C had been outside the range of both A and B, both loops would have been deactivated.

REPEAT Verbs.– This verb provides additional loop capability. The REPEAT statement differs from the loop statement in that it appears at the end of the loop and is used to repeat a series of steps directly preceding the REPEAT statement. For example, to execute five cycles of power cycling, the following program could be used:

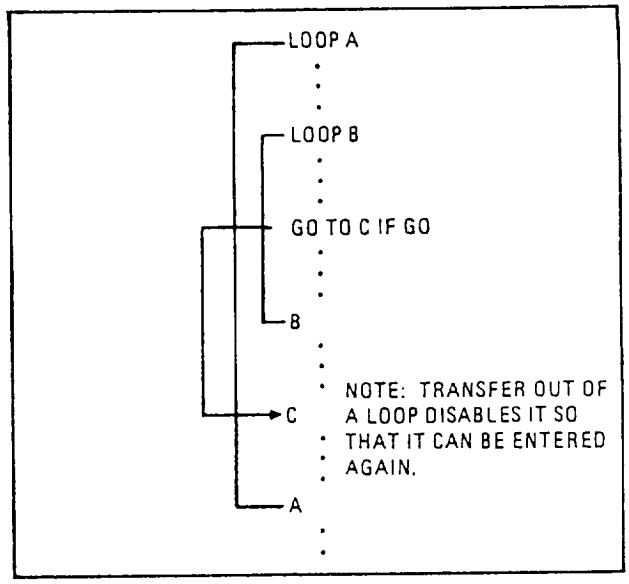


Figure 9-10.-Nested loops transfers.

```
000010 APPLY, DC SIGNAL, VOLTAGE 5V,
       CNX HI J1-1 LO J1-2 $

000020 DELAY, 20 SEC $

000030 REMOVE, DC SIGNAL, CNX HI
       J1-1 LO J1-2 $

000040 DELAY, 20 SEC $

000050 REPEAT, STEP 10 THRU STEP 40, 4
       TIMES $

000060 FINISH $
```

Note that the REPEAT statement in step 50 only specifies 4 repetitions, as the sequence is performed once before the REPEAT statement is reached. REPEAT statements may be nested in any manner, although unusual results may occur if the REPEAT range is overlapped. Transfers out of the range of the REPEAT statement do not disable the repetition cycle, and a subsequent transfer back within the range will continue the repetition where it was left off. The REPEAT count is reinitialized only if the statement is being encountered for the frost time, that it was never initialized previously, or if the previous repetition was successfully completed. Note that the last statement in a REPEAT loop must always be the statement immediately preceding the REPEAT statement.

Procedure

When a particular part of a program is to be performed more than one time or at several different

places in the test program, it may be separated so that it need only be written once as a procedure. A procedure is performed by executing a PERFORM statement specifying the procedure name. The following example illustrates the creation and use of a simple procedure to apply POWER to the UUT.

```
000010  DEFINE, 'POWER UP', PROCE-
        DURE ('VOLTS') $

000020  APPLY, DC SIGNAL, VOLTAGE
        'VOLTS', CNX HI J 1-1 LO J1-2 $

000030  APPLY, DC SIGNAL, VOLTAGE
        'VOLTS', CNX HI J1-3 LO J1-2 $

000040  END, 'POWER UP' $

. . .

001000  PERFORM, 'POWER UP', 5V $

. . .

002000  PERFORM, 'POWER UP', 10V $

. . .

003000  PERFORM, 'POWER UP', -15V $
```

In this example, step 10 defines 'POWER UP' as a procedure having 'VOLTS' as a single variable whose value is to be specified by the PERFORM statement. Steps 20 and 30 are the body of the procedure, and step 40 is the end of the procedure. All procedure definitions must be contained in the preamble of the ATLAS test program. When called upon to RUN, the ATLAS test program scans the preamble, sets up linkages, and begins execution with the first step following the last END statement. END statements are used only in procedures. Execution of the example test program will proceed until step 1000 is encountered. Execution of step 1000 causes the 'POWER UP' procedure to be performed with 'VOLTS' = 5. This has the effect of applying +5 VDC to J1-1 and ground to J1-3 of the UUT. When the END statement of the procedure is encountered, execution of the test program proceeds with the step following 1000. When step 2000 is executed, the 'POWER UP' procedure is performed again with the value of 'VOLTS' equal to 10 V. This results in applying +10 VDC to J1-1 and ground to J1-3. The test program then resumes execution following step 2000. Step 3000 executes the 'POWER UP' procedure again, this time with the value of 'VOLTS' equal to -15. This has the effect of applying -15VDC to J1-1 and ground to J1-3.

When a procedure is performed, execution commences with the statement immediately following the procedure definition statement and continues until the first END statement is encountered. A procedure may, of course, have more than one END statement if branching is involved. There is no limit to the length of a procedure, and any statement except DEFINE, COMMON, and DECLARE may be used. A procedure may perform other procedures or may even perform itself if such usage makes sense. The level of procedure nesting is limited only by the amount of temporary core storage available during program execution.

Values transmitted to the procedure by the PERFORM statement are known as actual parameters and may be numbers, constants, variables, or formulas of any complexity. The corresponding values in the procedure definition ('VOLTS' in the example above) are known as formal parameters. They can only be variables. Any formal parameters specified by the DEFINE PROCEDURE statement may have a label identical to a variable used in the copy of the test program. However, this can cause confusing results and is, therefore, not recommended. The correspondence between the values of actual PARAMETERS and formal parameters is determined by position in the PERFORMS statements and the DEFINE PROCEDURE statements, not by similarity in the label. No matter what value a formal parameter variable might achieve during execution of a procedure, its value will not be transferred through the parameters to the variable with the same name in the test program unless they occupy the same position in the specification of actual parameters and formal parameters.

Results or values maybe passed from the procedure to the test program in two ways.

1. Set a formal parameter equal to the value to be passed. The value will be passed to the actual parameter in the corresponding location so that a value assigned to the formal parameter 'VOLTS' in the test procedure may appear as the value of the actual parameter 'CURRENT' in the body of the test program if the variables are labeled that way. If the corresponding actual value is a constant, an attempt to pass a new value to it will cause a terminating error to be generated.

2. Within the test procedure, place the value to be transferred in any variable used in the main body of the test program. All variables within an ATLAS test program are treated as global variables (similar to Fortran Common) except those that are defined as formal parameter procedure. This means the procedure

can use and operate upon all the variables available to the entire ATLAS test program.

External Statement

A facility is provided to allow an ATLAS program to PERFORM a procedure that is not defined in its preamble but has been saved on the system disk. For such operation, the system disk must operate in the non-file-protected mode, as it is used for temporary storage during program execution. Procedures saved on a disk may be nested to any level permitted by the availability of disk storage. Arguments may be transferred to and from disk procedures only through the actual parameter list of the PERFORM statement or through variables declared as COMMON. Non-COMMON arrays will not transfer as actual parameters. The usual global variable characteristic is not available with disk procedures. Disk procedures need not use COMMON at all, but if they do, the COMMON declaration should be identical in both the PERFORMING program and the disk procedure.

The EXTERNAL statement is used by the test programmer to define all disk resident ATLAS programs and ATLAS procedures that are referenced by an ATLAS in EXECUTE or PERFORM statements. Any number of ATLAS programs or procedures may be declared as external in a single EXTERNAL statement. A typical EXTERNAL statement might be as follows:

```
000010 EXTERNAL, 'SUB1', 'SUB2', 'PROG4' $
```

All EXTERNAL statements should appear in the preamble section of the ATLAS program. A good programming practice is to use only a single EXTERNAL statement in a given program. This provides a quick cross-reference to all programs and procedures referred to by that particular program.

ATLAS Output

ATLAS provides a facility for presenting messages and variables to the test operator with the DISPLAY, PRINT, and INDICATE verbs. These three statements have identical structure and differ only in the equipment cm which the message appears. The DISPLAY statement directs messages to the CRT, the PRINT statement directs messages to the printer, and the INDICATE statement directs messages to the system indicator. In the discussion that follows, only the PRINT statement is considered, but all comments apply with equal validity to DISPLAY and INDICATE. The most common uses of the PRINT statement are as follows: to print the results of a test, to print a message for the test operator, to print a combination of the two, and to skip a line. There are two basic forms of the PRINT statement. These types are identified by the first two words of the statement, which may be either PRINT, RESULT or PRINT, MESSAGE.

The PRINT statement using MESSAGE is useful only for printing verbatim messages to the test station operator. It will print all characters between the comma and the dollar sign, and it will return the carriage at the end of the line. It may not be used to print the value of variables.

The format using RESULT is considerably more flexible. It maybe used to type out verbatim messages to the operator, the values of internal program variables, constants, formulas, or any combination of these. A single print statement may intermix character strings and variables. This intermixing may be done in any order, and the variables may be replaced by formulas. Character strings must be enclosed in double quotes, and the character strings and variables must be separated from each other by commas or semicolons.

Character strings are printed as they are entered, including all embedded spaces and adding none. All numeric values are printed left, justified in their fields, with leading zeros suppressed. No trailing spaces are appended to numeric values. For positive real numbers, the first character will be printed as a space. For negative real numbers, the first character will be printed as a minus sign.

Output Line Formats

The output line produced by DISPLAY, PRINT, and INDICATE statements is divided into five printing zones, starting at positions 0, 15, 30, 45, and 60. A terminator (comma or semicolon) controls the use of these zones. A comma (,) moves printing to the next zone, or if the fifth printing zone has been tilled, to the first printing zone on the next line. A semicolon (;) produces more compact output by inhibiting spacing between printing zones, acting only to separate quantities to be printed or to suppress the usual carriage return at the end of a DISPLAY, PRINT, or INDICATE statement. Output format can be further controlled by use of the TAB function. Insertion of TAB (17) causes the printer to move to print column 17. For this purpose, output columns are numbered 0 through 79. TAB can contain any expression as its argument. The value of the expression is computed, truncated, and its integer par taken. The output device then moves forward to this position. If the position has already been passed, the

TAB is ignored. If the result is greater than 71, the output device moves to position zero of the next line. The statement PRINT $ will cause a line to be skipped. Printing the character string "%%" will cause the output device to start a new page. Some sample PRINT statements and resultant printouts are shown below.

Sample PRINT statements:

```
000010 PRINT, MESSAGE, THIS UUT IS A
       BASKET CASE $

    20 CALCULATE, 'A' = 3, 'B' = 4 $

    30 PRINT, RESULT, ''A ='', 'A', ''B ='', 'B',
       ''SQRT = ''; SQRT ('B'**2+'A'**2) $

    40 PRINT RESULT, ''A = ''; 'A', ''B =;'' 'B',
       ''SQRT ='';SQRT ('B' **2+'A'**2) $

    50 PRINT, RESULT, ''A ='''; 'A', ''B =; 'B';
       SQRT ='';SQRT ('B'**2+'A'**2) $
```

The resultant printouts are as follows:

```
THIS UUT IS A BASKET CASE

A =        3          B =  4      SQRT = 5

A = 3      B=4        SQRT = 5

A = 3B = 4SQRT = 5
```

Number Formats

For printing all numbers, ATLAS provides a default format that automatically allocates a field large enough to contain the number and its sign. No leading or trailing spaces are included in the default format. A floating decimal point is used when required. The decimal point is omitted for integers. Exponential notation is used for real numbers outside the absolute range .1 to 999999. The form in which a number is printed is the same as the form in which it was entered or created. An octal number will be printed in octal format. A real number will be printed in the form of a real number.

FMT Function for Number Control

Additional format control is provided for printed numbers by the FMT function. The argument for the FMT function has two meanings, depending on the number to be printed. At run time, the interpreter determines if the number to be printed is a real number or a bit string number. The meaning attached to the argument is selected according to this determination. In either case, the argument is a five-digit integer. For example:

```
100 PRINT, RESULT, FMT (20805), 'B' $
```

ATLAS Input

There are times when it is desirable to have data entered by the test station operator during the running of a test program. Examples of this are entries of meter readings, the frequency at which a receiver is to be tested, or an indication of the state of illumination of an indicator on the UUT. With the WAIT FOR statement, the ATLAS interpreter provides three separate facilities for operator input. Examples of the use of this statement are as follows:

```
000100 WAIT FOR, MANUAL DATA, QUERY $

    10 WAIT FOR, MANUAL DATA, GO-NO
       GO $

    20 WAIT FOR, MANUAL DATA, 'A', 'B' $
```

Statement 100 is an example of the flying look option. Upon executing this statement, the test program interrogates the console to see whether the GO key has been depressed. If it has, the GO software flag is set, and if not, the NOGO software flag is set. Execution then proceeds with the next statement.

Statement 110 causes the computer to wait until the operator depresses either the GO or NOGO keys on the console keyboard. Execution of the program will not proceed until one or the other of these keys is depressed. When this has occurred, the appropriate software flag is set and execution of the program continues.

Statement 120 is an example of how numeric data is obtained from the test station operator. Execution of this statement causes the program to type a question mark on the system control device and to await the operator's entry oft two real numbers. These are entered as two real numbers separated by a comma, and the line is terminated by a $. The values of these real numbers are then entered into variables 'A' and 'B' and execution continues with the next program step. Any number of variables may be requested. The program will not continue until the operator has entered the required number of variables.

Frequently a WAIT FOR statement is combined with the DISPLAY statement to make sure that the

operator knows what the question mark is asking for. An example might be the following:

```
000200 DISPLAY, RESULT, ''TEST FREQUENCY'': $

10 WAIT FOR, MANUAL DATA, 'FREQ' $
```

On the CRT display, the operator would see the following:

TEST FREQUENCY

Data entered with a WAIT FOR statement is not saved with the program. Furthermore, it may take a long time to enter a large amount of data using WAIT FOR. Therefore, WAIT FOR should be used only when the need for entering data during the running of the program is unavoidable.

Program Linking

A program linking feature has been built into the ATLAS interpreter that permits test programs to be written that are larger than the available computer core memory. To use this feature, the user stores the program on the disk file in segments as individual programs. Segment names are usually assigned in an orderly fashion, often by appending a letter suffice to the program name. Any ATLAS program may cause the loading and execution of any other ATLAS program that is resident on the disk through the use of the EXECUTE VERB. When under ATLAS control, program segmentation is not required, since the program maybe as large as the available space on the applications disk. But sometimes it can be useful to overcome limitations existing on the number of variables allowed under ATLAS control. It is possible to pass variables between programs through the use of the COMMON statement.

EXECUTE VERB.– This verb is used in a statement that specifies the name of the program to be executed and, if desired, the entry point to be used. It will cause the specified program to be loaded on top of the program in memory and the execution of the program to begin. Note that the original program will be lost from computer memory. If no entry point is specified, the program is executed from its beginning. Two examples follow:

```
001000 EXECUTE, 'ALIGN TEST' $

001000 EXECUTE, 'ALIGN TEST', 3020 $
```

An EXECUTE verb can be used with a statement number in a test program to link the program to another part of the test. When the EXECUTE verb is used in the Immediate Mode; that is, with no statement number, the program referenced is loaded from disk memory and executed.

DELAY STATEMENT.– This statement is used to postpone execution of the following statement until a specified interval has passed. If a time dimension is not specified, seconds is assumed. Two examples follow:

```
0630 DELAY, 5 SEC $

0690 DELAY, 'X' –3 MSEC $
```

Delay times may range from zero to 3.2 x 10^7 milliseconds. Resolution is 1 millisecond or 0.001% of the delay time, whichever is greater.

FINISH STATEMENT.– This statement is required at the end of every ATLAS program to terminate testing, to reset test equipment to the quiescent state, and to return control to the operator. One example follows:

```
009990 FINISH $
```

TERMINATE STATEMENT.– This statement designates the last statement of a complete ATLAS program, and like the BEGIN statement, it has no function other than to improve readability of the program. If used, it is always the last statement in a program. One example follows:

```
009999 TERMINATE, ATLAS PROGRAM $
```

CHAPTER 10

WAVEFORM INTERPRETATION

Chapter Objective: *The primary objective of this chapter is to interpret common voltage and current waveforms as they are observed on an oscilloscope and a spectrum analyze.*

A waveform may be considered a pictorial representation of a varying signal as it is related to time. An unknown waveform can be graphically plotted by using a system of coordinates where the amplitude of the unknown signal is plotted linearly against time. An analysis of the resultant waveform provides valuable information in determining the characteristics of many electronic devices. The waveform of a signal may indicate the presence of harmonics or parasitic oscillations, or it may indicate how closely a device is following a desired cycle of operation. Distortion of a waveform is the undesired change or deviation in the shape of the observed signal with respect to some reference waveform.

One of the most important steps in waveform analysis-the one which usually proves the most difficult for the maintenance man-is the interpretation of patterns as viewed on the oscilloscope. For this reason, the following section of this chapter is designed to help you accomplish this task.

WAVEFORM AND PHASE DEVELOPMENT

Learning Objective: *Identify and interpret the information regarding complex waveform and place development to illustrate and explain normal waveforms, abnormal waveforms, and their causes.*

A complete understanding of the appearance and reason for a normal waveform will help you to recognize an abnormal waveform, and may help you to understand the reason for the abnormality. Terms such as *loss of high frequencies* and *loss of response* will have special meanings when referring to a particular waveform. The causes of deteriorated voltage and current waveforms will be discussed. The information regarding complex waveforms and phase development given in this section will illustrate and explain normal waveforms, abnormal waveforms, and their causes. This material does not include a mathematical analysis of waveforms; it is presented only as an aid to those individuals having a vital interest in waveforms and who cannot, or do not, desire to interpret mathematical explanations.

A voltage or current waveform, as encountered in the electronics field, may be graphically represented in both height and width. The height of a graphically displayed waveform represents the quantity or amplitude of voltage or current. The width of the displayed waveform represents the elapsed time or waveform duration. Also, a voltage or current waveform is normally represented in a two-dimensional (horizontal and vertical) plane without depth. The horizontal (X) axis on a graph will represent time measured in either whole or parts of a second; the vertical (Y) axis will represent amplitude, quantity, or intensity of the subject waveform measured, either in whole or in parts of volts or amperes. Any portion of the waveform extending above the horizontal (zero amplitude) reference line is considered positive, and any portion of the waveform extending below the horizontal reference line is considered negative.

SINUSOIDAL WAVEFORMS

The sine wave is the basis of all other waveforms. It represents the simple action of a swinging pendulum, a bouncing or vibrating spring, a free-running self-excited oscillator output, etc. When any outside source changes the shape of the sine wave, the wave is said to be distorted. However, you will find that the original sine wave is still present in combination with other sine waves introduced by the distorting agency to produce a single resultant waveform. Therefore, any waveform, no matter how complex, may be reduced to its individual sine wave components. The original sine wave components cannot be reduced further because they are the final remaining single-frequency basic components. No waveform that is composed of more than one frequency is a true sine wave.

Since the primary purpose of this portion of the section is to familiarize you with the basic structure of a sine wave, the sine wave is graphically presented in figure 10-1. A 60-hertz wave is represented by one complete cycle; the total time duration for this cycle is 1/60 of 1 second. Half of this cycle time is above the horizontal reference (zero amplitude) line and is considered positive, while the other half is below the horizontal reference line and is considered negative. The two halves of the sine wave do not cancel out or nullify each other because each half-cycle occurs during a different time. During the positive half-cycle, the negative half-cycle has not occurred; therefore, it does not exist. During the time that the negative half-cycle is present, the positive half-cycle does not exist. The single cycle illustrated in figure 10-1 must not have any bumps or kinks on either its increasing or decreasing side. The top (positive peak) and the bottom (negative peak) must be smoothly curved with no appearance of either a point or a flat spot in this region. The positive and negative half-cycles of the sine wave must be exactly equal in both amplitude and time duration. In figure 10-1, the maximum positive peak amplitude is represented by positive 1 volt, while the maximum negative peak amplitude is represented by negative 1 volt. The time duration illustrated is exactly 1/120 of a second for each half wave.

Considering that a sine wave represents a complete mechanical revolution or circle, it requires 90 degrees to traverse one-fourth of the circle; 180 degrees to traverse one-half of the circle, 270 degrees to traverse three-fourths of the circle; and 360 degrees (or back to 0 degrees) to complete the entire circle. There are an infinite number of points represented on a sine wave. For example, there are 360 different points, each representing an advance of 1 degree, on each sine wave. However, only four points are shown to illustrate the shape of a basic sine wave. This was done because it is only necessary to become familiar with the general features of the sine wave curve rather than to provide a point-by-point analysis.

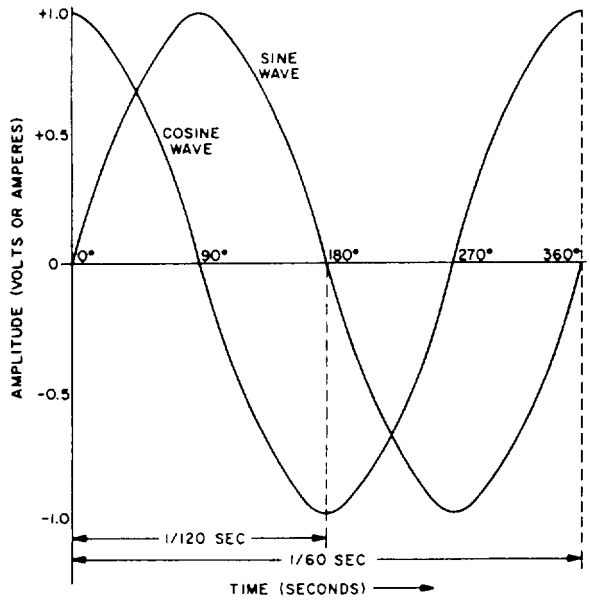


Figure 10-1.-Sine and cosine waveforms.

Amplitude will not affect the general outline of a sine wave provided that the positive and negative portions of the waveform contain equal amplitudes. If the sine wave is viewed on some form of oscilloscope, the instrument controls may be incorrectly positioned, thus presenting the wave as either too narrow or too wide for its height. This will not affect the actual waveform, but may affect the viewers perspective; the peaks of the normal waveform may appear sharp or peaked for the narrow vertical horizontal version, and flat or broad on the wide or stretched horizontal version. If you disregard its disproportional appearance, a point-by-point examination of the wave will prove that it is a true sine wave. A cosine waveform is also shown in figure 10-1.

The cosine wave is the same in all respects as the sine wave except for one difference; it leads, or begins 90 degrees (1/240 of a second in this case) before the sine wave time begins. The cosine wave is superimposed on the same graph as the sine wave to illustrate the cosine lead. These two waveforms arc not used to provide a resultant waveform.

The half-sine waveform consists of a series of unidirectional pulses, each resembling a half-cycle of a sine waveform. The half-sine wave may exist either above (positive) or below (negative) the horizontal reference line. The half-sine waveform is produced by removing any amplitude variations from the complete sine wave in one direction for a period of one-half the time duration of the complete cycle. As shown in figure 10-2, there are two types of half-sine waves. In view A of the figure, the negative portion of the sine waveform has been removed and its one-half cycle time interval remains as a zero dc reference level. In this type of half-waveform, the frequency remains the same as the original full sine wave frequency. In view B of the figure, the negative half of the original full sine wave has been inverted over the horizontal reference line; consequently, the average dc voltage level is increased. Since each alternation occurs in one-half the time

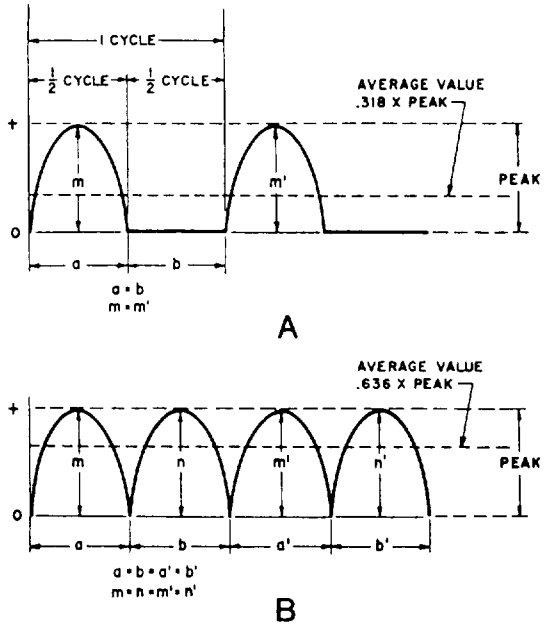


Figure 10-2.-Half-sine waveforms.

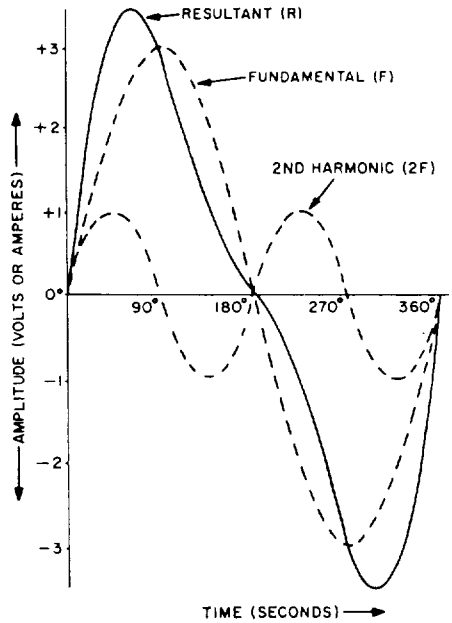


Figure 10-3.-Waveform resulting from algebraic addition of a fundamental sine wave with its second harmonic in phase.

interval of the original full sine waveform, inverting the negative alternations to occupy the empty spaces between the positive alternations causes the frequency to double. Half-sine waveforms are composed of the original fundamental frequency in conjunction with a dc component and an infinite series of even numbered harmonics of progressively decreasing amplitude.

NONSINUSOIDAL WAVEFORMS

The sine wave is the basic or standard alternating current (or voltage) waveform used in combinations with phase or time differences and amplitudes to algebraically form all other waveforms. The sine wave is the wave most commonly used as an input to circuits under test because it does not introduce distortions commonly associated with nonsinusoidal waveforms.

All nonsinusoidal waveforms can be reduced to their individual component sine waves. A nonsinusoidal waveform is composed of more than one sine wave; other frequencies, usually harmonically related, are algebraically added to the fundamental frequencies to produce the resultant nonsinusoidal waveform. In this case, the sine wave of lowest frequency is normally considered to be the fundamental frequency, and higher frequencies that are exact multiples of the fundamental frequency are considered as harmonics of the fundamental. However, in some cases, the nonsinusoidal waveform being considered may be composed of only harmonic frequencies because the fundamental sine wave may have been intentionally removed. The algebraic addition of the fundamental sine wave (f), and the second harmonic of the fundamental (h = 2f) will provide a resultant nonsinusoidal waveform (R), as shown in figures 10-3, 10-4, and

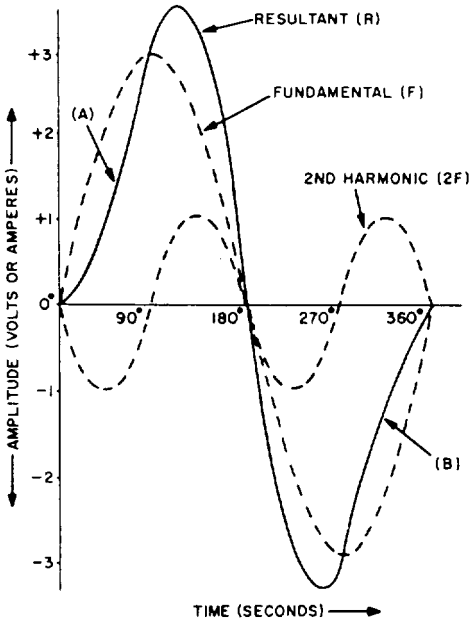


Figure 10-4.-Waveform resulting from algebraic addition of a fundamental sine wave with its second harmonic delayed 180 degrees.

10-5. However, only the resultant would be shown on an oscilloscope or equivalent test instrument. The second harmonic of figure 10-3, is shown in phase with the fundamental because its amplitude increases in the same direction as the fundamental from the horizontal and vertical zero reference level. The second harmonic of figure 10-4 is shown 180 degrees out of phase with the fundamental because it proceeds from the horizontal and vertical zero reference level in a direction exactly opposite (negative direction) from the fundamental. The second harmonic shown in figure 10-5 is shifted 90 degrees behind (lagging) the fundamental. The resultant waveforms contained in figure 10-3, 10-4, and 10-5 are the only waves you will see on an oscilloscope or other equivalent test instrument.

The amplitude of the second harmonic, relative to the fundamental sine wave, will either increase or decrease the amount of dip at points (A) and (B) of figure 10-4, represented by two heavy arrows. However, if the phase of the second harmonic is changed with respect to the fundamental sine wave, the appearance of the resultant waveform will change completely. As shown in figure 10-5, the positive half of the resultant waveform looks like part of a sine wave, but the negative half does not. Therefore, this resultant is definitely not a true sine wave. At any point along the horizontal zero reference line, the vertical amplitude of the fundamental can be added directly to the vertical amplitude of the harmonic to obtain the final amplitude of the resultant waveform at that particular point (that is, if f = 3 units and h = –1 unit, then R = 2 units; and, if f = –6 units and h = –3 units, then R = –9 units). The resultant

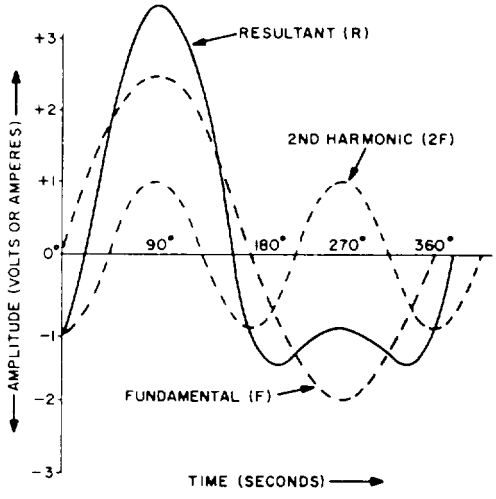


Figure 10-5.-Waveform resulting from algebraic addition of a fundamental sine wave with its second harmonic delayed 90 degrees.

waveforms (figs. 10-3, 10-4, and 10-5) show that the algebraic addition of a harmonic waveform to the original fundamental sine wave produces a new waveform that is no longer a sine wave. In many cases these new waveforms are created deliberately to perform functions beyond the capabilities of the original signal. For example, the new waveforms resulting from the addition of a fundamental and its harmonics maybe used as timing pulses.

Phase Distortion

Figure 10-5 represents the results of feeding the fundamental and its second harmonic through a network that delays the second harmonic by 90 degrees. This normal resultant waveform obtained by the algebraic addition of the second harmonic (without phase shift) to the fundamental sine wave is known as "phase distortion." This type of time delay can be recognized only by becoming familiar with the proper waveforms. If the phase of a fundamental sine wave is shifted, its shape will not change. Therefore, special methods must be employed to recognize any change in phase. These methods are described in the following text.

Harmonic Distortion

The addition of harmonics to the fundamental wave shape creates a new resultant waveform. The resultant is a distortion of the original waveform and, if undesirable, is termed *harmonic distortion*. The resultant waveform created by the addition of only one harmonic to the original waveform can probably be recognized. However, the addition of several harmonics to the fundamental sine wave, in and out of phase, will create a resultant waveform of pure confusion. As was mentioned previously, any waveform can be separated or removed from its resultant with the aid of suitable filters. Therefore, by removing all except one frequency component, you can extract a pure sine wave of some specific frequency that was not evident in the original wave or its harmonics. This newly extracted sine wave can then become the fundamental sine wave input to a circuit under test.

Complex Waveforms

The resultant waveforms discussed in this section are created either by adding harmonics to the fundamental waveform, by changing the phase of the harmonic with respect to the fundamental, or by a combination of harmonic addition and phase change. Therefore, all resultants are termed *complex waveforms*

no matter how simply or easily they are recognized. Actually, other effects are realized in the resultant waveform, depending on the addition of even (2d, 4th, 6th, 8th, etc.) harmonics or odd (3d, 5th, 7th, 9th, etc.) harmonics; on the percentage of harmonic waveform amplitude injected; and on the phase of the introduced harmonic with respect to the fundamental sine wave. All so-called "distorted waveforms" are classified as complex waveforms. These complex waveforms are grouped into types of complex waveforms. Figures 10-6 through 10-8 will help to familiarize you with some of the primary features of nonsinusoidal waveforms that consist of a single harmonic addition to the fundamental

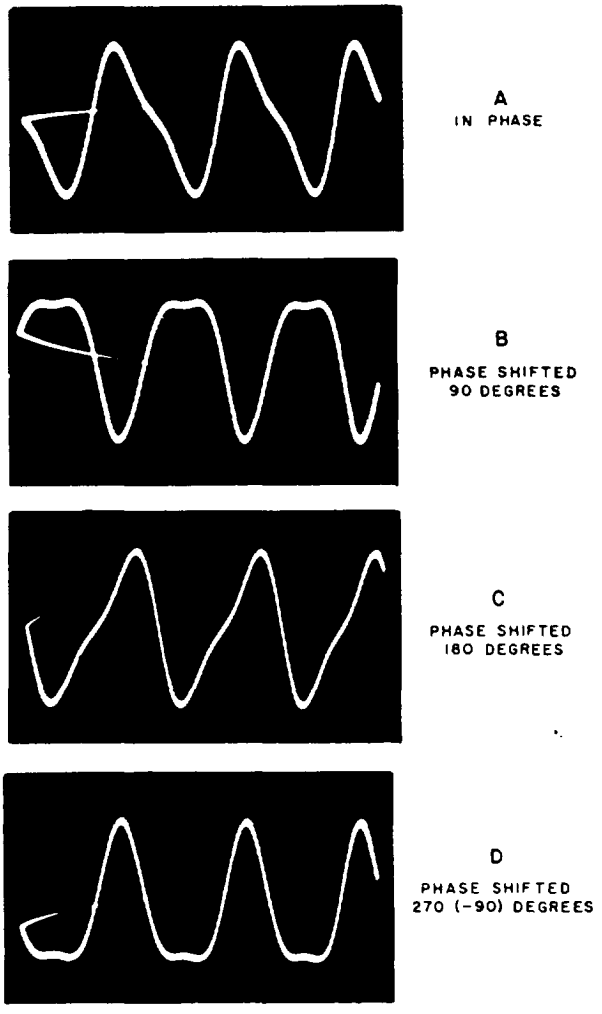


Figure 10-6.-Resultant waveforms created by algebraic addition of second harmonic to fundamental sine wave when second harmonic amplitude is 30 percent of fundamental.

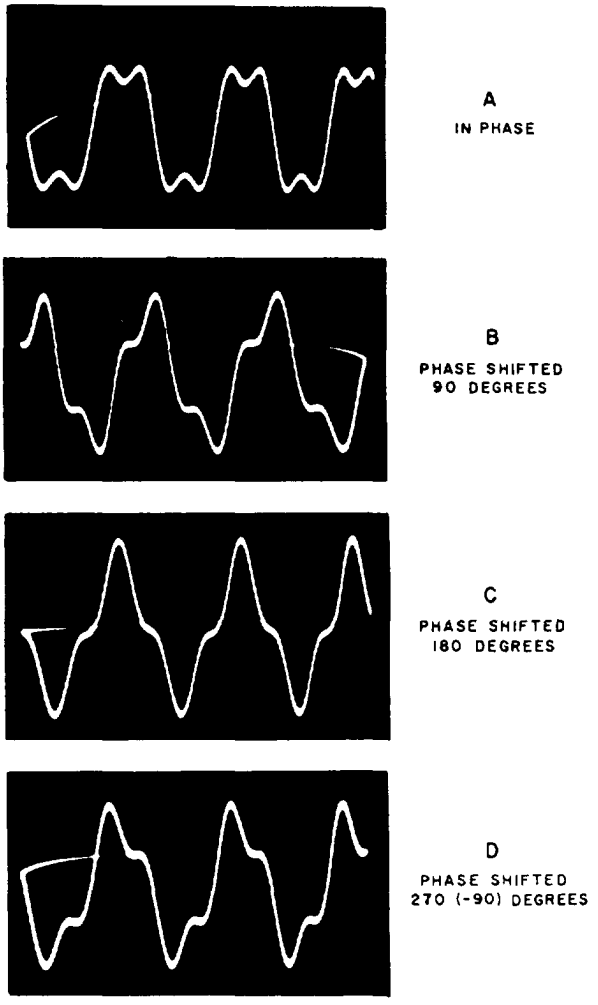


Figure 10-7.-Reaultant waveforms created by algebraic addition of third harmonic to fundamental sine wave when third harmonic amplitude is 30 percent of fundamental.

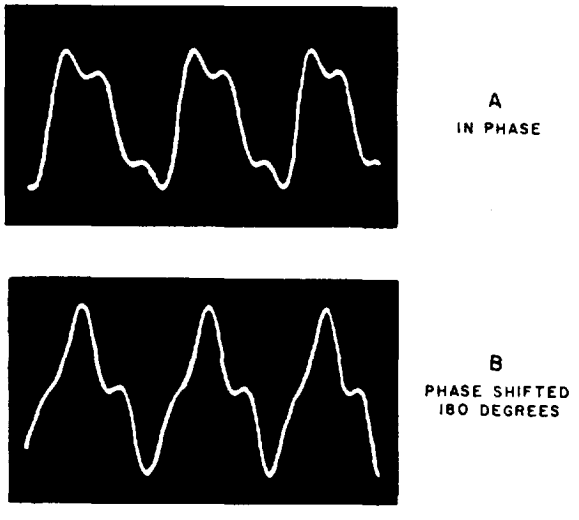


Figure 10-8.-Resultant waveforms created by algebraic addition of third harmonic to fundamental sine wave when third harmonic amplitude is 15 percent of fundamental.

Mirror Symmetry

The term *mirror symmetry* refers to the fact that if the positive part of the resultant wave is inverted over the horizontal reference line, it will exactly match the negative part of the resultant waveform; or conversely, if the negative part of the resultant wave is inverted over the horizontal reference line, it will have the same shape, outline, and appearance of the positive part of the resultant waveform. By viewing the resultant waveform, you can definitely determine whether even harmonics (2d, 4th, 6th, etc.) were added to the fundamental sine wave to create the resultant, or whether odd harmonics were used for this purpose. If even harmonics were algebraically combined with the fundamental sine wave, there will be a lack of mirror symmetry, as shown in view A of figure 10-9. If odd harmonics (3d, 5th, 7th, 9th, etc.) were algebraically combined with the fundamental sine wave, there may be mirror symmetry, as shown in view B of figure 10-9, but only when there is an adequate percentage of the odd order harmonics amplitude.

SQUARE WAVEFORMS

The square waveform is a resultant waveform type composed of a sine waveform in conjunction with odd harmonics. Unlike the original sine waveform, the application of a square waveform to either a capacitive or inductive circuit will result in an output of a completely different waveform shape. As shown in figure 10-10, the leading edge of a square wave rises from zero reference value to its maximum value, where it remains as a constant-amplitude wave over a set period of time. It then drops back toward its original zero reference level, or beyond, until it reaches a minimum value, where it remains as a constant-amplitude wave over an exact period of time that matches its positive excursion. The rise and fall times are negligible in an ideal square wave. View A of figure 10-10 shows the ac square waveform; it is so called because the waveform extends in a negative direction below the horizontal reference line as well as above. Views B and C of figure 10-10 show the pulsating dc square waveforms; they are so called because they contain a dc component that prevents the waveforms from crossing the horizontal zero dc reference level. However, all three forms are identical except for amplitude. All corners must be square, the sides perpendicular, and the extremities flat.

Unfortunately, this idealized square waveform cannot be attained because the waveforming equipment is not perfect. The square wave is formed by the algebraic addition of the fundamental sine wave and an infinite number of odd harmonics of the fundamental sine wave. However, as shown in figure 10-11, as few as three added odd harmonics will produce a reasonable facsimile of a square wave even though a minimum of 10 added harmonics are required to produce a usable square wave. The fundamental sine wave may start at any phase. For illustrative purposes, figure 10-11 shows the sine wave as beginning at the horizontal and vertical

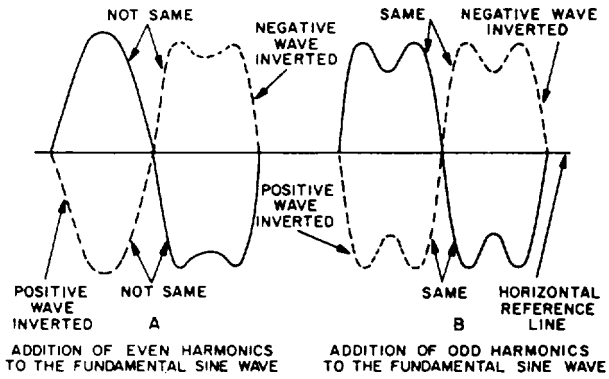


Figure 10-9.-Presence or absence of mirror symmetry due to harmonic addition to fundamental sine wave.

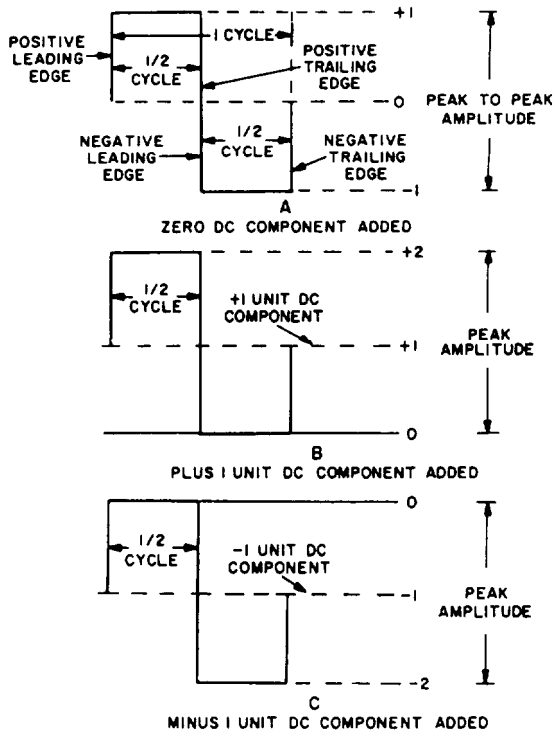


Figure 10-10.-Square waveforms.

zero reference level. Figure 10-11 shows the resultant waveforms as additional harmonics are progressively added to the fundamental sine wave. Although this illustration shows the algebraic addition of only three odd harmonics, it conveys the true impression that as each additional harmonic is added, the leading and trailing edges of the resultant waveform become flatter. The frequency (f) of any order odd harmonic can be determined by calculating the value with the aid of the formula:

$$f_N = (2N + 1)f_1$$

where f is the fundamental frequency and N is the order of the harmonic. For example, if the fundamental frequency is 100 Hz, the frequency of the 6th odd harmonic can be determined as follows:

$$f_6 = [(2 \times 6) + 1] \times 100$$

$$f_6 = [(12) + 1] \times 100$$

$$f_6 = (13) \times 100$$

$$f_6 = 1300Hz$$

It also follows that each odd harmonic shown in figure 10-11 has been added in phase (zero phase difference) with the original fundamental sine wave. In addition, another factor of figure 10-11 should be inspected. The amplitude of each harmonic is in direct proportion to the harmonic order; that is, the third

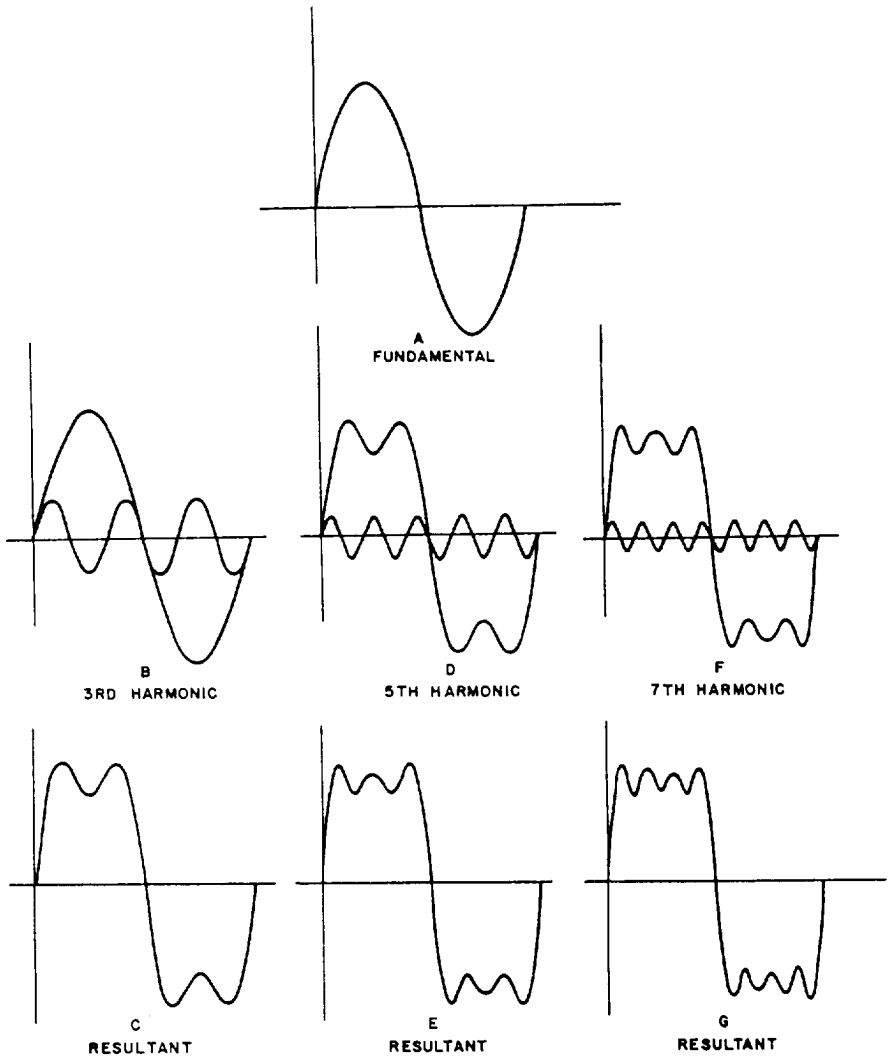


Figure 10-11.-Formation of a square wave.

harmonic contains one-third the amplitude of the fundamental, the fifth harmonic contains one-fifth the amplitude of the fundamental, etc. If the harmonic is not in phase, or has incorrect amplitude, etc., the resultant square wave is said to be distorted. However, the type of distortion observed may indicate the kind of trouble, and even the source of trouble, within a circuit.

RECTANGULAR WAVEFORMS

The rectangular waveform contains all but one feature of the square waveform discussed previously, and it is shown in figures 10-10 and 10-11. This one different feature is that the former wave has identical periods of positive and negative pulsations, whereas the latter wave has unlike periods (time duration) of the positive pulse with respect to the negative pulse. The rectangular pulse can be either bidirectional or unidirectional in that the waveform may be entirely above or entirely below the horizontal zero reference level, as shown in figure 10-12. The period of the rectangular waveform, like that of the square waveform, is the total time required to complete both half-cycles together as one unit. For example, in view A of figure 10-12, if the positive half-cycle duration is 50 microseconds and the negative half-cycle duration is 150 microseconds, the total period for one cycle of this frequency is 200 microseconds. Considering that the frequency of a cycle is the reciprocal of the time required for that cycle, 1/200 μs, then the frequency of this example waveform is 5000 Hz. This means that this particular cycle will repeat itself 5000 times every second; it is said to have a frequency repetition of 5000 Hz. The shorter pulse durations require the presence of higher frequency components, whereas longer pulse durations require the presence of lower frequency components. The rectangular waveform is rarely used as a test voltage. However, it may be used in many special applications to perform a specific function. Figure 10-13 illustrates such a function.

The rectangular wave in this case has a rectangular wave riding atop the first wave. This is a practical situation in the transmission and reception of a television signal cycle. The video information is shown riding on the minimum amplitude portion of the first rectangular wave between the blanking pulses that represents the positive excursions of this rectangular pulse. However, rectangular synchronization pulses are shown riding atop the blanking pulse. This rectangular pulse represents the so-called “front porch,” the synchronizing pulse proper, and the so-called “back porch.”

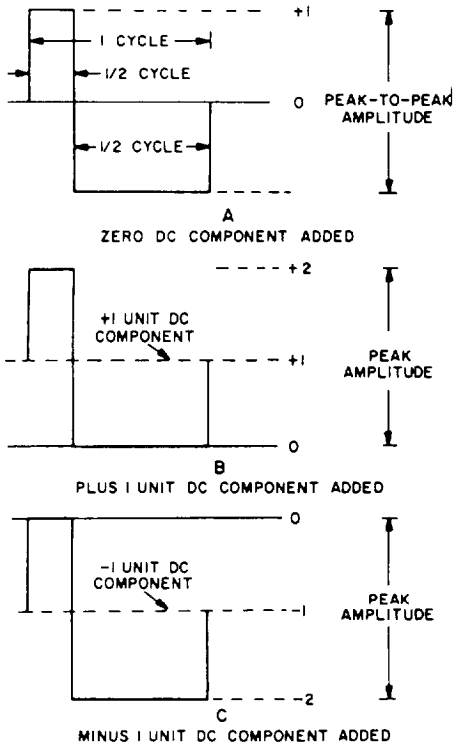


Figure 10-12.-Square waveform.

SAWTOOTH WAVEFORMS

The sawtooth waveform, like all other waveforms except the fundamental sine wave, is composed of sine-wave components. As shown in figure 10-14, the wave consists of a gradual linear change from a maximum negative-going peak to its maximum

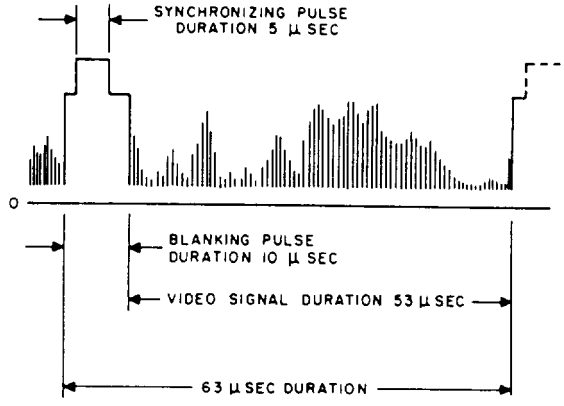


Figure 10-13.-Rectangutar waves used in television.

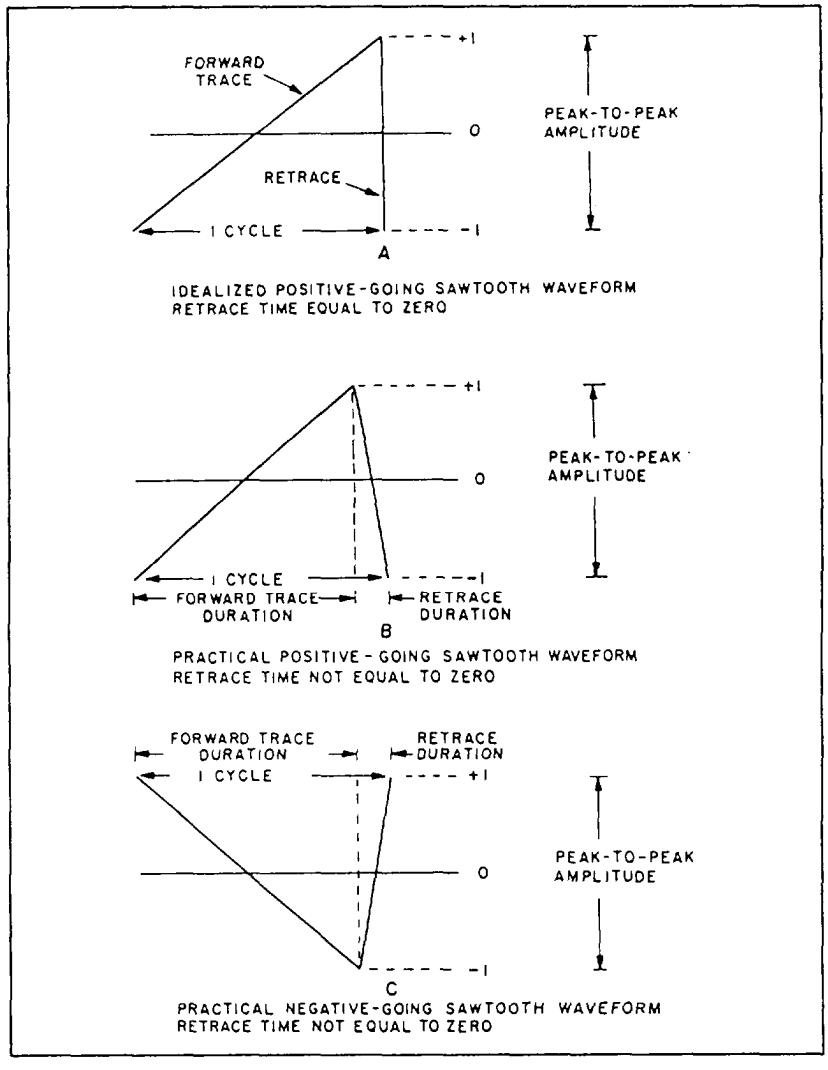


Figure 10-14.-Sawtooth waveforms.

positive-going peak. It then follows a rapid drop to its original amplitude. Considering that this waveform is composed of many sine-wave components that may differ in both frequency and phase, you cannot apply the sawtooth waveform to any inductive or capacitive device to cause different lead or lag times between the sine-wave components that compose the sawtooth waveform. Therefore, the output from any component device, other than a pure resistance, will not be the same as the original sawtooth input. In the ideal waveform, shown in view A of figure 10-14, the retrace time is shown as zero seconds. This is not the true practical case because any action or reaction requires a definite time for accomplishment. Views B and C of figure 10-14 reflect a practical case where the retrace time is some finite time rather than zero. However, the retrace time is normally assigned the smallest practical duration consistent with the design of the equipment with which it is to be used. If the voltage amplitude increases at a constant rate during the forward trace, the waveform is called a "linear sawtooth." The fact that half of the waveform shown in figure 10-14 is above the horizontal zero reference level and the other half is below this reference level will normally not be seen on an oscilloscope because the reference level line (time base line) is absent from the display. Unlike the square waveforms that were produced by the algebraic addition of odd frequency, or in-phase harmonic waves with the fundamental sine wave, the sawtooth waveform is the resultant wave produced by the algebraic addition of both even and odd frequency harmonics to the fundamental sine waveform.

A positive-going sawtooth waveform is produced by the algebraic addition of all harmonics to the fundamental sine wave, but the fundamental harmonic components must begin in-phase and start in a negative

direction, as shown in figure 10-15. A negative-going sawtooth is the resultant of the same sine-wave components, but the fundamental and the in-phase harmonics must start in a positive direction. Figure 10-15 shows the method of progressive algebraic addition of each higher frequency harmonic to the fundamental sine wave to gradually obtain the ultimate sawtooth waveform. However, only the first two harmonics (2d and 3d) have been combined with the fundamental sine wave to form the resultant shown in view D of figure 10-15. As more harmonics are progressively added, the resultant wave will approach more and more closely the required sawtooth form.

TRAPEZOIDAL WAVEFORMS

The trapezoidal waveform is the resultant of the algebraic addition of sine waves, but it is more easily understood in terms of a sawtooth rectangular waveform since both are composed of basic sine waves. As previously stated, a sawtooth of voltage applied to the input of either an inductive or capacitive device will not appear at the output of the reactive component or device as a sawtooth waveform. Therefore, those applications that require a sawtooth current waveform must obtain the sawtooth current from a device that has a sawtooth current output produced from a trapezoidal voltage input. The trapezoidal waveform has the necessary characteristics to cause a linear change in the amplitude of the current with respect to time as it passes through the resistive and inductive components of a coil.

Figure 10-16 shows the resultant output of a resistor versus a coil with an applied input sawtooth waveform. The sawtooth wave passing through the pure resistive element produces no change in the output waveform. However, the output from the coil with an applied

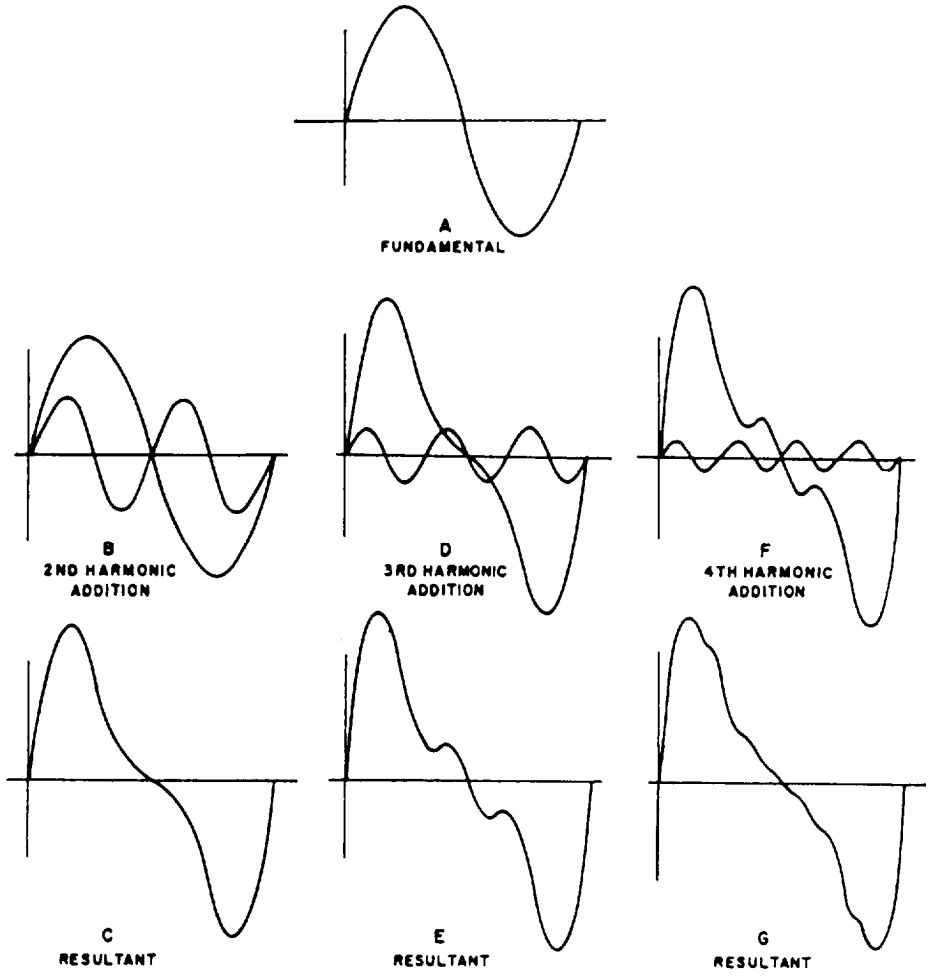


Figure 10-15.-Formation of sawtooth waveform.

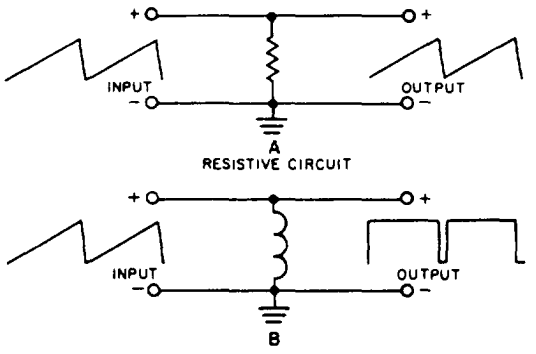


Figure 10-16.-Output current waveforms for resistive and inductive circuits resulting from a sawtooth voltage input.

sawtooth waveform is essentially a rectangular waveform.

Figure 10-17 shows the algebraic addition of a sawtooth waveform to a rectangular waveform in order to produce a resultant trapezoidal waveform for application to a series resistive-inductive circuit, the output of which is a sawtooth current waveform.

The trapezoidal waveform occurs in numerous varieties because of the amplitude differences of the sawtooth voltages and rectangular voltages prior to algebraic addition. A comparison of two varieties of trapezoidal waveforms is shown in figure 10-18. The

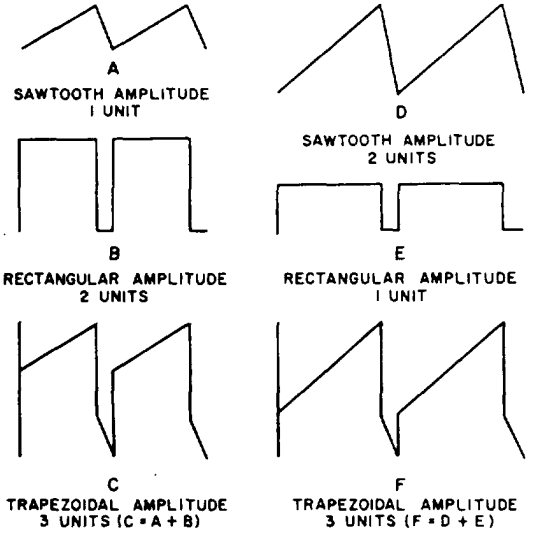


Figure 10-18.-Trapezoidal voltage waveform varieties.

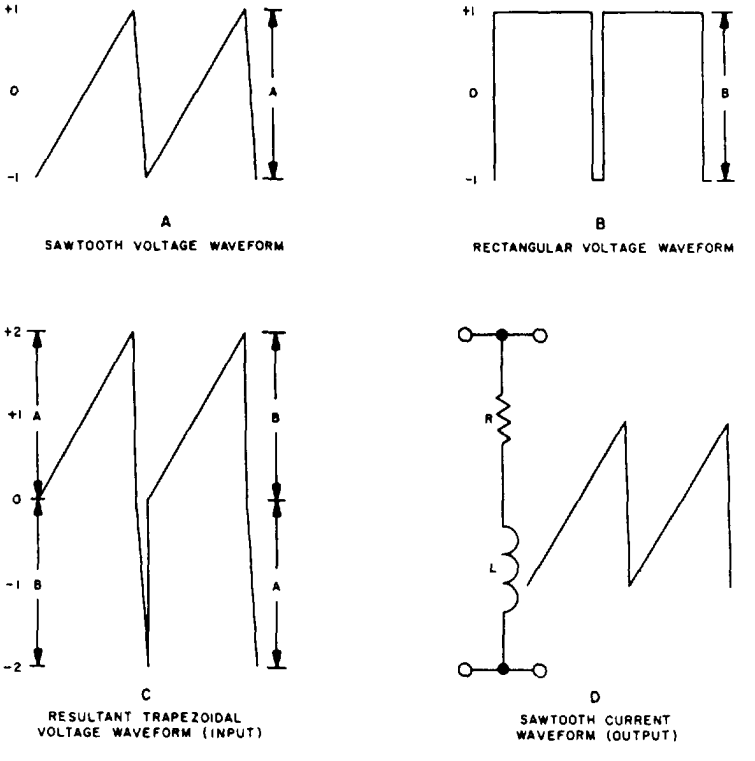


Figure 10-17.-Output current sawtooth waveform resulting from application of a trapezoidal input voltage waveform to an inductor.

resultant waveform in view C is not the same as the resultant waveform in view F. View C is the resultant trapezoidal waveform most commonly used in electronic technology. Sawtooth current waveforms are generated by deflection circuits for cathode-ray-tube deflection coils. When a deflection coil has a small internal resistance as compared with its inductive reactance, the sawtooth current waveform is produced by a sweep voltage that is a combination of a small sawtooth waveform and a large rectangular waveform.

DIFFERENTIATED VOLTAGE WAVEFORMS

Various complex waves can be resolved into their component sine-wave frequencies, and any group of frequencies can be extracted from a complex wave by means of a falter. In the case of differentiation, the differentiated waveform extracts the high-frequency sine-wave components, while the integrator extracts the low-frequency sine-wave components. A differentiated waveform is obtained by the process of differentiation. This process is simply the procedure whereby a waveform is passed through inductive or capacitive components to provide a voltage output proportional to the rate of change of the input voltage waveform.

The most popular method of differentiation employs a capacitive-resistance (RC) network. The time constant of the circuit, in microseconds, is the product of the resistance and capacitance in ohms and farads, respectively. A rapid change occurring in the input voltage waveform will produce a narrow sharp-peak (spike) in the output. The peak amplitude of the output pulse is directly proportional to the rate of change in the input waveform. Figure 10-19 illustrates a square-wave input to each of the two most common differentiating circuits, and to the output obtained from each circuit.

A square wave was used because of its rapid amplitude change and high harmonic frequency content on both the leading and trailing edges of the applied pulse. The flat horizontal portions of the square wave will produce zero output because they contain zero slope (change), and because the time constant of the differentiator will not pass the lower frequencies contained in the square wave.

A sine wave is not used as an input voltage to a differentiation circuit because these circuits accomplish their function by shifting the phase of the input waveform. In the case of a sine wave, the output will be shifted in phase and will have a smaller amplitude, but will still be a sine wave. This output maybe particularly useful as a phase-shifted wave formed from a continuously variable time-constant function.

RECTANGULAR VOLTAGE WAVEFORMS

The rectangular waveform has the same characteristics as the square waveform with respect to differentiation. It is important to note that the output voltage has a sharp peak only when the differentiating circuit contains a short time-constant. Also, it should be realized that the sharp peak is produced only during the rapid rise or rapid fall of the input voltage. Therefore, the differentiator circuit is known as a “peaker” circuit. In a circuit containing a time constant of less than one-tenth the time required for one cycle of input voltage, the time constant is said to be “short”; it is said to be “long” if the circuit components permit a time of 10 times the duration of one cycle of input voltage.

A square waveform produces an output differentiated wave with evenly spaced positive and negative excursions, whereas a rectangular waveform produces a positive and negative peak spaced close

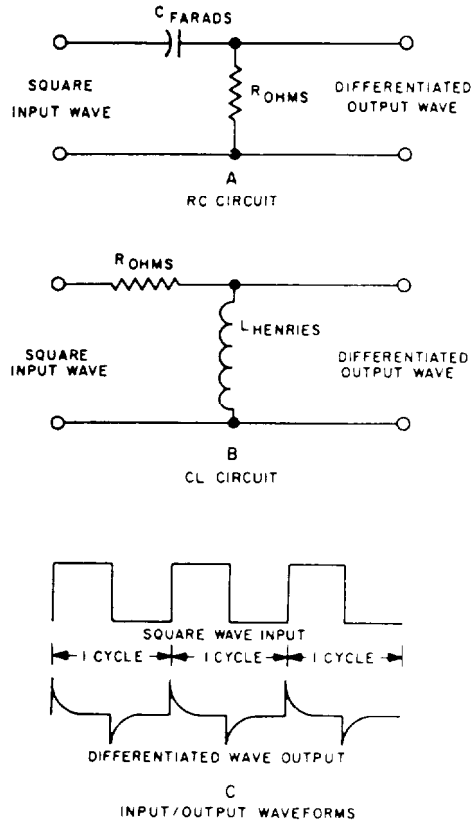


Figure 10-19.-Input to and output from diffrentiating circuits.

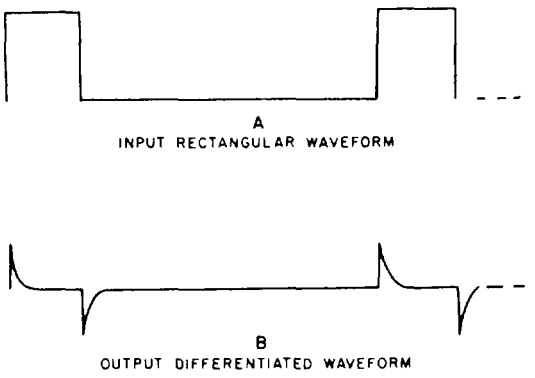


Figure 10-20.-Rectangular input and resultant differentiated out put waveform.

together (paired), with a distance separation from the next pair of peaks as shown in figure 10-20.

Sawtooth Voltages

When a sawtooth voltage is applied to a differentiating circuit containing a short time-constant, the output from that circuit will be a rectangular waveform. If the applied sawtooth is positive-going, the negative spike of the output rectangular waveform will increase in amplitude as the retrace duration time is made smaller. This condition is shown in figure 10-21. However, as the time-constant of the differentiated circuit is increased, the output progressively takes on the appearance of the input sawtooth waveform, shown in figure 10-22.

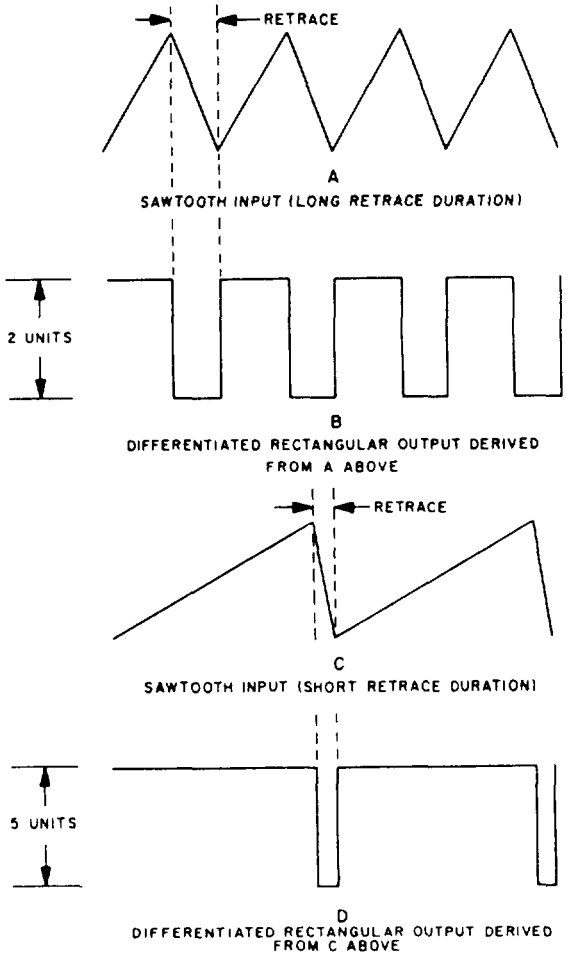


Figure 10-21.-Differentiated wave amplitude changes resulting from sawtooth input rate-of-change.

Resistor-Inductor Differentiation

The RL differentiator consists of a resistor and an inductor in series, and serves the same purpose as the RC differentiator. The output of the RC circuit is taken from across the resistor, whereas the output of the RL circuit is taken from across the inductive element. However, using either form of differentiator, the time-constant of the circuit represents the actual time required for the voltage to charge the capacitor in the

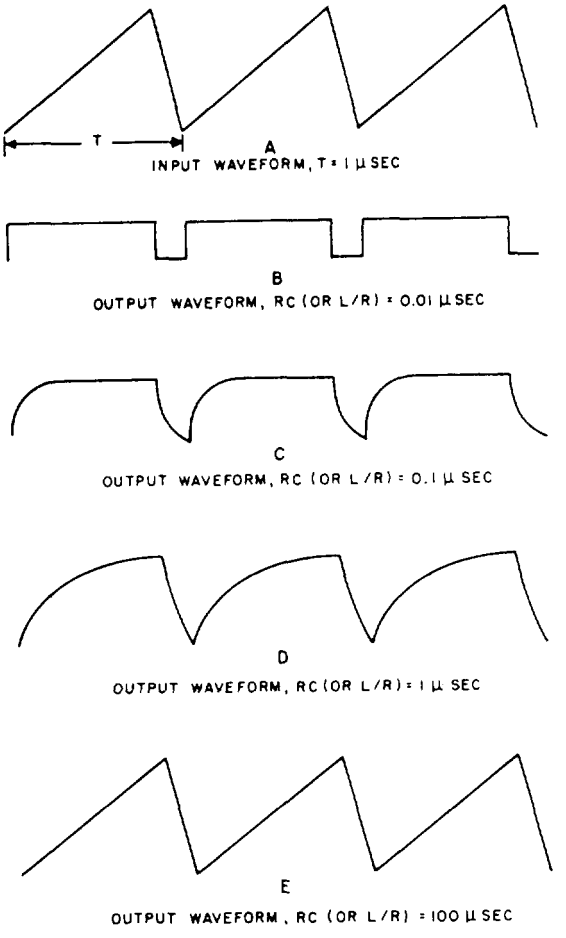


Figure 10-22.-Differentiated output waveforms for sawtooth input waveform progressively illustrating an increasing RC or RL circuit time-constant.

RC network or for the current to charge the coil in the RL network.

The actual time-constant of a differentiator (in microseconds) can be obtained by use of the applicable formula:

$$T = R\ C$$

or

$$T = L / R$$

where R is in ohms, C is in microfarads, and L is in microhenries. The shape of the voltage waveform across the capacitor and the waveform of the current through the coil are identical. Therefore, technical data pertaining to the output voltage waveform of the RC network is the same for the output current waveform of the RL network.

INTEGRATED VOLTAGE WAVEFORMS

In contrast to the differentiator circuit, which is actually a high-pass falter, the integrator sums up the applied voltages and discriminates against high frequencies, which makes it a low-pass filter. The integrator can use exactly the same components as the differentiator circuit. However, in the case of an RC integrator circuit, the output is taken from across the capacitor, whereas in the differentiator circuit, the output is taken from across the resistor. The reverse is also true of the RL integrator circuit. The output is taken from across the resistor, whereas in the differentiator circuit, the output was taken from across the coil. Figure 10-23 illustrates the two common forms of integrator circuits.

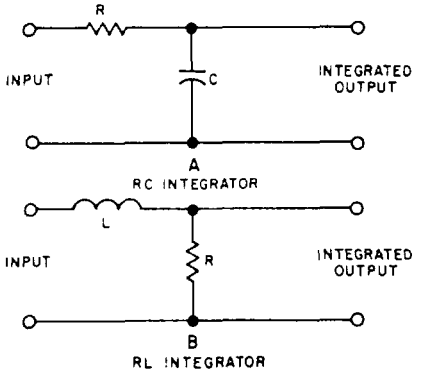


Figure 10-23.-Typical Integrator circuits.

When a square wave is applied to an integrator circuit and the integrator circuit time-constant is increased, the output waveform will gradually take on the appearance of a sawtooth waveform and will decrease in amplitude. However, the shorter the circuit time-constant, the more closely the shape of the output will resemble the shape of the input. Figure 10-24 shows the various forms and representative amplitudes for an input square or rectangular waveform.

When a square waveform is applied to the input of an integrating network having along time-constant, the output waveform will approximate a sawtooth wave with the charge or trace portion equal to the discharge or retrace portion. However, if a rectangular waveform is used as the input to the same integrating circuit, the output will not have equal charge and discharge times; therefore, the resultant waveform will build up in either a positive or negative direction. A positive buildup of the output waveform as a result of an input rectangular waveform with longer positive pulse durations than negative pulse durations is illustrated in figure 10-25.

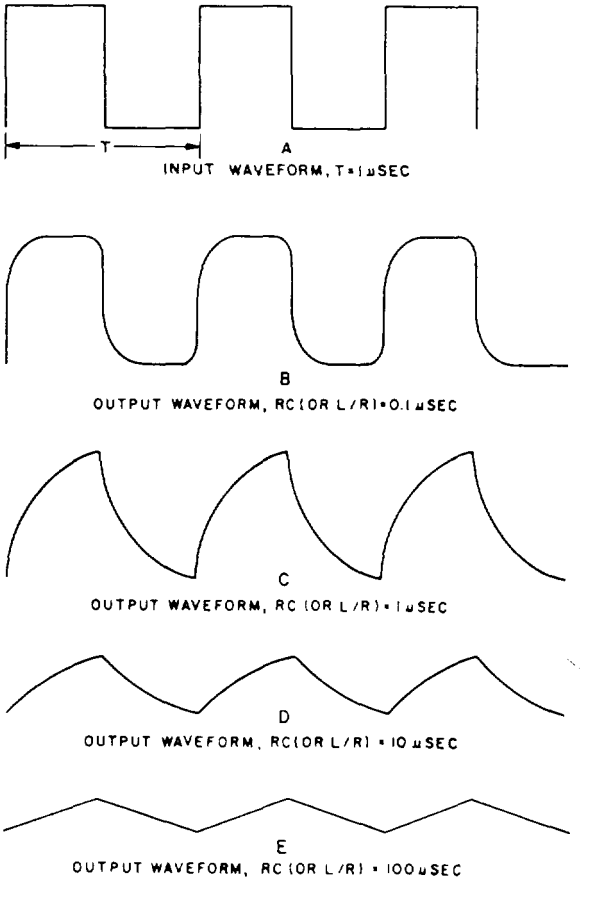


Figure 10-24.-Integrated output waveforms progressively illustrating an RC or RL circuit time-constant.

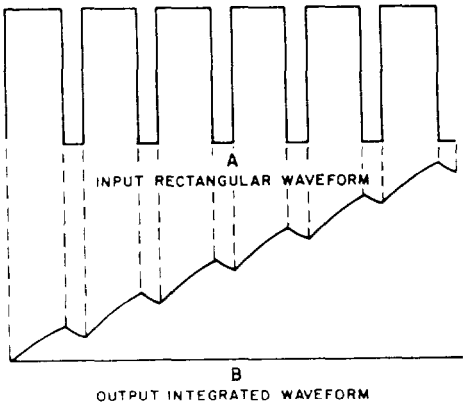


Figure 10-25.Cumulative wide integrated pulse obtained from narrow pulse rectangular waveform.

The shorter time duration provided by the negative portion of the input rectangular pulse will provide less time for the output discharge, and, as a consequence, the output waveform will charge more rapidly to its maximum value.

MODULATED WAVEFORMS

The art of superimposing on, combining with, or changing the original earner frequency by the addition of intelligence in the form of electrical energy is termed *modulation.* The three primary types of modulation are amplitude modulation, frequency (or phase) modulation, and pulse modulation.

Amplitude Modulation

The radio-frequency (RF) carrier is normally generated with the characteristics of a constant frequency and a constant amplitude. However, the amplitude of this carrier can be varied in direction in direct accordance with the intelligence to be transferred (the spoken word, music, etc.) by simply adding the amplitude of the intelligence algebraically to the amplitude of the RF carrier. This is accomplished by means of some form of mixing circuit. Figure 10-26 illustrates a hypothetical waveform representing the intelligence; a hypothetical RF carrier; the resulting composite signal obtained by algebraic addition of the carrier and intelligence frequencies; and the spectral display obtained through mixing the two signals. This resultant amplitude-modulated waveform is transmitted and received, and the amplitude is then detected (separated from the carrier) and converted back to a facsimile of the original intelligence information.

The amplitude-modulated carrier illustrated in view C of figure 10-26 is actually a composite of one carrier frequency component, view B, in addition to two modulating frequency components originally contained in view A. The two frequency displacements from the carrier, as shown in view D, are the sidebands, upper and lower, generated from linear mixing of the carrier and the modulation voltages. For example, if the carrier frequency shown in view B of the figure is 1000 kilohertz and the modulation voltage shown in view A is 10 kilohertz, the resultant modulated RF carrier shown in views C and D will contain the original carrier frequency component of 1000 kilohertz, plus a lower sideband component obtained by algebraically subtracting the modulating frequency from the carrier frequency (1000 kHz - 10 kHz = 990 kHz) and an upper side band component obtained by algebraically adding the modulating frequency to the carrier frequency (1000 kHz + 10 kHz = 1010 kHz). If the same carrier is modulated with two modulating frequencies, such as 10 kHz and 20 kHz, the resultant modulated carrier will be composed of five frequency components (that is, 1000 kHz, 990 kHz, 1010 kHz, 980

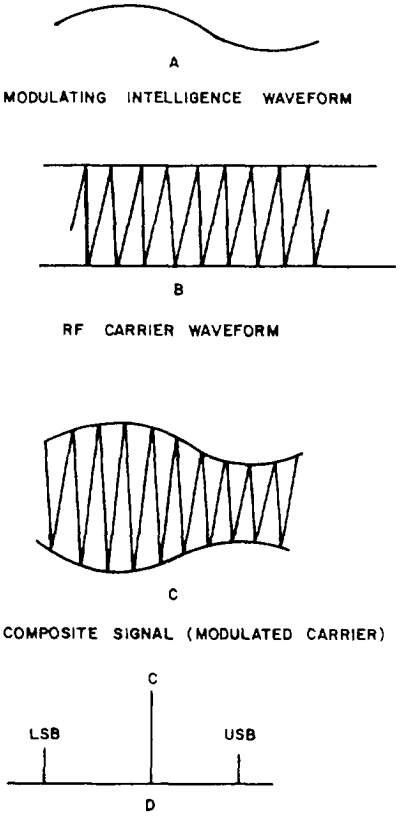


Figure 10-26.-RF carrier amplitude modulated by a sine wave.

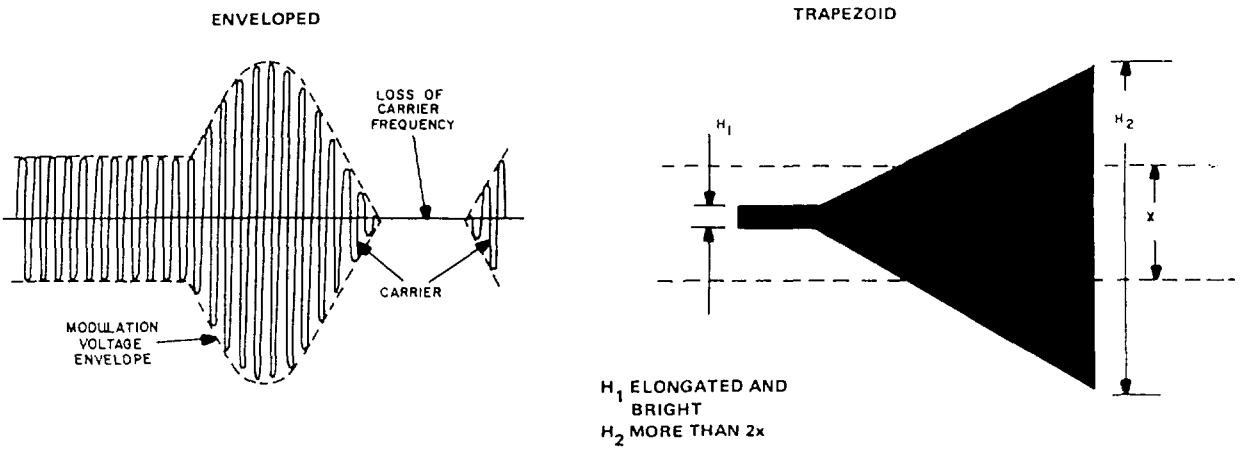


Figure 10-27.-Overmodulated RF carrier.

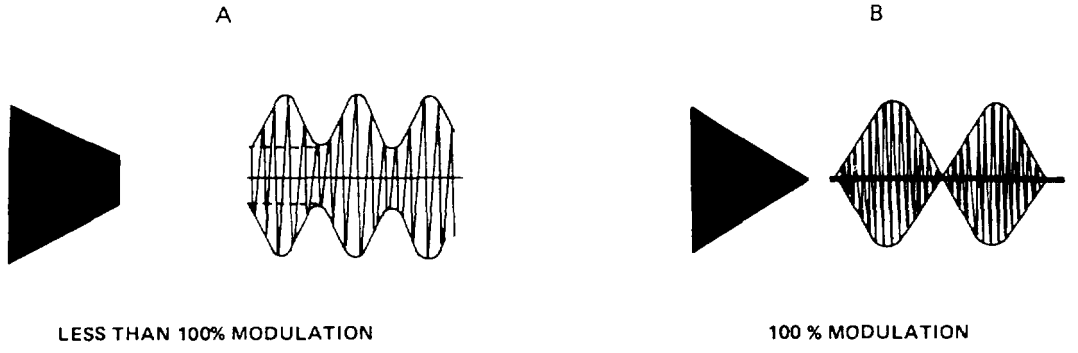


Figure 10-28.-Undistorted modulation.

kHz, and 1020 kHz). Thus the original frequencies of the modulating voltages are not apparent, and the intelligence is now contained within the parameters of the sidebands created.

Do not assume, however, that the original carrier now contains intelligence, because the carrier will be completely eliminated after it has served its purpose. The modulating voltage can increase the amplitude of the carrier any amount above and below the horizontal zero reference level without creating any operational difficulty. However, if it decreases the amplitude of the carrier to the zero reference level, it will remove the existing carrier frequency and thus create distortion. This type of distortion is termed *overmodulation,* and is shown in figure 10-27. Figure 10-28 shows the modulation pattern displayed on an oscilloscope when 100-percent modulation or less is employed.

Superimposed modulation is normally undesirable. It is readily recognized because the negative portion of the modulated carrier is not inverted. Superimposed modulation is normally encountered as a result of hum or noise modulation. Figure 10-29 provides an example of superimposed modulation without the aid of actual modulator equipment. In addition, no sidebands are evident. Thus, only the modulation frequency and the carrier frequency are present. The superimposed distortion signal shown in figure 10-29 has limited application in that this type of signal is sometimes applied to the input of an amplifier. If the output from

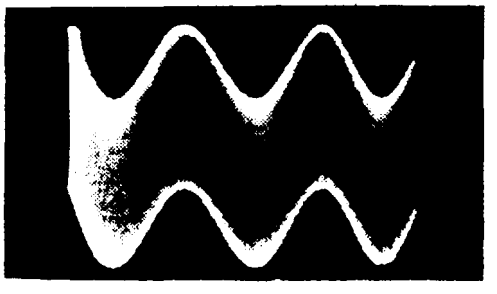


Figure 10-29.-Superimposed modulation.

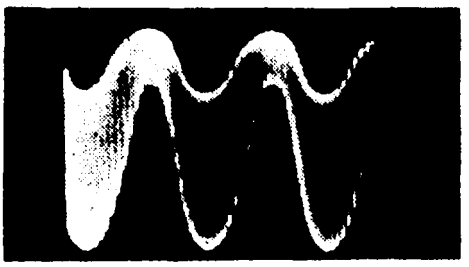


Figure 10-30.-Intermodulation distortion.

the amplifier is reasonably close to being identical with the input, the amplifier stage is relatively distortionless, and therefore linear. However, if the amplifier stage is not distortionless, or linear (amplifies all applied frequencies within its range equally), the output will be amplitude-modulated by the amplifier stage. This output will contain the upper and lower sideband-frequency harmonic components that were not contained within the original waveform. These frequency components will then modulate each other. This effect, termed *intermodulation distortion,* is shown in figure 10-30.

Frequency Modulation

This method of modulating the constant-frequency, constant-amplitude carrier will also permit the transference of intelligence. However, considering that the majority of man-made or natural noise interference is amplitude modulation, the frequency modulation method is relatively free from noise or other interference during the process of intelligence transference. In the frequency modulation technique, the modulation signal represents the intelligence by frequency changes, rather than by amplitude changes. Therefore, when the modulation signal is used to modulate the constant-frequency carrier signal, no amplitude change occurs. The resultant RF-modulated carrier signal will contain no amplitude variation, but its frequency will vary in accordance with the frequency and amplitude of the modulating voltage, as shown in view B of figure 10-31. Increasing the modulation voltage causes the carrier frequency to decrease proportionally. Decreasing the modulation voltage has the reverse effect on the carrier. Increasing the frequency of the modulation voltage will increase the rate at which the carrier changes frequencies, and decreasing the frequency of the modulating voltage will decrease the rate at which the carrier frequency changes. The percentage of modulation is not a consideration in frequency modulation since FM is measured in terms of deviation and modulation index.

Deviation is defined as the amount of shift to either side of the carrier frequency, and is directly proportional to the amplitude of the modulating signal. For example,

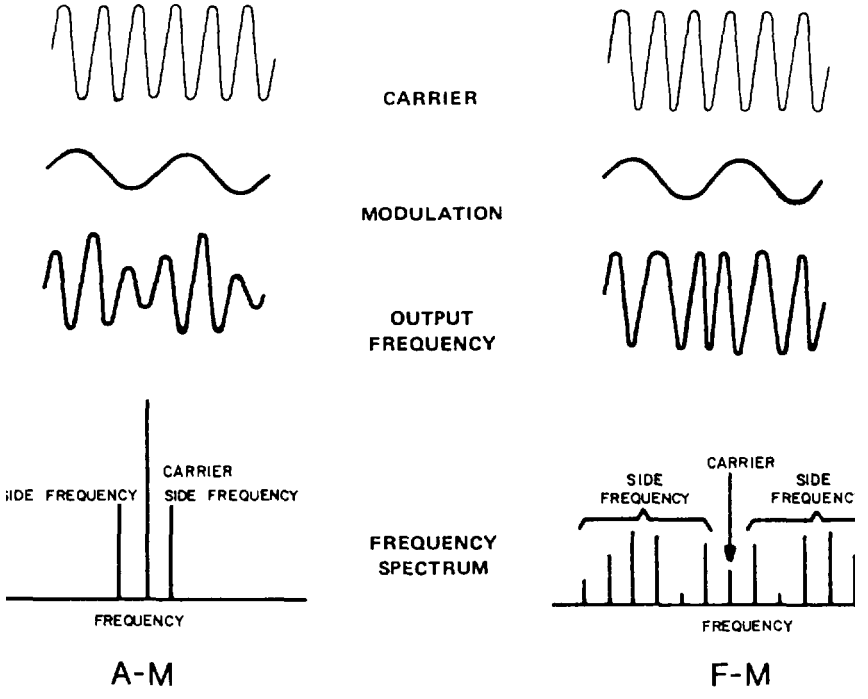


Figure 10-31.-FM patterns compared to AM patterns.

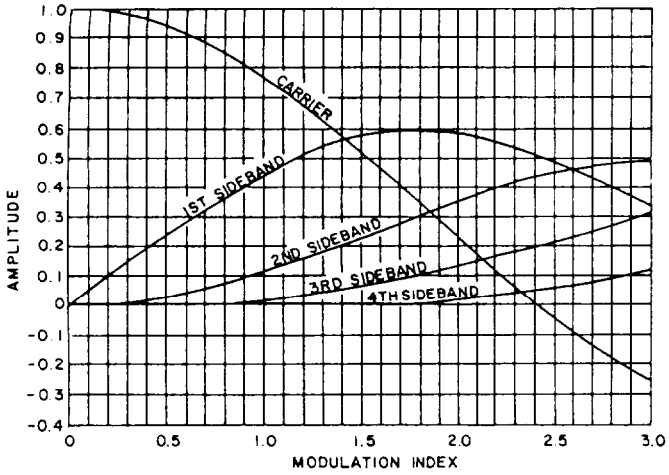


Figure 10-32.-Bessel curve for frequency modulation.

if a 1-MHz carrier were to be shifted 10 kHz to either side of its center frequency for each cycle of the modulation frequency, the resulting deviation would be 10 kHz. Modulation index, on the other hand, is defined as the ratio of deviation frequency to modulation frequency. Thus, if the same 1-MHz signal were to be modulated by 2 kHz at the same amplitude as in the foregoing example, the modulation index would then be 5 (10 kHz deviation/2 kHz modulation). Figure 10-32 shows the variations in carrier and sideband amplitude as the modulation index increases. The spectral display of figure 10-31 shows that the power within the FM spectrum is distributed throughout the sideband in an amount proportional to the amplitude of the modulation voltage. In the AM spectrum, only one-fourth of the rated output power can be attained in the sidebands, and then only with 100-percent modulation. Furthermore, the side band frequencies in the FM spectrum are separated from the carrier and from each other by an amount equal to the frequency of the modulation voltage. Theoretically, each modulating frequency creates an infinite number of sideband frequencies, but, in actuality, these are limited by the response of the transmitter circuitry.

Phase Modulation

Phase modulating of a constant-amplitude, constant-frequency carrier will result in basically the same type of transference characteristics as FM modulation. Phase modulation involves changing the carrier phase in direct accordance with the intelligence. The primary difference between frequency modulation and phase modulation is that in FM the deviation frequency of the carrier is a function of the modulation signal's voltage, whereas in phase-modulation, it is a function of the modulation signal's frequency and voltage, as shown in figure 10-33.

Pulse Modulation

Pulse modulation is accomplished by periodically interrupting the carrier frequency. Either the amplitude or the width of the pulse can be varied as a means of transferring intelligence. In some applications, both width and amplitude are varied. The most common applications of pulse modulation are found in CW (continuous-wave) keying and in radar circuitry.

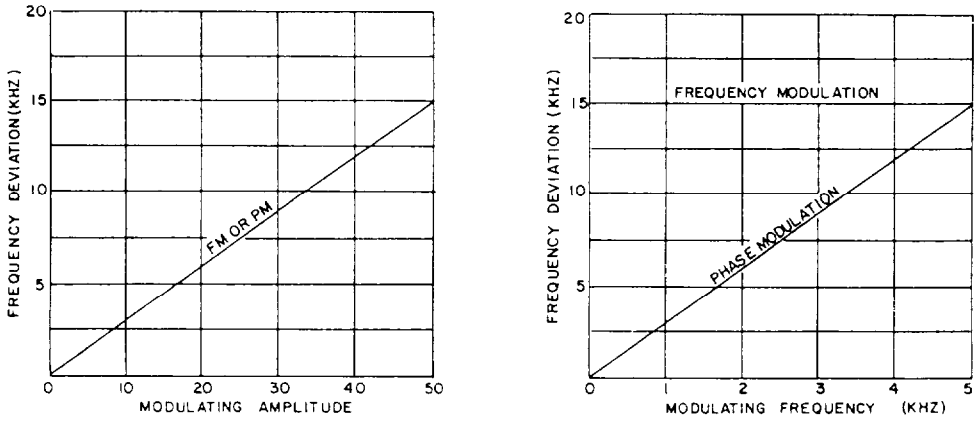


Figure 10-33.-Frequency and phase modulation characteristics.

RESPONSE AND DISCRIMINATOR WAVEFORMS

Learning Objective: *Understand response and discriminator to include single-, double-, and triple-peaked response curves, as well as discriminator type "S"* curves *and types of waveform distortion.*

A response curve is a form of graph showing the relationship between output voltage and frequency. The response curve can indicate the degree of acceptance, amplification, or rejection by either a component or a circuit as the signal frequency is varied over a desired range. There are three primary types of response curves: single-peaked, double-peaked, and triple-peaked, as shown in figure 10-34. As shown in the figure, frequency is plotted along the horizontal axis, while amplitude of the output current or voltage is plotted along the vertical axis. The circuit response for a given input frequency is the measured amplitude separation between that point on the response curve representing the frequency and the horizontal zero reference line. The amplitude of the response curve may be shown either above or below the horizontal zero reference line, as illustrated in figure 10-35. The half-power points shown in figure 10-35 are 3 dB down from the peak or maximum amplitude point on the curve. The term 0.707

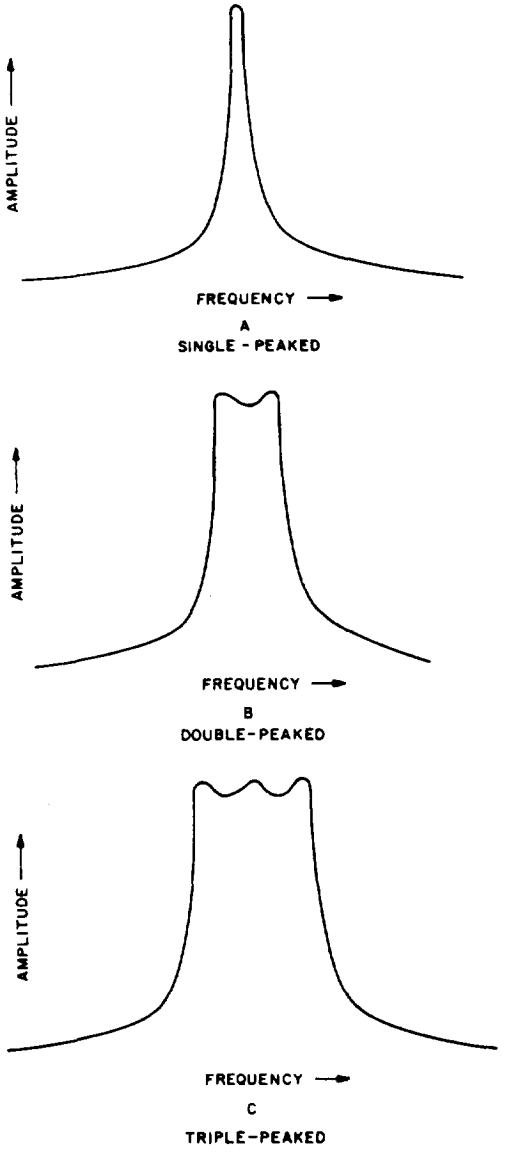


Figure 10-34.-Primary types of response curves.

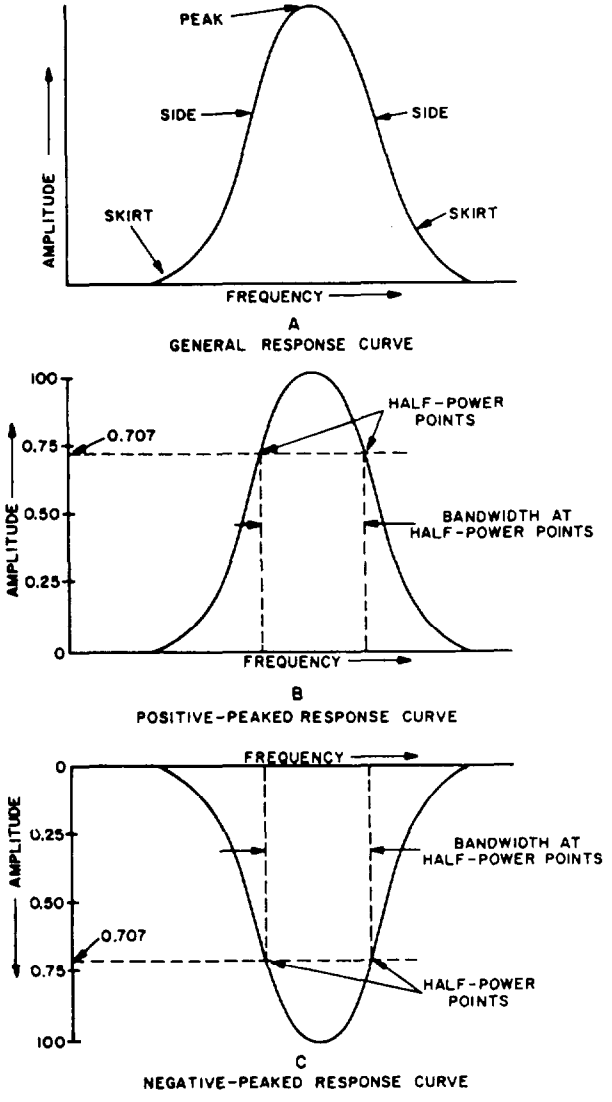


Figure 10-35.-Positive and negative single-peaked response curves.

is an amplitude value above or below the horizontal zero reference level, and is obtained from the reciprocal of the square root of two $(1/\sqrt{2})$.

A single-peaked response curve indicates the circuit is tuned to a single frequency, and will naturally provide a very narrow frequency pass band.

A double-peaked waveform is the result of the deliberate design of transformer-type circuits. A transformer-type circuit, when tuned to a single frequency, will provide a voltage maximum peak above the resonant frequency, and a voltage maximum peak below the resonant frequency. The resonant frequency will be represented by the dip between the two peaks, as shown by view B of figure 10-34. The purpose of this type of waveform is to increase the frequency pass band by increasing the amplitude of a greater number of frequencies adjacent to the center frequency. The greater the dip between the two peaks, the greater the coupling between the primary and secondary windings of the transformer. Too great a dip is undesirable.

A flat-topped curve is ideal because all frequencies within the pass band will then be the same amplitude. Several response curves may be algebraically added through a mixing circuit to produce a flat-topped, broad resultant response curve, as shown in figure 10-36.

The terms *overcoupled* (close coupling) and *undercoupled* (loose coupling) refer to the spacing between the primary and secondary windings of the transformer. For example, if the primary is brought closer to the secondary (overcoupled), all frequencies within the bandwidth will be transferred from the primary to the secondary with approximately the same amplitude; this provides a wider pass band, less frequency selectivity, and greater overall amplitude. However, if the primary and secondary windings are moved farther apart, more impedance is effectively placed between the two windings, and only the frequencies containing the greatest amplitude will have sufficient energy to bridge the gap. This will create a sharply peaked waveform in the output, representing a very narrow bandwidth, high-frequency selectivity, and less overall amplitude, even though the waveform peak is more pronounced. These effects are shown in figure 10-37.

Broadbanding, or the technique of increasing the bandwidth to permit a greater number of frequencies to pass, is accomplished by two primary methods: overcoupling, as was just discussed; and *stagger tuning.* The term stagger tuning refers to the tuning of a series of circuit stages to slightly different frequencies. For example, three stages could be tuned 1,000 cycles apart from one another, as shown in figure 10-38, to combine as a resultant waveform. This resultant waveform is considered as a triple-peaked response curve.

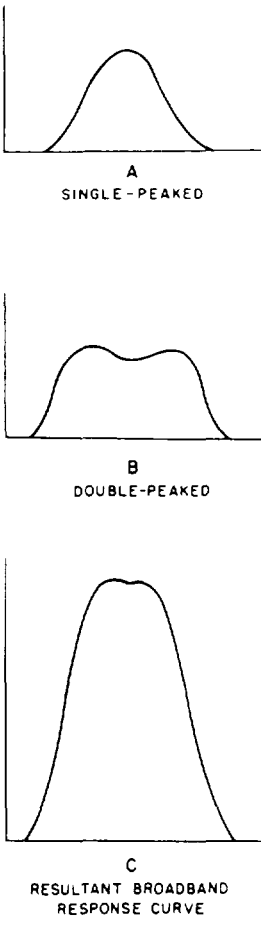


Figure 10-36.-Response curve combinations to produce a required resultant wide-band response curve.

DISCRIMINATOR CURVES

The output from a discriminator circuit is sometimes referred to as an "S" curve. Figure 10-38 shows the ideal form of the "S" curve used as a reference standard. Any deviation from this shape represents incorrect tuning of the primary or secondary transformer windings, or other improper circuit adjustment.

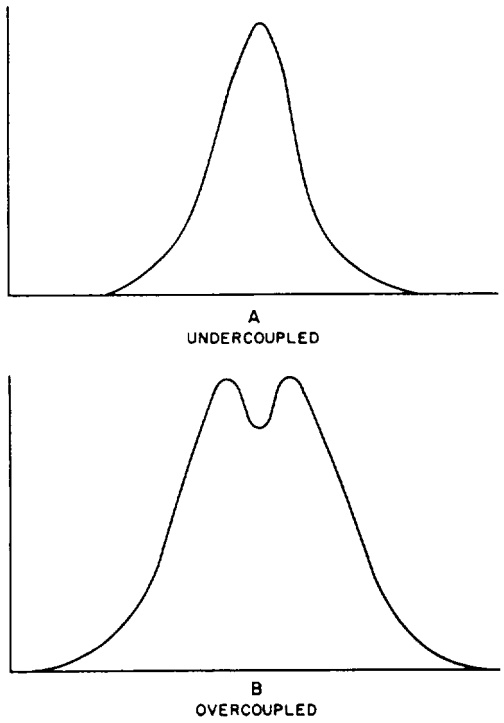


Figure 10-37.-Response curve coupling.

The "S" curve is linear and always crosses the horizontal zero reference axis at the point on the curve representing the center frequency. Many times, a marker pulse is electronically added so that it appears at some point on the curve. However, this marker will disappear

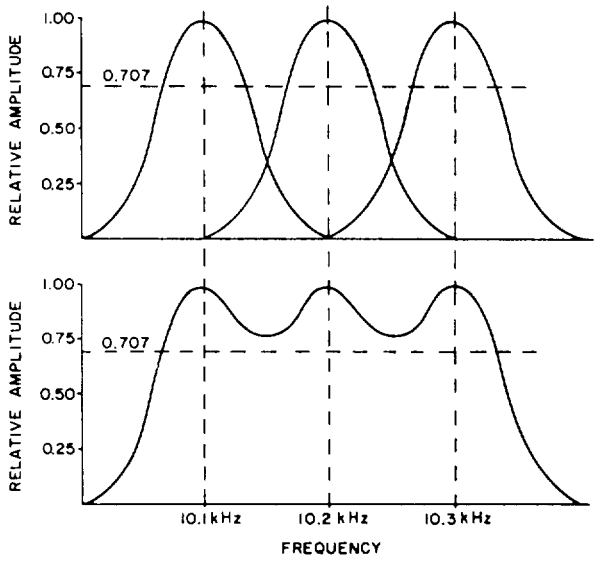


Figure 10-38.-Response curve resulting from stagger-tuned stages.

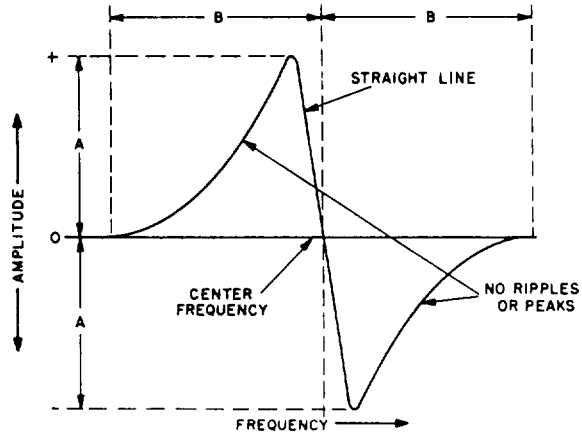


Figure 10-39.-Discriminator "S" curve.

at the center frequency because this point occurs at zero voltage amplitude. The positive amplitude and low-frequency components on one side of the center frequency should equal the negative amplitude and high-frequency components, respectively, on the opposite side of the center frequency. In other words, A = A (amplitude) and B = B (frequency separation), as shown in figure 10-39. The audio frequency response curve shown in figure 10-40 is ideal. The constant height of this response curve proves that the circuit under test has a flat response from its lowest to its highest frequency.

The horizontal zero reference base line is useful for measuring relative amplitudes. Considering that the portion of the wave below the reference or baseline is exactly the same as the wave above the baseline, only the top half of any "S" curve maybe observed for full information. Any peaks extending above the average amplitude of the waveform represent accentuation of the frequencies within that region of the pass band, and valleys or dips reflect attenuation of the frequencies at those points. Therefore, this waveform as an input not only shows the circuit behavior as a whole, but instantly reflects any unusual frequency characteristic of any

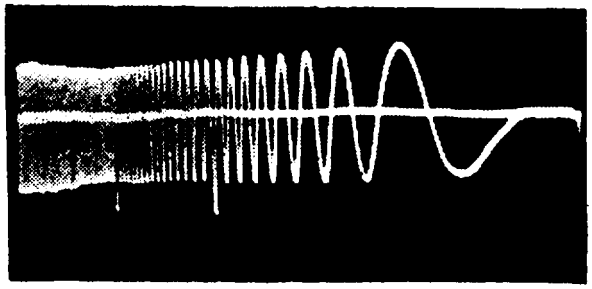


Figure 10-40.-Ideal audio-frequency response curve.

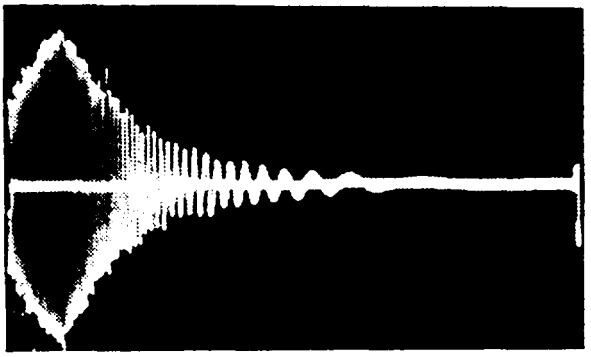


Figure 10-41.-Resonant circuit audio-frequency response curve.

recently-added components, filters, or circuits. For example, figure 10-41 represents an undesirable voltage-frequency characteristic within an LC filter circuit resonant at 4,000 Hz.

Video and other high-frequency response curves are similar to low-frequency (audio) response curves. However, in high-frequency curves, the frequency band pass is wider (broader), with an extremely low-frequency limit (60 Hz) and an extremely high-frequency limit (in the megacycle range). Two different types of markers may be used to designate exact frequencies. The first marker is a disturbance along the response curve at a particular frequency, whereas the second is produced by a tuned circuit that removes or absorbs energy from the response curve at a particular frequency. Both types of markers are shows on the typical high-frequency response curve shown in figure 10-42. Figure 10-43 shows still another type of high-frequency response curve. However, the curve shown in figure 10-43 has not been demodulated, and is not popular because frequency markers are very difficult to discern on this type waveform.

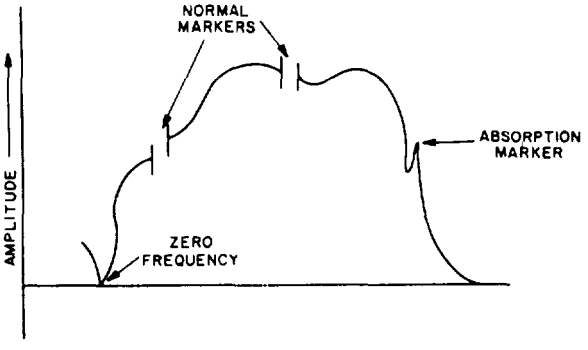


Figure 10-42.-High-frequency response curve.

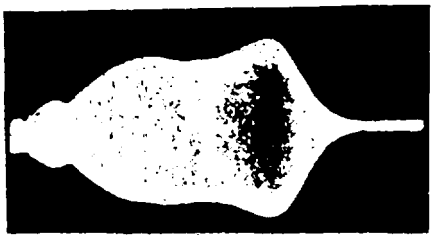


Figure 10-43.-Nondemodulated high-frequency response curve.

INTENSITY MODULATED PRESENTATIONS

The most common usage of intensity (relative brightness) modulation occurs in television. However, intensity modulation is also used to display signals on a raster (continuous display), or to brighten the mark to space transition when taking end-distortion measurements of teletype signals with a digital data distortion test set.

Intensity modulation is also employed in comparing frequencies in excess of 10:1, provided the frequency ratio involved is an integer such as 10:1, 20:1 and 30:1. In all instances, the information is obtained from the display by noting the degree of intensity of the display. This intensity can vary from zero magnitude to a very bright illumination.

Intensity modulation is accomplished by modulating the Z axis-the electron beam–of the display scope. In television, the video signal modulates the beam. In a raster display, intensity modulation could result from a received signal at a specific frequency and in a given spectrum. On oscilloscopes, intensity variation is accomplished by the signal input to the Z-axis. Shaping circuits are incorporated in the Z-axis circuitry of an oscilloscope to ensure a definitive presentation, regardless of the type of signal being applied.

COMPARING TWO FREQUENCIES

As the ratio of two frequencies being compared increases, the Lissajous pattern becomes more difficult to retain in a stationary position, and counting multiple loops becomes a more difficult task. For these reasons, the intensity modulation method of obtaining frequency ratios can be used to advantage. A circular pattern, obtained from the low-frequency signals, is passed through an appropriate phase-shifting network and applied to the vertical and horizontal inputs of the oscilloscope, as shown in figure 10-44. The high-frequency signal is connected to the intensity

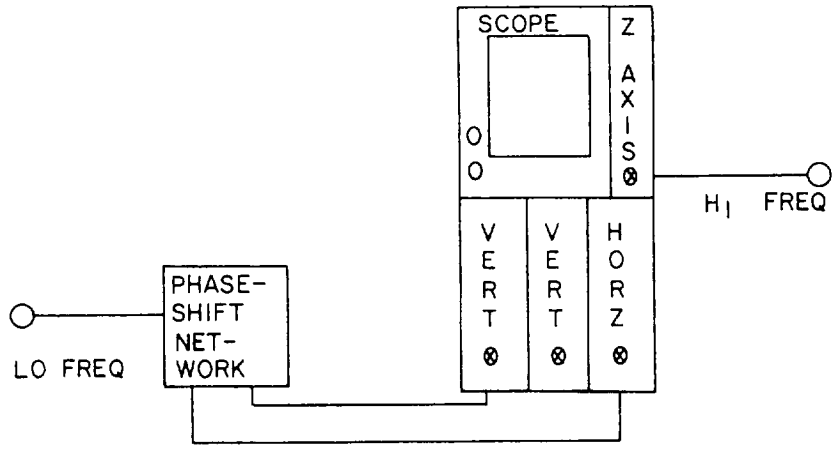


Figure 10-44.-Phase-shifting circuit for circular sweep displays.

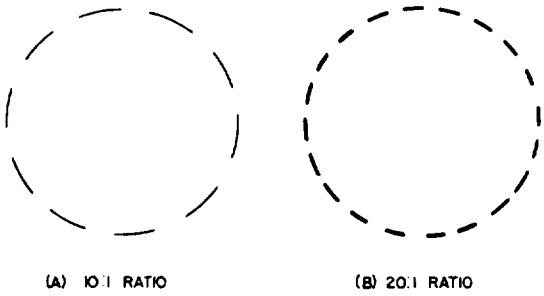


Figure 10-45.-Spot-wheel patterns.

modulation terminal of the oscilloscope, and the low-frequency signal then serves as the reference signal. In view A of figure 10-45, the frequency ratio is 10:1. There are, therefore, 10 blanked-out segments of the original circular display. In view B of figure 10-45, the frequency ratio is 20:1; therefore, there are 20 blanked-out segments in the pattern. The number of blanks in the pattern is equivalent to the ratio of frequencies. Because of this appearance, such displays are often called "spot- wheel" patterns.

CIRCULAR SWEEP PRESENTATIONS

Figure 10-46 shows the circuit connections for employing the circular sweep method of obtaining

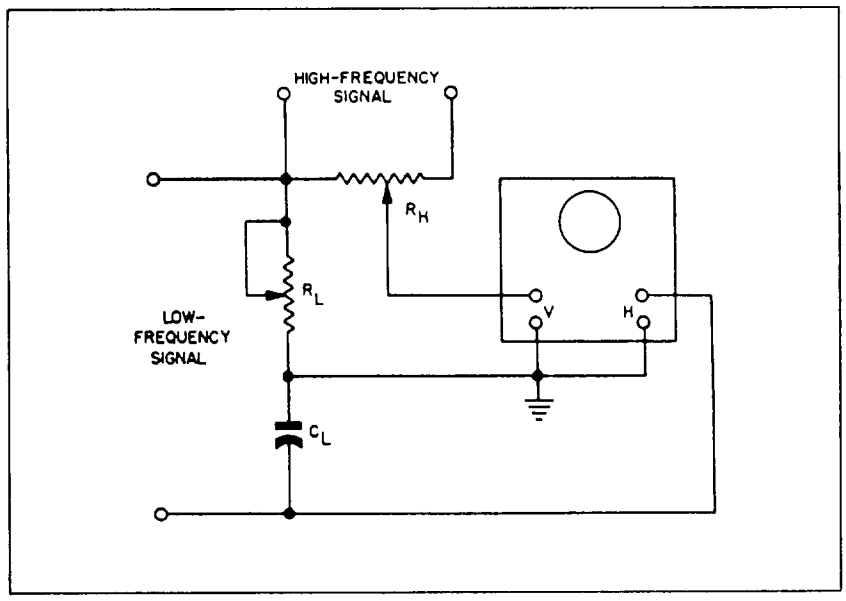


Figure 10.46.-Circular sweep comparison circuit, using deflection systems only.

frequency ratios greater than 10:1. The circular sweep is developed by the low-frequency signal. The high-frequency signal is then applied to either input terminal of the oscilloscope. Figure 10-46 shows the circuit for only one situation–the case where the high-frequency signal variations are superimposed on the circular sweep through the action of the vertical deflecting plates in the circuit. The resultant patterns of figure 10-46 are shown in figures 10-47 and 10-48.

WAVEFORM DISTORTION

Distortion is normally considered as a deviation from the desired waveform. However, the undesirable waveform in one application may be the desired waveform in some other applications. Therefore, the term distortion refers to a particular waveform application, and is meaningless if no application is being considered.

A normal high-frequency current is characterized by its amplitude, frequency, and phase relationships, and can be altered by changing any one of these characteristics. Actually, any two (or possibly all three) characteristics may be altered by a circuit change. If the circuit change produces the desired signal, this new signal is termed an *undistorted* or pure waveform; if the circuit change produces an undesired signal, it is termed a *distorted* waveform. The factors contributing to waveform distortion in one application maybe the same factors required to produce a desired waveform in some other application. The following paragraphs discuss only those undesirable factors which contribute to distortion.

The primary cause of distortion created within an actual circuit can be traced to overloading of the active component in that particular circuit. For example, overloading an active component will cause it to operate on a nonlinear portion of its characteristic curve, and the output waveform will not retain the same shape as that of the input. The same overloading effect can occur if the active component is defective or if one or more of the applied operating inputs is incorrect.

For other than active components or circuit defects and incorrect operating inputs, distortion can be eliminated by simply decreasing the input amplitude (volume control) or intensity. The placement of components, wires, or leads may create undesirable feedback voltages in a phase relationship that results in distortion. Distortion-removing circuits, designed to eliminate feedback may be defective. Neutralization circuits may be used to remove or balance out distortion

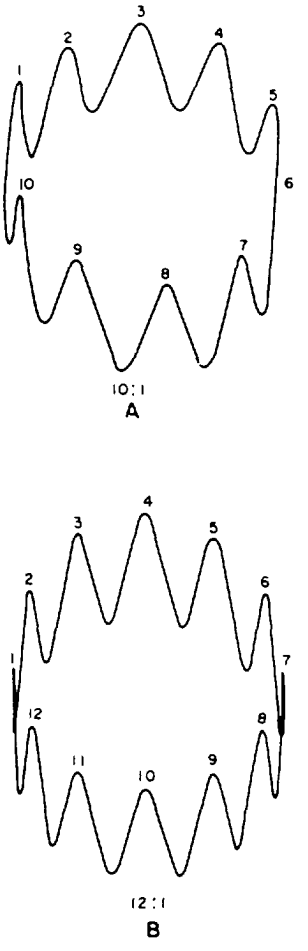


Figure 10-47.-Integral frequency ratios in circular sweeps.

resulting from undesirable feedback. The elimination or overemphasis of the amplitude of particular frequencies, within a desired band or range of input signal frequencies, will create distortion. The primary and secondary windings of frequency-sensitive transformers may be incorrectly tuned or be spaced an incorrect distance from each other. Therefore, the sideband frequencies, which form an important part of the resultant desirable signal, may be missing from the output signal. Finally, defective input or output components may blank out certain pass-band frequencies or permit undesirable voltages to pass, thereby causing distortion.

Amplitude Distortion

Amplitude distortion may be caused by a limitation of bandwidth or by irregularities within the bandwidth. In either event, amplitude distortion is normally expressed in terms of attenuation because it is a logarithmic quantity that is algebraically added for

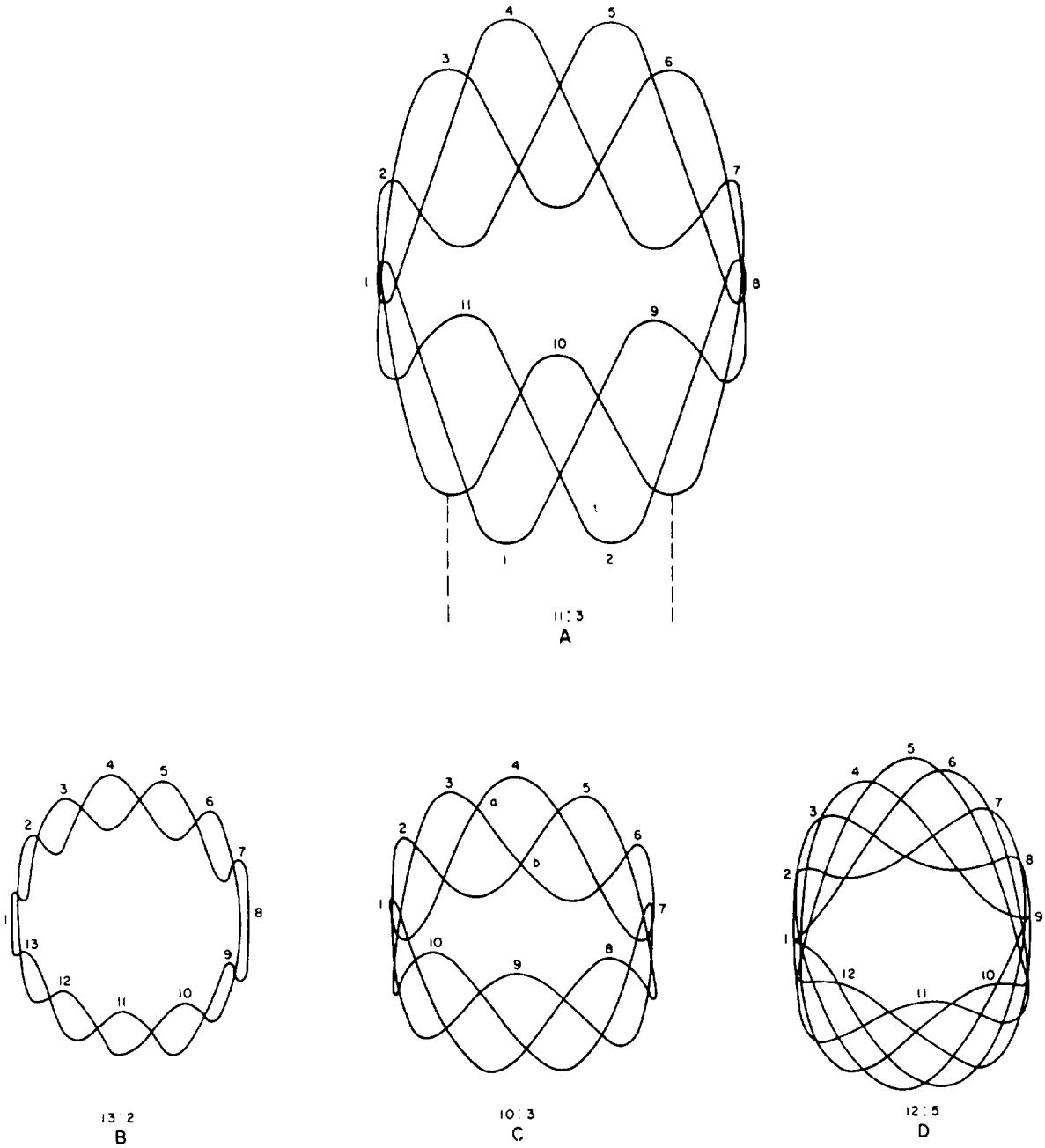


Figure 10-48.-Fractional frequency ratios in circular sweeps.

cumulative stages. Amplitude distortion, free of phase distortion, cannot change the symmetry of a symmetrical input pulse. The response of a circuit should be the same for all frequencies present in the input signal voltage. However, if the circuit response is not the same for all input frequencies, suppression or exaggeration of the amplitude of some frequencies will create distortion. The fundamental plus harmonics will be seen or heard in the output waveform when amplitude distortion exists. In the case of an amplifier stage, it can be determined whether amplitude distortion is present by applying a signal voltage of known characteristics to the amplifier input and then viewing or measuring the output signal. The output waveform should be a replica of the input waveform.

Amplitude distortion caused by an amplifier is the result of the generation, by the amplifier, of frequencies that were not contained in the input. The result of generating additional frequency components is seen by the change in waveform amplitude.

Frequency Distortion

Frequency distortion occurs when different frequency inputs are not all amplified equally. The distortion may be audible or inaudible, depending on the circuit frequency response limits. In addition, if the circuit output load is composed of reactive components, the low-frequency resonance and the increase in inductive resistance at high frequencies will increase the nonlinear (amplitude) distortion and modify the response. If a feedback network contains reactive elements, then the overall gain of an associated stage is a function of frequency, and frequency distortion due to feedback will be obtained. However, negative feedback even when reactive elements are present, will decrease the total circuit distortion at the expense of maximum gain.

The distortion in linear amplifiers as a result of the relationship between the input voltage and output voltage is a type of frequency distortion as well as of amplitude distortion. With a square waveform applied to the input of a linear circuit, the output should also be a square waveform. However, if the circuit response is not the same for all frequencies, the output waveform will not be the same shape as the input waveform. Figure 10-49 shows output waveforms for nonlinear circuit response.

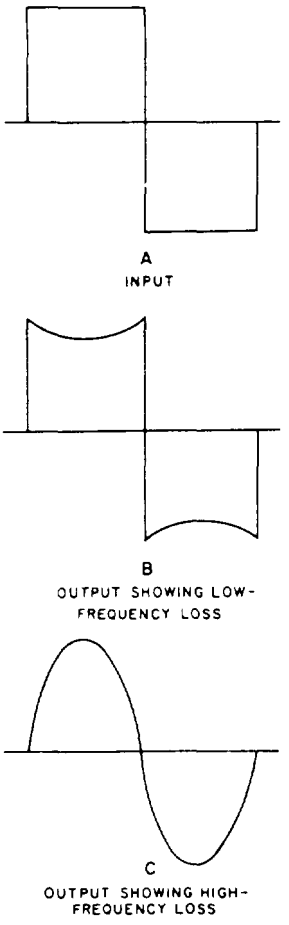


Figure 10-49.-Output waveforms resulting from poor circuit frequency response.

Frequency modulation distortion is often termed *flutter distortion*. This type of frequency distortion is generally the result of speed fluctuations as a recording is driven by the recorder or reproducer motor. The flutter effect may also be caused by a loudspeaker when it is reproducing two frequencies simultaneously. This is true because the sound pitch is a function of the relative velocities and sources with respect to the listener. Both linear and nonlinear loudspeakers produce this type of distortion.

Interference Distortion

Figure 10-50 shows two signals, separated slightly in frequency and differing in amplitude. The third waveform is the resultant obtained when the desired and undesired signals are combined algebraically at every point. The amplitude of the resultant varies at a rate equal to the difference in frequency between the two original signals. If both signals differ in frequency by 100 Hertz, the resultant waveform amplitude will change 100 times per second.

In an AM receiver, this amplitude would be separated by the detector and be heard as a whistle from the speaker. The resultant waveform may at times lead, lag, or be in phase with the desired signal; therefore, the resultant is phase-modulated. This phase modulation (and, indirectly, frequency modulation) is directly proportional to the amplitude difference between the two signal carriers. When the amplitude ratio between

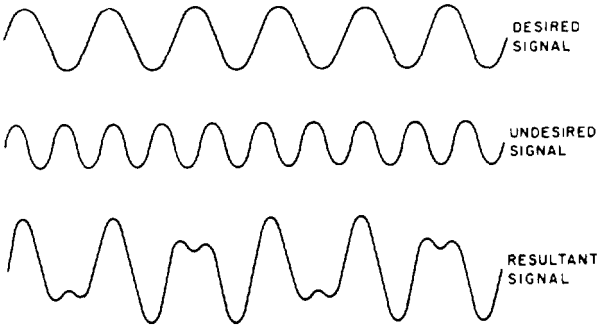


Figure 10-50.-Combination of two signals forming an amplitude-and-phase-modulated resultant.

the two signals is 2 to 1, the phase angle shift is slightly under 30 degrees. The rate of phase shift change is in direct proportion to the frequency difference between the two original signals.

Static is primarily a form of amplitude distortion caused by uncontrolled electrical waves associated with thunderstorms and other natural phenomena. The strength of these waves is sometimes great enough to drown out the desired station or prevent clarity of reception. Limiter stages will limit incoming bursts of static amplitude, and, by selecting a narrow bandwidth, much of the continuous crackling variety of static can be removed through frequency selectivity. For FM reception, transmission allocations are in the higher frequency bands, where static amplitude changes are not very effective; most of the outburst energy is limited to lower frequencies. Even if no external natural disturbances or other station interference is present, internal active component and circuit noises exist that will limit the weakest received signal to some minimum amplitude. Any signal lower than this minimum amplitude will not be amplified with clarity.

Thermal agitation is the term applied to the noise created by the random motion of electrons in any conductor. The thermal noise produced is proportional to the amplifier bandwidth.

Tube hiss is the term applied to the noise created by *shot effect,* which refers to the fact that electrons moving through a vacuum tube are a congregation of separate particles that do not impinge on the anode as a continuous fluid movement, but rather as sporadic fluctuations. This tube hiss noise is normally distributed evenly throughout the frequency spectrum.

Transistors generate noise by the shot effect in the bias current, and their thermal noise is caused by inherent resistance in the base region. Surface recombination of electrons may also be a source of semiconductor noise that only becomes significant at very low frequencies.

Impulse noise, as distinguished from random noise, consists of external sharp bursts of energy. Normally, this noise is associated with automobile ignition systems and sparking gaps in electrical machines. A limiter stage is required to decrease the effects of this type of interference.

Hum interference is normally a result of insufficient filtering in the power supply, of heater-cathode leakage of unshielded transformers and chokes in the DC line, or of defective coupling between circuits. This type of interference, as is true of other types of circuit noise, will combine with the desired signal and produce distortion.

USE OF LISSAJOUS FIGURES

Learning Objective: *Understand the use of Lissajous figures to include phase relationship of Lissajous patterns.*

Lissajous patterns are a useful method of determining the frequency ratio of one signal to another (sine wave). If one of the signals is known, the other can be determined from the displayed Lissajous pattern. The known signal is applied to the horizontal axis input of the oscilloscope, and the unknown is applied to the vertical deflection input. Figure 10-51 shows examples of various Lissajous patterns. In the first five examples, the ratio is always a selected multiple of one. The sixth example shows that odd ratios can also be displayed.

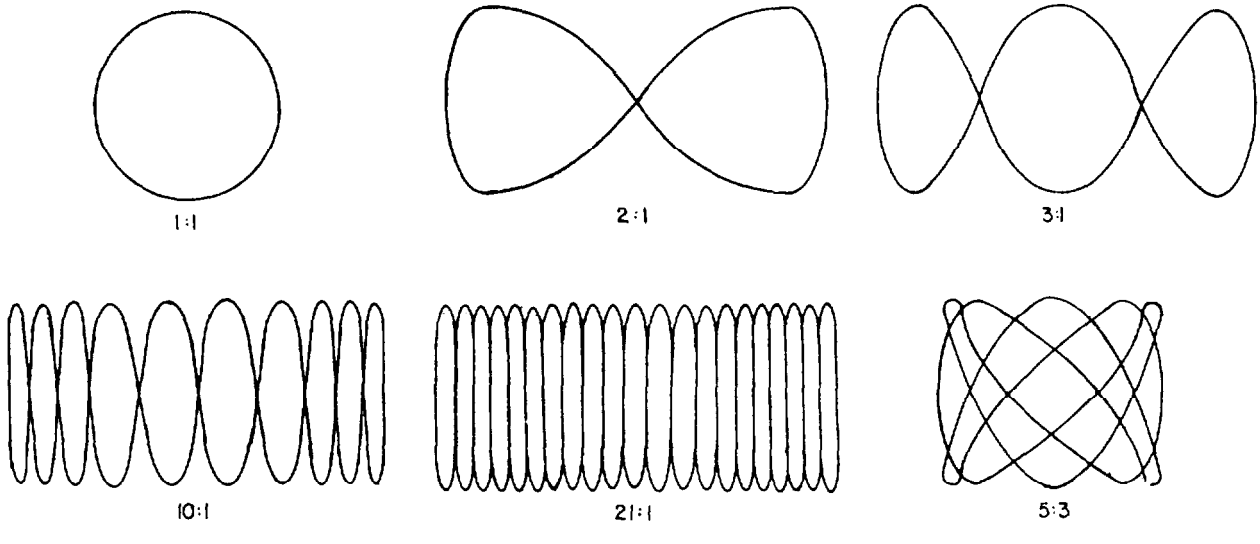


Figure 10-51.-Lissajous patterns, showing frequency ratios.

The accuracy of frequency measurements obtained by Lissajous patterns is limited by the accuracy of the reference frequency and by the care exercised in obtaining a stationary display and in counting the loops. The practical ratio limit in this type of measurement is 10: 1; however, by using extreme care, it is possible to count frequency ratios as high as 30:1.

Lissajous figures can be used to measure the phase relationship existing between two voltages of the same frequency. The patterns involved appear as ellipses with different degrees of eccentricity. As shown in figure 10-52, the pattern is formed when two sine waves of the same frequency are applied to the vertical and horizontal input terminals of the oscilloscope. Point-to-point plotting of like-numbered projections will verify the formation of the resultant pattern.

To measure the angle of phase displacement, it is necessary to us an oscilloscope with a cross-section screen, called a *graticule,* to provide a graph of the X- and Y-axis coordinates. If two sine waves of unequal amplitude are used, the resultant pattern will always be elliptical in form and could not be used intelligently. In actual phase measure-merit, unequal amplitudes of the input to the scope are compensated for by adjusting the horizontal and vertical gain controls. The vertical gain is first reduced until a straight horizontal line is obtained. The horizontal gain control is then adjusted for some convenient length; for example, 2 inches.

The next step is to place the horizontal function or selector switch in the horizontal input position. A small spot near the center of the oscilloscope will now be obtained, depending upon the relative adjustments of the vertical and horizontal positioning controls. Apply the signal to be measured to the vertical input terminals, and increase the vertical gain control to the same length of line as previously obtained; in this case, 2 inches. Since there is no horizontal deflection, the 2-inch trace will be only a thin vertical line. At this point the gain of both amplifiers is equalized, and the technician may apply the comparison signal to the horizontal input terminals and proceed with the phase measurement technique. However, it is to the technician's advantage to make one further check for equalization.

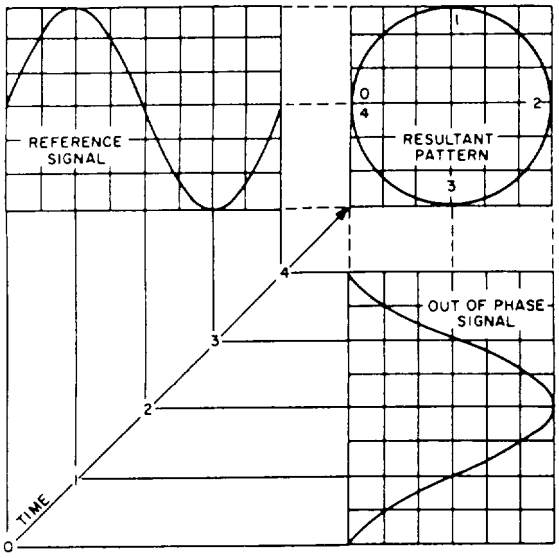


Figure 10-52.-Formation of a Lissajous figure, illustrating 90 degrees of phase difference.

Connect a jumper wire from the vertical input terminal of the scope to the horizontal input terminal so that the same signal is applied simultaneously to both amplifiers. The pattern should tilt over to a 45-degree line intersecting the corners of the 2-inch square, as shown for 0 degrees in figure 10-53. The procedure just given is not the only method of equalizing the oscilloscope amplifiers, but it is applicable to any oscilloscope, and is not subject to any specials switching positions provided on a specific oscilloscope.

If the phase angle displacement between two input frequencies remains at a fixed angle, the phase angle may be calculated. Count the number of divisions along the vertical Y axis to the point where the ellipse intersects the Y axis. This number is known as the "Y-axis intercept," or Y_1. Next, count the number of divisions along the vertical Y axis to the point that indicates the maximum vertical amplitude of the ellipse. This number is known as the "Y-axis maximum," or Y_2. The angle of phase difference, θ, is found by performing the following calculation:

$$Sine\ \theta = \frac{-Y\text{-}axis\ intercept}{Y\text{-}axis\ maximum}\ or\ \frac{Y_1}{Y_2}.$$

Reference to a trigonometric function table, or use of an electronic calculator or conventional slide rule will enable you to convert this ratio into a phase angle expressed in degrees and minutes.

A similar procedure, using the horizontal X axis in the same manner, will produce the same results. The direction in which these values are measured can be either positive or negative. The ratios obtained arc independent of the direction taken when counting. Assume the calculated ratio is 0.5, which, when converted into angles from the trigonometric tables, gives the phase angle difference of 30 degrees. Figure 10-54 shows that the major axis of one of the ellipses lies in the first and third quadrants, and that the phase angle could therefore be 30 degrees or 330 degrees. This

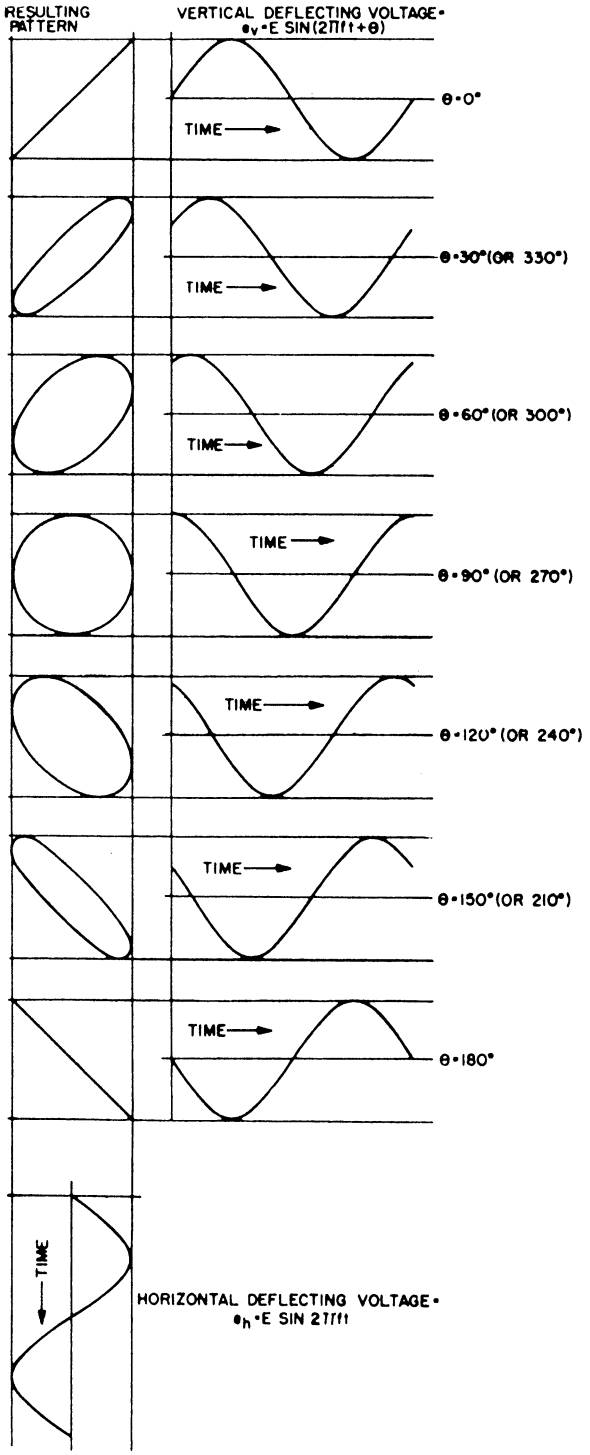


Figure 10-53.—1:1 Lissajous patterns, showing effect of phase relationships.

ambiguity of two possible phase angles is not surprising. Whether a signal is 30 or 330 degrees ahead of a second signal, the difference in phase is still 30 degrees. It is this difference that a Lissajous figure has the ability to indicate, and not which signal leads or lags the other.

Fortunately, it is not difficult to learn which is the leading signal when the information is not known from

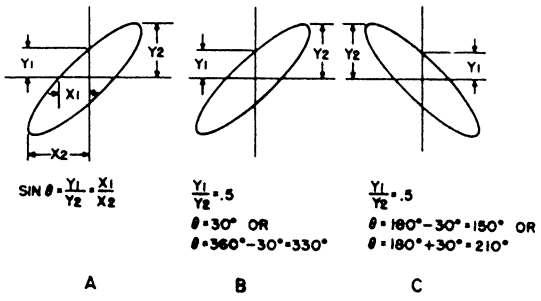


Figure 10-54.—Computation of phase difference.

other sources. Assume that a phase difference of 30 degrees is computed. If it happens that the signal applied to the vertical deflecting plates leads the horizontal signal by 30 degrees, an additional phase advancement of the vertical signal will reduce the eccentricity of the ellipse; that is, it will be made to resemble a circle. On the other hand, if the vertical signal lags by 30 degrees (equivalent to leading by 330 degrees), an advancement in phase will bring the two signals more nearly into phase. Consequently, the ellipse will continue to contract until eventually it becomes a straight line.

There are a variety of circuits for shifting the phase of a signal, such as the one shown in figure 10-55. One

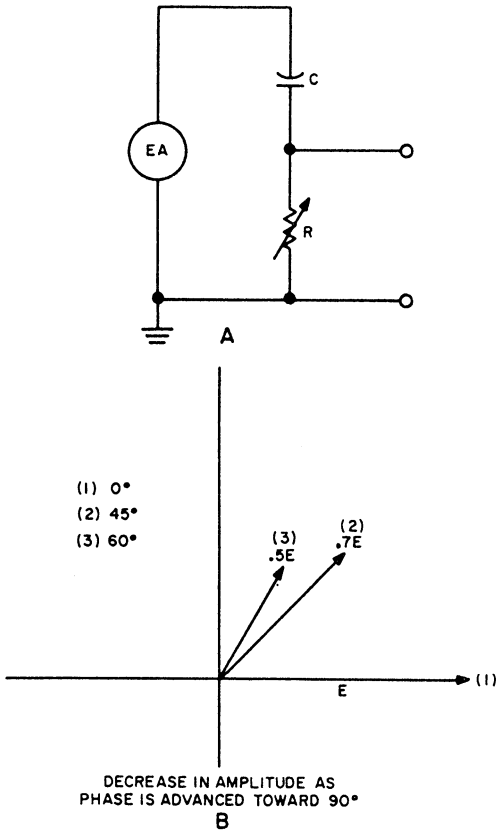


Figure 10-55.—Phase-shift network.

of the two signals under investigation, such as the signal applied to the vertical deflecting plates, can be impressed across a series circuit consisting of a potentiometer and a capacitor. At the frequency concerned, the resistance of the potentiometer should be about 10 times the reactance of the capacitor. An output from the network can be taken from either the resistance, as shown in figure 10-55, or from the capacitor. If the signal developed across the capacitor is desired, the ground connection should be made to the input side of the capacitor. If the output signal is derived from the resistance, its phase will be advanced relative to the original signal; if taken from the capacitor, the phase will be retarded. It will be assumed in this case that the signal across the resistance is applied to the vertical terminals of the oscilloscope. If the vertical signal leads the horizontal signal, the ellipse will become broader as the resistance of the potentiometer is decreased. Most likely a circle will not be obtained since the amplitude of the signal also decreases as the resistance becomes less.

In those cases where one of the two applied frequencies is constantly changing phase with respect to the other, the resultant ellipse changes form, and the plane of the ellipse appears to rotate around either of two imaginary diagonal axes. As the phase difference increases from zero to 90 degrees, the plane of the ellipse appears to rotate around one of the imaginary diagonals; and as the phase difference increases from 90 to 180 degrees, the plane of the ellipse rotates around the opposite diagonal axis.

When an oscilloscope is used to determine phase relationships, several precautions must be observed It is imperative, first of all, to know whether the circuits in the oscilloscope ahead of the deflecting plates have unequal phase-shift characteristics. If there is an inequality, the indicated phase relationship of the two signals undergoing investigation will be in error by the amount of the inequality. To determine the amount of phase error introduced by the oscilloscope circuits, apply a sine wave simultaneously to both the horizontal and vertical input terminals of the oscilloscope. If a straight line is displayed in the first and third quadrants, no phase shift is introduced by the oscilloscope amplifier. If the straight line appears in the second and fourth quadrants, a 180-degree phase shift is introduced by the amplifying stages of the oscilloscope, probably because the number of stages in the two sections are unequal. It is important to check this possibility, as the design requirements of the sections are not generally the same.

The appearance of an ellipse, however, discloses an inherent disparity in the phase characteristics of the two amplifiers, rather than a mere difference in design. This phase difference (in degrees) must be determined, and then added to or subtracted from the result of a phase measurement of the two signals according to which amplifier has the leading characteristic. This check of amplifier characteristics, which should be made over the entire frequency range of interest, is especially important in the low-and high-frequency portions of the band passed by the amplifiers.

Astigmatism in an oscilloscope may be so pronounced that accuracy in measuring Y-maximum and Y-intercept is difficult. In this case the trace maybe in focus over one region of the tube face, but out of focus in other regions. Wherever the trace is poorly defined, there will be uncertainty in a measurement of distance. For an accurate determination of the sine of the phase angle, it is necessary that Y_1 and Y_2 be measured accurately. This means that the intersection of the X and Y axes must be placed in the exact center of the ellipse.

(2:1) LISSAJOUS PATTERNS

The concept of Lissajous patterns was developed on a limited basis with respect to multiple patterns and time basis, and with respect to the phase-angle measurement in the preceding paragraphs. A detailed discussion of Lissajous figures is presented in the following text, with special attention given to frequency comparisons. There are many possible configurations for any ratio of applied frequencies.

One consideration is whether the higher or lower frequency is applied to the horizontal deflecting plates. The most significant consideration, however, is the "phase" of the high-frequency signal with respect to that of the low-frequency signal when the latter is beginning a cycle. Strictly speaking, "phase" in this sense is a misnomer, as the definition is normally in terms of a single frequency. Nevertheless, a cycle of the high-frequency signal is often well advanced at a time when a cycle of the low-frequency signal has just commenced; for convenience, this condition is usually referred to as a "difference in phase."

Figure 10-56 shows the situation that prevails when both applied signals start at the same time. The resulting pattern can be likened to a figure "8" resting on a side. In views A and B of figure 10-57, a line drawn against the top edge of the pattern, called a "tangent," would make contact with the pattern at two places. Similarly, a line drawn against a vertical side would be tangent at

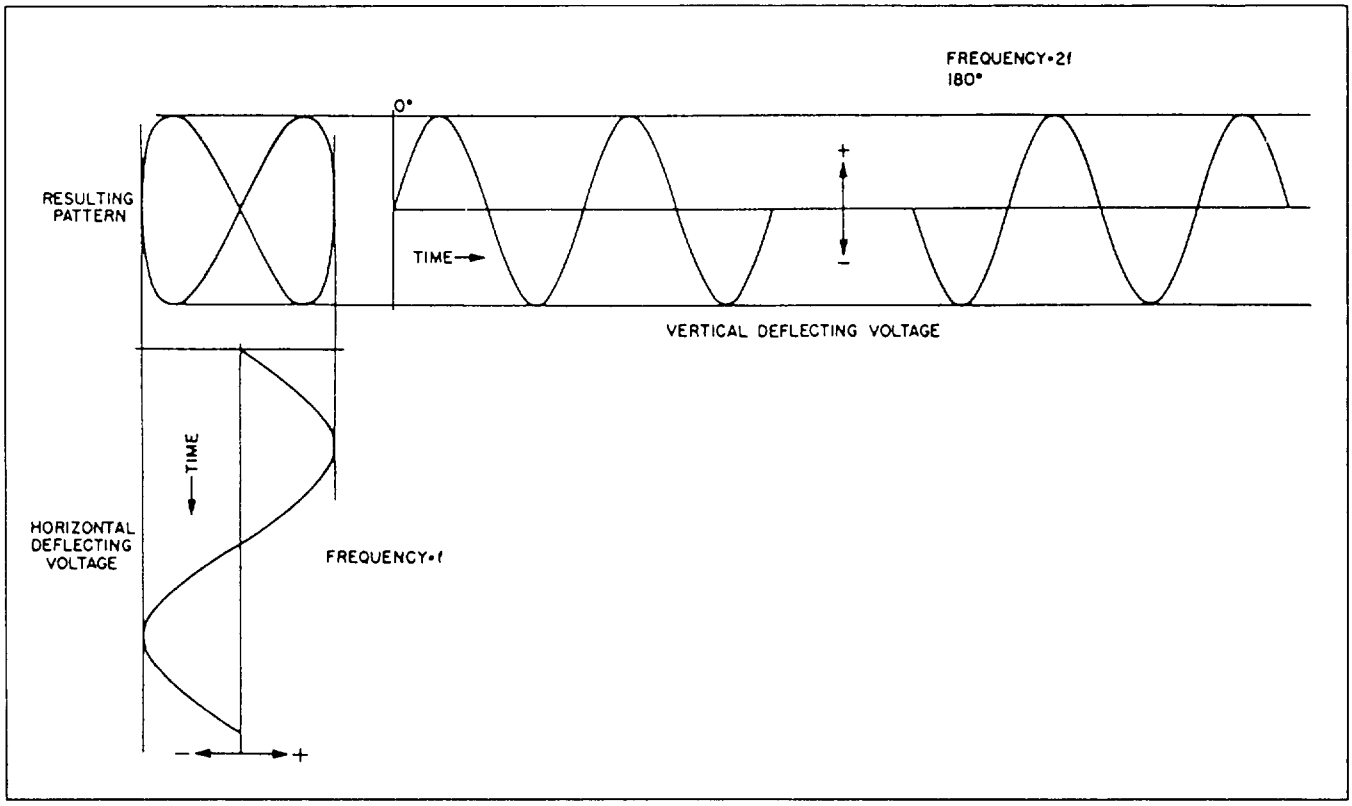


Figure 10-56.-2:1 Lissajous pattern.

only one place. Notice that the horizontal tangents correspond to the vertical deflecting voltage, and that the vertical tangents correspond to the horizontal deflecting voltage. Hence, the ratio of the vertical deflecting frequency to the horizontal deflecting frequency is 2:1. If the two signals were applied to the opposite sets of deflecting plates, the resulting pattern would be rotated 90 degrees.

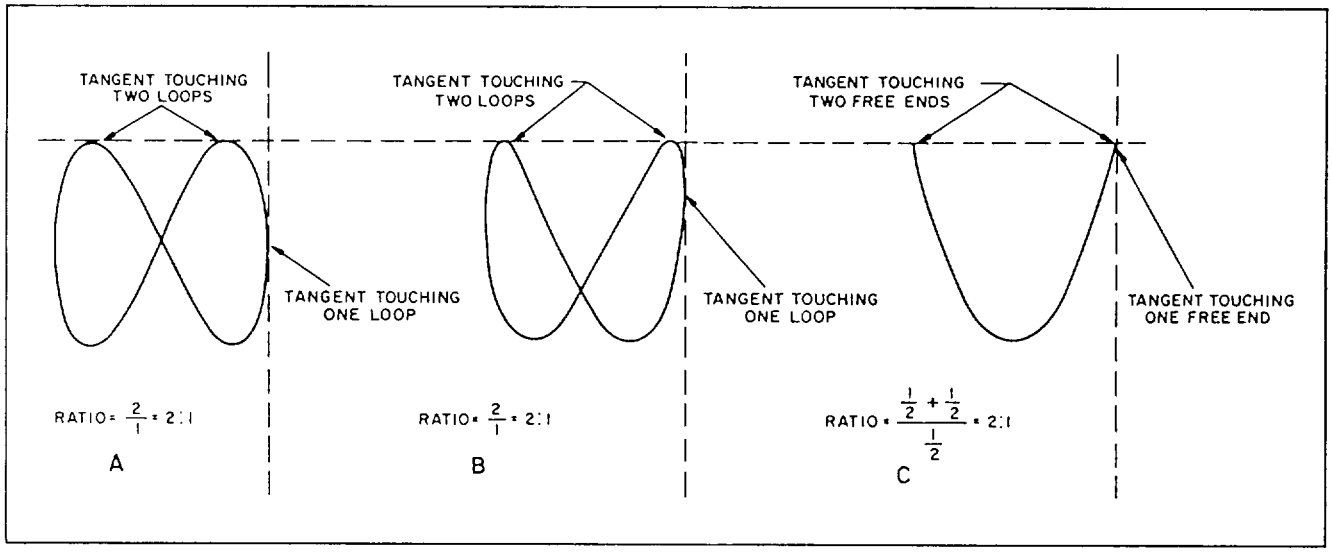


Figure 10-57.-Calculation of 2:1 frequency ratio.

An interesting situation exists when the high-frequency signal is shifted ahead 90 degrees in phase. As shown in view B of figure 10-58, the high-frequency signal may be at its maximum value when the low-frequency signal is just beginning a cycle. When this condition occurs, the two loops are closed into the form of a parabola, with its cup pointing downward. Similarly, if the high-frequency signal is at its most negative value when the low-frequency signal is commencing a cycle, the pattern is a parabola with its cup pointing upward, as shown in view D of figure 10-58. This type of pattern is commonly referred to as a "double image" because the electron beam, after reversing its direction, traces out the same path. A

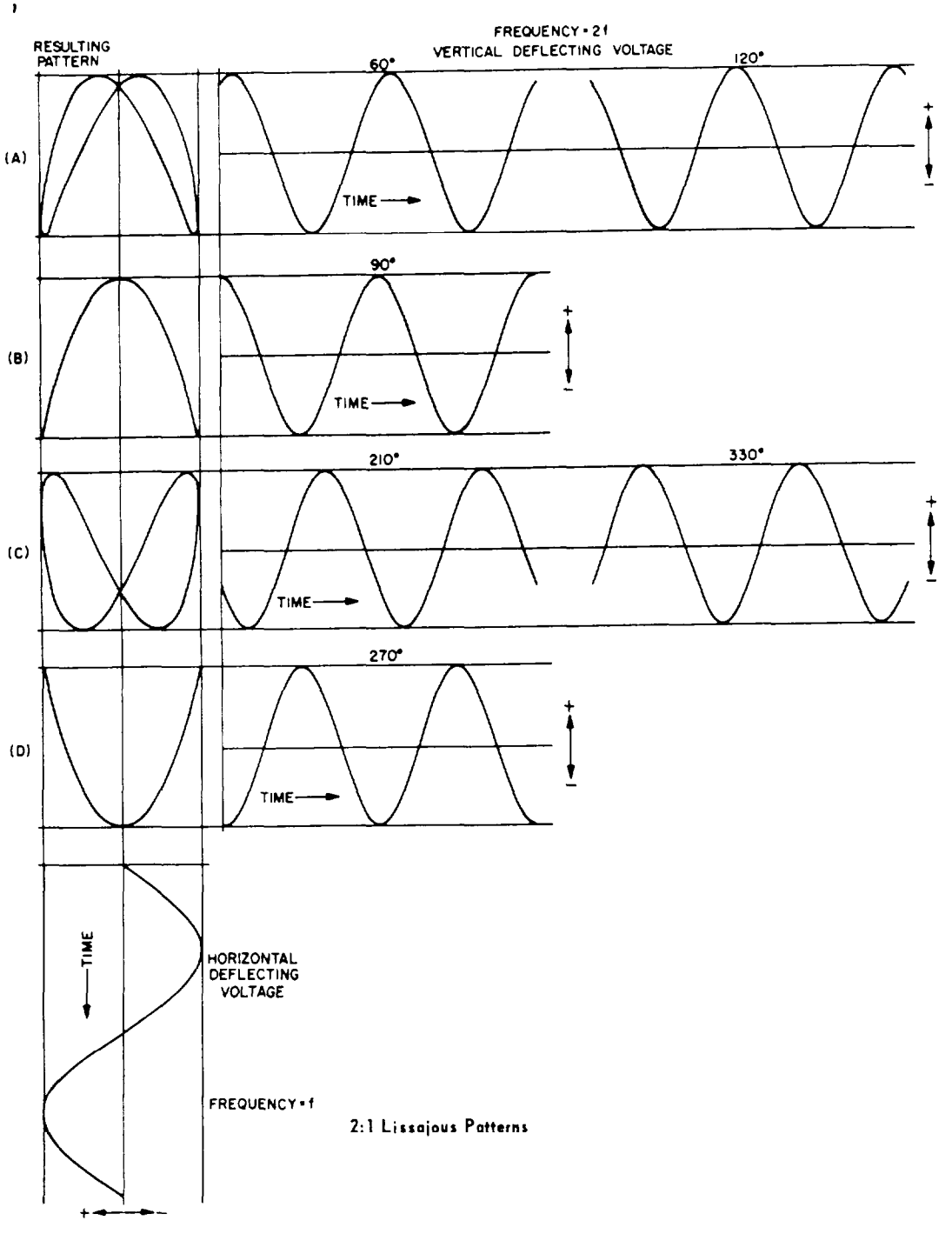


Figure 10-58.-2:1 Lissajous patterns for various phase relationships.

double-image pattern is also called an "uncompleted loop" or "closed pattern."

Each type of 2:1 Lissajous pattern, except the parabola, is developed for two phase relationships. For example, the pattern of figure 10-56 is also generated when the high-frequency signal is 180 degrees out of phase with the low-frequency signal. These alternative phases are shown in figures 10-56 and 10-58 by the high-frequency signals that produce the vertical deflection. When a double image such as the parabola is developed, a somewhat different method of evaluating the frequency ratio must be employed.

If the contact of an open line against the side of a pattern is counted as one-half, the correct ratio can be determined as shown in view C of figure 10-57. There is only one contact, against the vertical line, giving a figure of 1/2. There are two contacts against the top horizontal line, giving a total of 1. The ratio of the vertical deflecting-plate signal frequency to the horizontal deflection-plate signal frequency is therefore 1:1/2, or still 2:1. Now assume the horizontal line had been drawn against the bottom edge of the pattern. Here, the rounded end, or closed loop, of the parabola clearly has a single point of tangency with the line, giving a total of 1. The ratio is now, therefore, 1:1/2, which is still 2:1.

3:1 LISSAJOUS PATTERNS

Analogous conditions hold when the frequency ratio is 3:1. Representations of the various patterns that may be obtained are shown in view A of figure 10-59. If the signals applied to the deflecting plates are interchanged, the resulting patterns are exactly the same except for an axis rotation of 90 degrees. In the case of the S-shaped curve, the frequency ratio is computed by the same procedure as described for closed patterns in the previous paragraph. To illustrate, there would be a tangency and a contact with respect to a horizontal line drawn against the pattern shown in view B of figure 10-59. This gives a ratio of 1 1/2, or 3/2. If a vertical line were drawn, there would be a single contact, giving a figure of 1/2. The ratio of these two numbers is 3:1, which is consistent with the ratio of frequencies. Figures

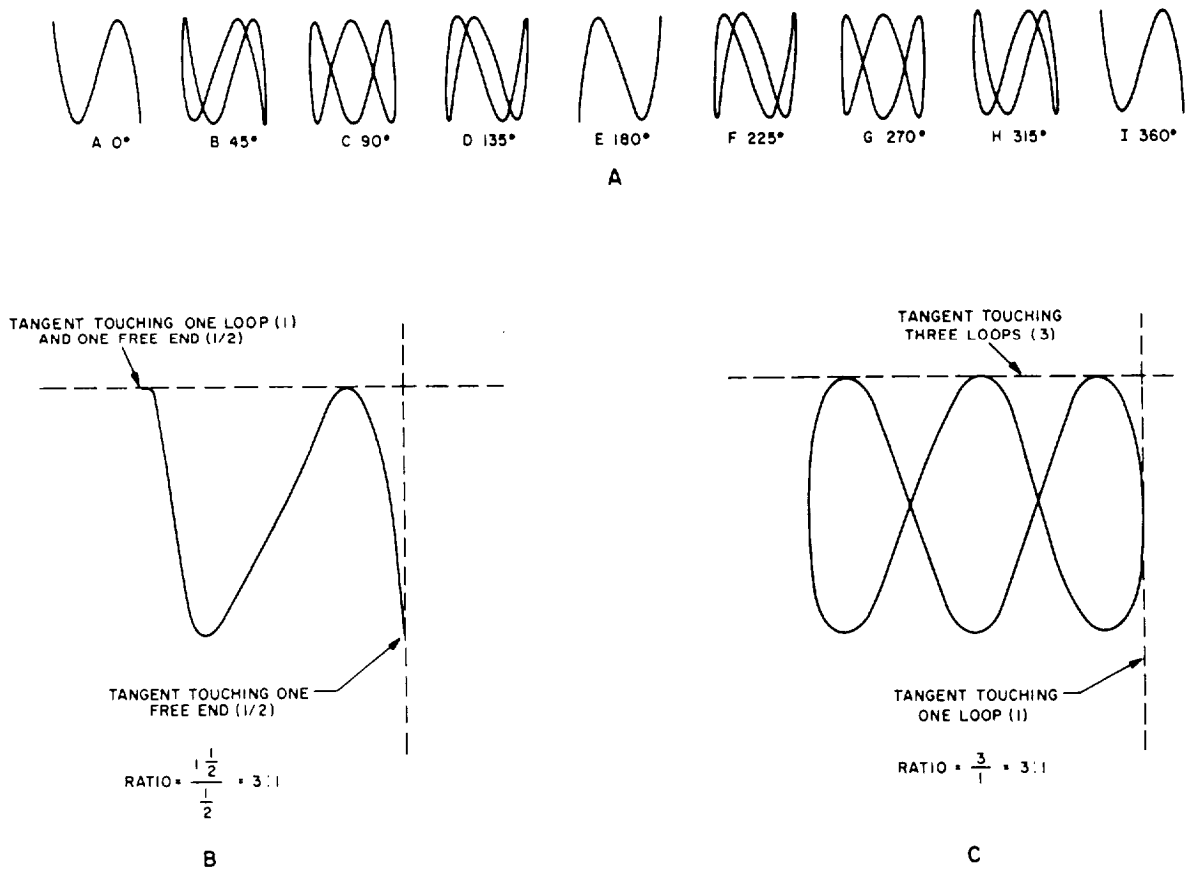


Figure 10-59.-3:1 Lissajous patterns and calculation of frequency ratio.

10-60 and 10-61 show how the phase relationship of the signals affects the resultant Lissajous pattern.

In figure 10-60, the high-frequency signal starts from its maximum amplitude just as a cycle of the low-frequency signal is ready to begin. This results in a symmetrical pattern comprising three loops. (This figure is also shown in view C of figure 10-59 for ratio computation purposes.) As shown in figure 10-59, view A, the same pattern is formed when the phase difference is 270 degrees instead of 90 degrees. Figure 10-61 shows how an S-shaped pattern is formed when the two signals are in phase. If the high-frequency signal began to swing negative as the low-frequency signal began, the pattern shown in view C of figure 10-59 would result.

OTHER LISSAJOUS PATTERNS

There are two restrictions on the frequencies of the signals applied to the deflecting plates. One was mentioned previously; namely, the frequency must lie within the useful pass band of the oscilloscope. Second, the relationship between the applied frequencies must not result in a pattern too involved for an accurate evaluation of the frequency ratio. As a rule, ratios as high as 10:1 and as low as 10:9 can be readily determined.

Patterns in which the ratios are 6:1, 8:1, and 10:1, including the corresponding double images, are shown in figures 10-62, 10-63, and 10-64. The discussion thus far has been limited to integral ratios, such as 1:1, 2:1, 8:1, and 10:1. In addition to these patterns, there are many patterns for which, even though the numerator and denominator of the ratio are whole numbers (or integers), the ratio is not an integer. For example, there are the 3:2 patterns of figure 10-65, the 5:4 patterns of figure 10-66, the 5:3 patterns of figure 10-67, the 7:2

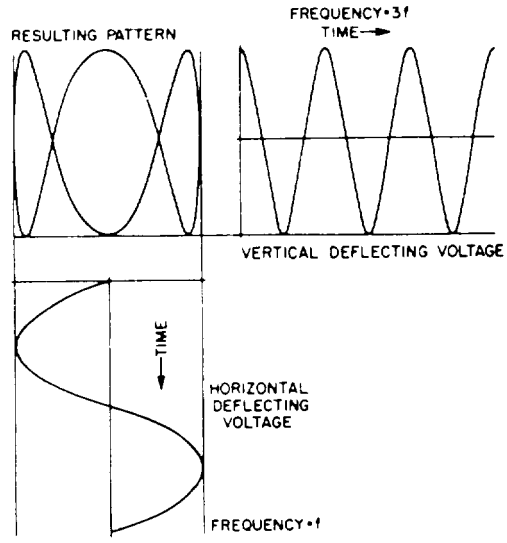


Figure 10-60.-OPEN 3:1 Lissajous pattern.

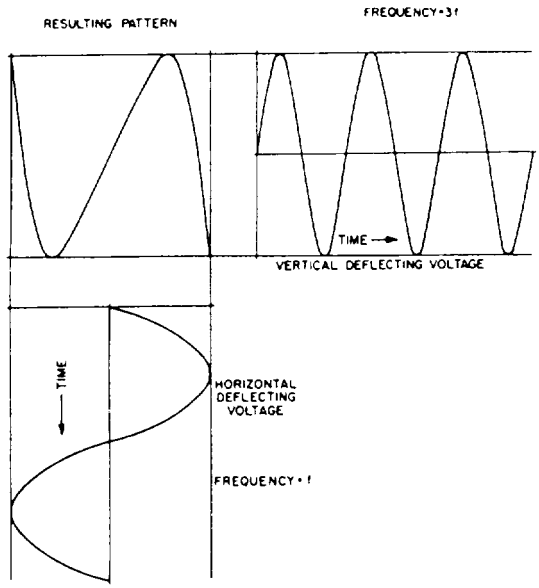


Figure 10-61.-CLOSED 3:1 Lissajous pattern.

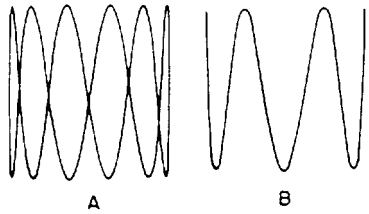


Figure 10-62.-6:1 Lissajous patterns.

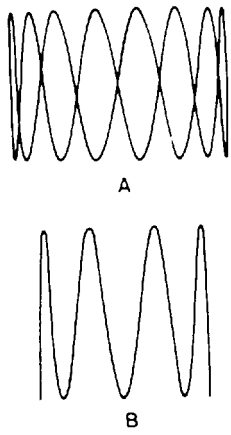


Figure 10-63.-8:1 Lissajous patterns.

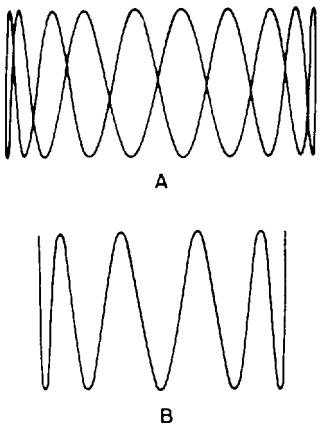


Figure 10-64.-10:1 Lissajous patterns.

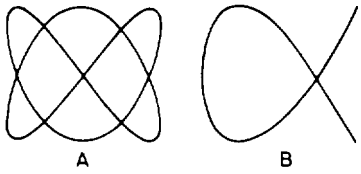


Figure 10-65.-3:2 Lissajous patterns.

patterns of figure 10-68, etc. In every case, however, the methods for determining the ratio of the applied frequencies are the same as those previously described.

TRANSIENT RESPONSE MEASUREMENT

Learning Objective: *Understand transient response measurements to include measurement technique, transients, and the measuring equipment.*

The ability of a linear device to cope with intermittent pulses is determined by the degree of transient response of the device. The quiescent state

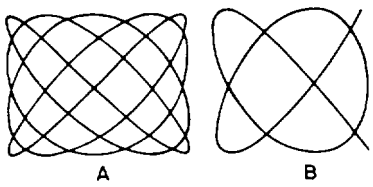


Figure 10-66.-5:4 Lissajous patterns.

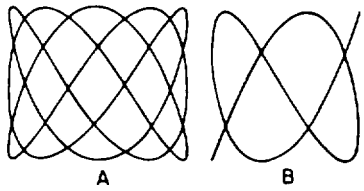


Figure 10-67.-5:3 Lissajous patterns.

must be considered and calculated as being zero while the technician determines the transient response of the device. (The calculation of transient response by this method will apply to linear devices only.) When a linear constant parameter device is excited by an input pulse, such as that shown in figure 10-69, the quiescent component of the device remains constant with respect to time (static-state current). Therefore, when making calculations, the total response shape can be considered as dependent only as the input pulse. The response of a linear constant-parameter device can be displayed and measured directly on an oscilloscope. The technician

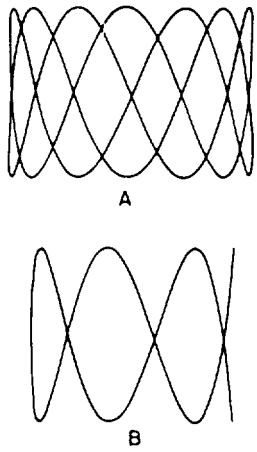


Figure 10-68.-7:2 Lissajous patterns.

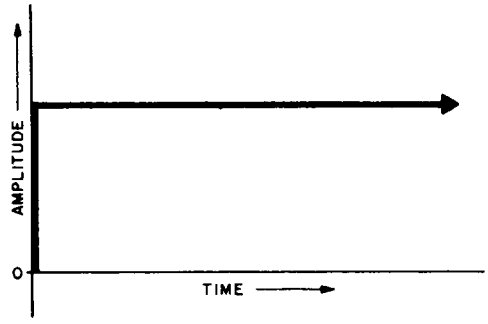


Figure 10-69.-Input step function.

must be aware that if this method is applied to a device that does not have constant parameters, it will be extremely difficult to separate the quiescent response from the transient response. The quiescent component cannot be considered as being "zero" in a nonconstant parameters device. The response to a single pulse input, as shown in view A of figure 10-70, is actually the time duration of the device's response to a single pulse input, as shown in view B of figure 10-70.

MEASUREMENT TECHNIQUE

The input signal to the amplifier or other electronic device should be a rectangular pulse with the stage output being applied to an oscilloscope to provide a visual display. The half-period of the input rectangular pulse must require a longer time than the transient response duration. The oscilloscope should be adjusted to show the width of a single complete pulse or less. The result of applying a proper pulse versus a too narrow pulse is shown in figure 10-71.

The transient response of a stage, with an input square wave, is measured by viewing two separate response characteristics. The first of these is related to the leading edge of the output response curve, and is composed of the rise time, time delay, and overshoot.

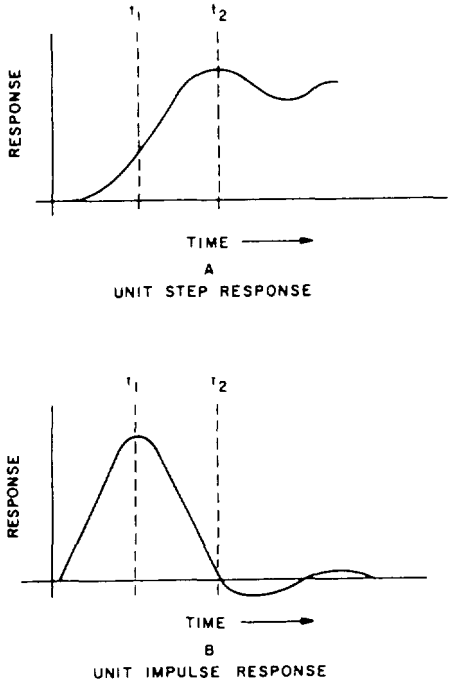


Figure 10-70.-Linear constant parameter amplifier responses.

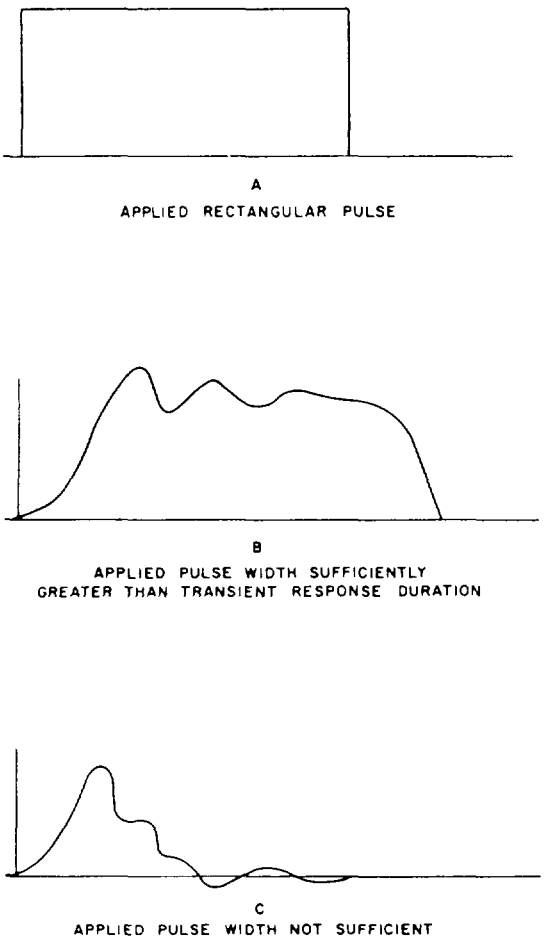


Figure 10-71.-Comparison of applied pulsewidth and transient response times.

The second of the response curve characteristics is related to the flat-top portion of the curve, called "sag." Sag is possible in a circuit only if the circuit is not capable of passing dc currents. In order to examine the leading edge of the wave, as shown in view B of figure 10-72, a fast sweep rate on the oscilloscope should be used, whereas a slow sweep rate should be used to illustrate the flat-top, as shown in view C of figure 10-72. Wide-input pulses normally have long rise-times. Therefore, narrow-input pulses with short rise-times are used to obtain response pulse leading-edge measurements. Wide pulses should be applied to the input for response pulse sag measurements. Figure 10-72 shows a pulse wide of 5 microseconds as being adequate for leading-edge measurements, but 1000-microsecond input pulses are required for the flat-top sag portions of the applied pulse. When a radio frequency signal, rather than a rectangular pulse, is used as the input to a tuned stage or band-pass device undergoing transient response measurements, the

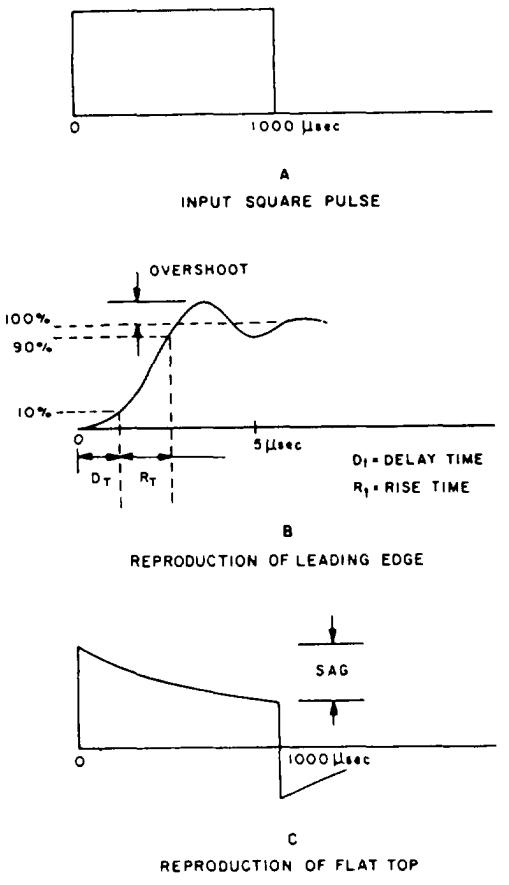


Figure 10-72.-Transient response characteristics.

parameters of the response are related only to the leading edge because there will be no flat top characteristics. The response of a band-pass device or tuned stage is shown in figure 10-73.

Transients

The total response in a linear circuit includes all of the individual transients due to the store of energy in each inductor, capacitor, and external energy source connected to the circuit, plus the steady state (forced response) of each external applied energy source. The response can be computed by starting at any arbitrary time (T = O) where all of the initial energy conditions of the proposed circuit are known. At high frequencies, the transit time contributes a conductance element to the grid input admittance. This occurs because as an electron passes the grid, it will introduce a grid current, even if it does not strike the grid. In this case, the grid conductance increases indirect proportion to the applied frequency.

You can visualize the transit time effect on an average tube if you realize that the input resistance to a typical electron tube may be several megohms at a frequency of 5 megahertz, while at 100 megahertz, the resistance drops to about 1500 ohms. In triode tubes, the transconductance of the tube decreases and lags behind an increasing amount as the transit time increases. The amplification factor also decreases and the phase angle increases. Much of the tube transit-time difficulty has been solved, however, by placing the electrodes closer together.

Reactive Elements

The discharge of a capacitive element through a resistor requires the time T_2 minus the original starting time (T_1) for the voltage or current to decay to 37 percent of the original value. The same analogy applies if the capacitive or inductive component charges to 63 percent. This type of analysis may be used for any periodic applied voltage. The steady-state current and voltage for an applied voltage are determined, the periodic voltage is resolved into its individual harmonic components, and then the transient is determined. The transient waveform does not bear a relationship to the applied voltage, as the transient waveform depends only on the circuit constants and the initial current and voltage conditions.

Resistive Elements

Time is considered to be zero when no reactive elements are present. Pure resistance elements do not charge or discharge with time.

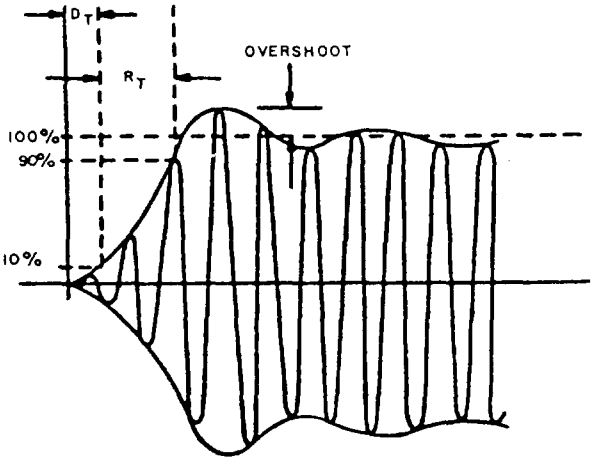


Figure 10-73.-Typical transient response of a tuned stage.

High frequency elements

At ultra-high frequencies, the transit time of an electron traveling between the cathode and the plate of a diode constitutes an appreciable part of the input cycle. This causes the dynamic plate resistance of the diode to decrease. Therefore, the cathode-to-plate resistance must be represented by a series resistor capacitor. Figure 10-74 shows the curves of both the resistance and the capacitance, each with respect to transit time, as the frequency increases. The resistance R/R_P eventually oscillates about a zero reference level, and is sometimes negative.

MEASURING EQUIPMENT

The device being tested contains a definite transit or rise time to be measured However, the equipment used to test the device may affect the rise time. In fact, if the rise times of the pulse generator and oscilloscope are each less than 10 percent of the rise time of the device being measured, an accuracy within less than 1 percent can be obtained. To partially compensate for the sag on the rectangular pulse, introduced by the testing equipment, the individual sags may be subtracted from the final measurement to obtain the correct value; that is, assuming that the sag introduced by the equipment is small. Figures 10-75 and 10-76 show the required test setup for the measurement of transient response within linear equipment.

The delay time measured with the aid of test setups, shown in figures 10-75 and 10-76, will result in a larger delay time than is actually contained in the equipment itself because of the additional time introduced by the

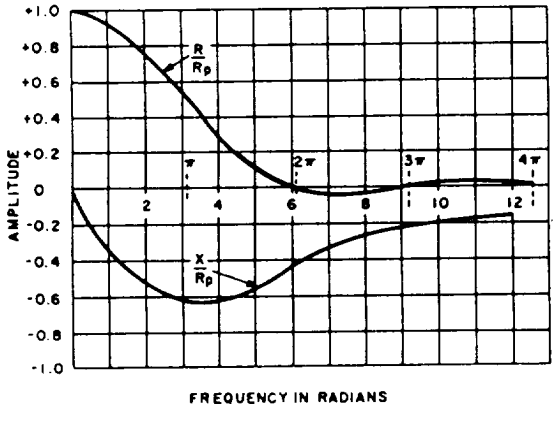


Figure 10-74.-Series resistive and reactive diode components represented as a function of frequency and transit times.

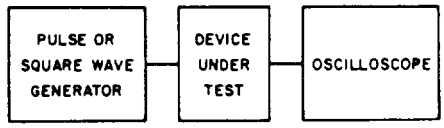


Figure 10-75.Typical test setup for measurement of transient response in low-pass equipment.

test equipment. However, tie test equipment time can be directly subtracted from the final measured result, and the measuring equipment will also tend to reduce the leading edge overshoot of the waveform in the device being tested.

The requirements of test equipment used to measure transient response are extremely rigid, as shown by several special features of the oscilloscope. A triggered oscilloscope, referred to as a "synchroscope," requires a variety of sweep speeds. The sweep circuit may be triggered by the trigger pulse that starts the pulse generator; the response being measured may be used to trigger the sweep so that later transient responses may be measured; or the applied signal maybe used to trigger the sweep circuits of the synchroscope, and then be passed through a delay line to the circuit under test. The pulse or square wave generator must contain a wide range of available pulsewidths and frequencies; the leading and trailing edges of the output pulse must be short as compared with the pulsewidth; and the sag should be flat and contain no ringing or oscillation. Finally, the carrier frequency pulses must be relatively free from frequency modulation during the active pulse.

TEST EQUIPMENT CONNECTION

The connection of test equipment to the device being tested for transient response is extremely critical because the internal impedance of the test equipment can load down the equipment under test, and thus cause

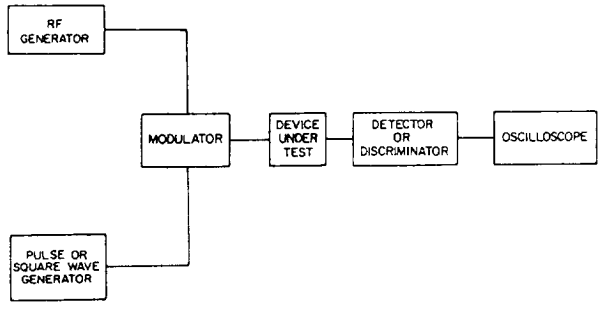


Figure 10-76.-Typical test setup for measurement of transient response In bandpass of FM equipment.

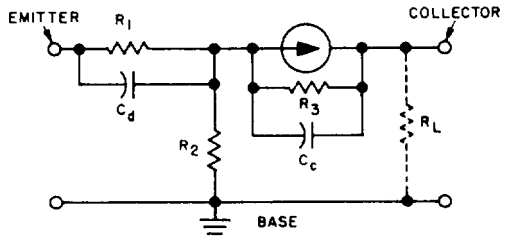


Figure 10-77.-Transistor equivalent switching circuit.

response distortion. A cathode-follower stage or other isolating device should be employed between the test equipment and the device under test to minimize the loading effect. All connecting leads should be maintained as short as possible, and the connecting lines must be matched to prevent feedback reflection along the line at high-frequencies, which would cause spurious ripples or ringing.

TRANSISTOR CONSIDERATIONS

The transient response of a transistor used as a switch is important because of the time required to turn the transistor switch from the "off" to the "on" position, and vice versa. Normally, either a step or a pulse of current applied to the input is required to turn the transistor from "off" to "on." Referring to figure 10-77, which is the high-frequency equivalent circuit, some assumptions and simple calculations can be made that will provide an approximate rise time within a transistor circuit operating in an active region.

First, assume that load resistor R_L is small enough to represent a short circuit; therefore, R_3 (the collector resistance) and C_c (the barrier capacitance) are effectively in parallel with R_2 (the barrier resistance), as shown in figure 10-77. Now, since R_2 is very much smaller than R_3 in this parallel circuit, the value of R_3 can be disregarded Furthermore, as C_c (the barrier capacitance) is much larger in reactance than R_2 at the assigned frequency, $2\pi F, C_c$ also can be neglected. Thus, for a grounded base configuration, the equivalent circuit of figure 10-77 can be reduced to the more practical equivalent circuit shown in figure 10-78. The value of C_d (the diffusion capacitance) is equal to the reciprocal of the assigned frequency, $2\pi F$, and the emitter resistance, R_1;

$$\frac{1}{2\pi F_1 R_1}.$$

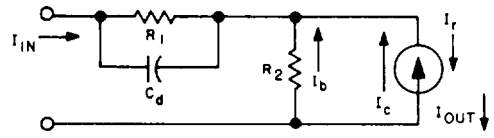


Figure 10-78.-Simplified transistor equivalent switching circuit with small output load.

Next, include a subsidiary circuit (R and C), as shown in figure 10-79. The time of RC equals the reciprocal of $2\pi F_1$. When the time T equals zero, output I_R equals zero, and when T equals infinity, output I_R equals the input (I_{in}). The time constant is the reciprocal of the

assigned frequency $\frac{1}{2\pi F_1 R_1}$, and the rise time is $\frac{1}{2\pi F_1}$.

This is the time required for the output current to rise from 0.1 to 0.9 percent of its final value.

The effect on the transient response of C_c (the collector capacitor) was not taken into account in the previous calculations because load resistor R_2 was considered small enough to be an effective short circuit. Figure 10-80 shows a simplified equivalent circuit, which recognizes that R_L does not affect the response when the input is from an infinite series-impedance current and $R_L C_c$ is very much larger than the time-constant

$$\frac{1}{2\pi F_1}$$

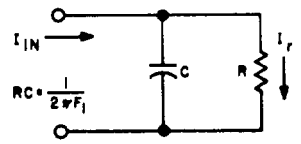


Figure 10-79.-RC circuit, simulating frequency dependence, used to calculate I_R.

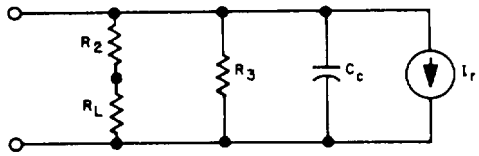


Figure 10-80.-Simplified equivalent transistor switch circuit with large output load.

In fact, if $\frac{R_2}{R_3}$ is very much smaller than 1 and $2\pi F_1 R_3 C_c 2$ *is* much larger than 1, then the time-constant, calculated for the response with a very small or shorted value of R_L, would be increased by the amount, $2\pi F_1 R_L C_c + 1$. Therefore, the rise times and turn-on times would be increased by this same factor.

TRANSISTOR DELAY TIME

In the previous calculations of transistor response, the finite time required for the impulse signal to diffuse across the base region was neglected. Under this condition, if a current pulse is applied to the transistor emitter, a response will appear at the collector only after some delay in time. The value of I_R may be obtained for use in the equivalent circuits shown in figures 10-78 and 10-80 by representing this delay time with an equivalent circuit, as indicated in figure 10-81. The line attenuation will increase in direct proportion to the frequency. The resistive-capacitive delay line indicates a time interval before a response is indicated at the output.

TRANSISTOR STORAGE TIME

Up to this point, the discussion has concerned the turn-on transistor process. If the transistor is in the active region, the turn-off process consists simply of applying a pulse of reverse polarity, and the required time-constant is calculated in the same manner, using the equation for the turn-on process. Unfortunately, if the transistor is in the "on" condition and is operating in the saturation region, an abnormally large time-delay will occur before the transistor responds to the turn-off signal. This peculiar delay, termed *storage-time delay,* is shown in figure 10-82, which shows the minority carrier density in the base region for three situations.

The first situation shown is the cutoff condition, with both the emitter and collector junctions back-biased. The minority carrier density is therefore zero at both junctions, and very small throughout the base region.

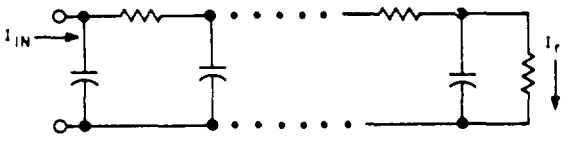


Figure 10-81.-Transistor switch equivalent delay time circuit.

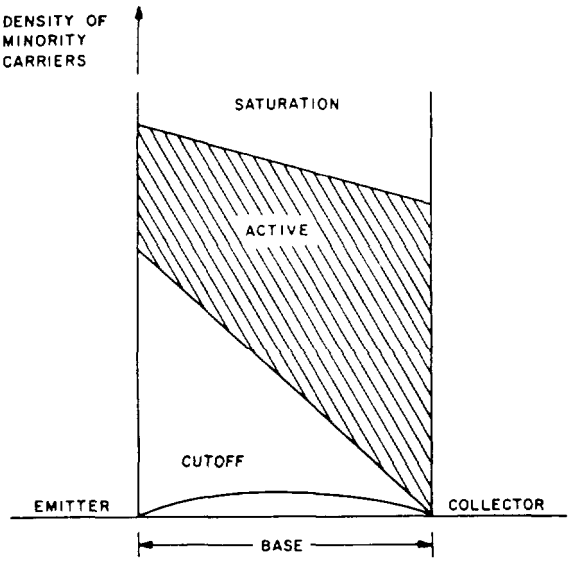


Figure 10-82.-Comparative minority-carrier densities in transistor base for cutoff, active, and saturation regions of operation.

The second situation is shown by the active curve in figure 10-82, where the minority carrier density is high at the emitter junction and zero at the collector junction. The change in density between the two junctions is the result of the diffusion process, which accounts for current flow across the base of the transistor. However, if an input signal drives the emitter junction to a back-bias condition, the diffusion process will not continue until the minority carriers in the base region have been removed.

The third situation is shown by the saturation curve in figure 10-82. Unlike the cutoff and active conditions, the previously discussed equivalent circuits used to determine time delays, response, etc., do not apply in this case because both the emitter and the collector are emitting carriers into the base region. In addition, as both junctions are forward-biased, the junction voltages will be small and the collection process at the junctions will be slow. This, in turn, causes the density of minority carriers in the base region to build up to a relatively large value. This high-density level in the base region must be permitted to decrease before the turn-off process begins to take effect. This long-storage time delay may represent two or three times the normal rise or fall time in the active region. Therefore, it is evident that when a transistor switch is used in an application requiring high switching speeds, it must be restrained from entering the saturation region.

TRANSISTOR RESPONSE

Figure 10-83 shows the approximate waveforms required to represent the response from a transistor driven from cutoff to saturation and back again. A grounded-emitter switch is used because this type of configuration is the most useful and its response can represent other configurations.

In figure 10-83, the delay time is represented by the symbol T_D, the rise time by T_0, the storage time by T_S, and the decay or fall time by T_1. The input current reverses at the end of the pulse rather than falling only to zero. As a result, the output current response falls toward a negative value rather than toward zero. The fall time of the pulse is thereby reduced. The voltage input waveform of figure 10-83 was terminated in a minus E_2

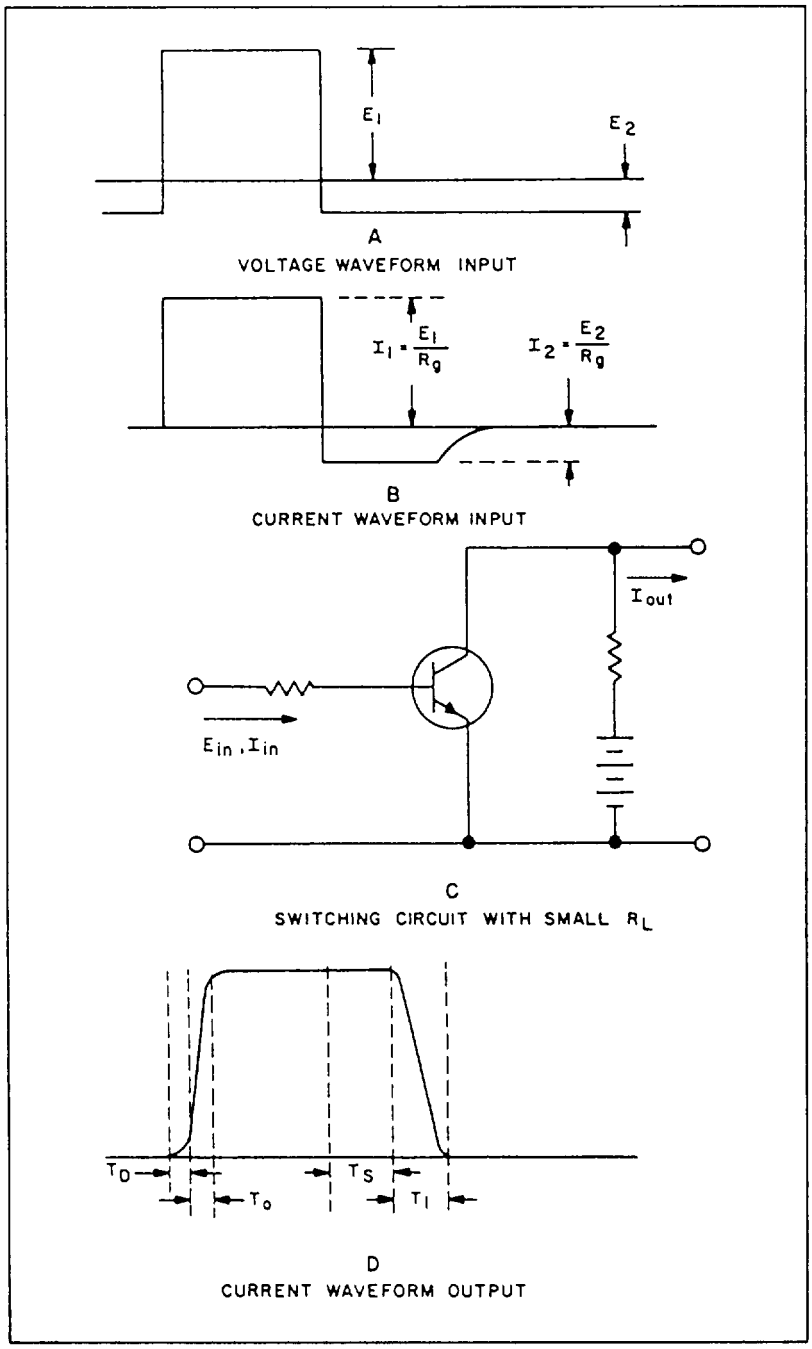


Figure 10-83.-Grounded-emitter switch circuit with input voltage, current waves, and output current response waveform for small R_L.

voltage because the storage of minority carriers in the base region does not permit the transistor input impedance to immediately attain a large value. In fact, the input impedance remains small until the minority carriers at the transistor junctions are swept away. At this point the input impedance increases and causes the input current to decrease in direct proportion to the speed with which the minority carriers throughout the base region drift to the junctions.

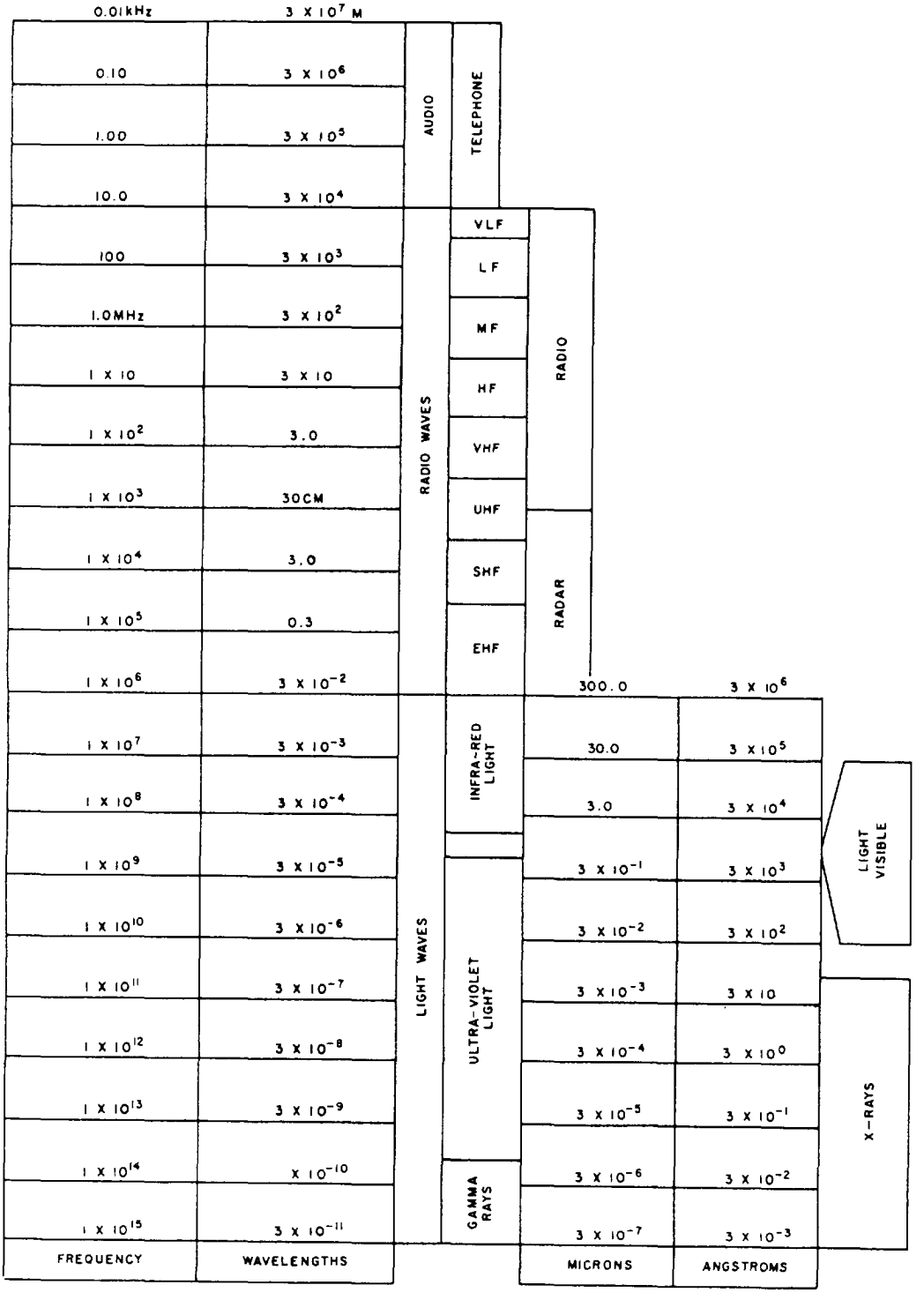


Figure 10-84.-Electromagnetic frequency spectrum.

SPECTRUM WAVEFORM ANALYSIS AND MEASUREMENTS

Learning Objective: *Understand the function and usage of a spectrum analyzer to include analyzing the spectrum pattern as well as the operation of the analyzer.*

The analysis of a complex waveform, prepared in terms of a graphical plot of the amplitude versus frequency, is known as spectrum analysis. Spectrum analysis recognizes the fact that any waveform is composed of the summation of a group of sinusoidal waves, each of an exact frequency and all existing together simultaneously.

ELECTROMAGNETIC FREQUENCY SPECTRUM

A chart showing the electromagnetic frequency spectrum is given in figure 10-84. This chart indicates the frequency and wavelength of the various frequency bands.

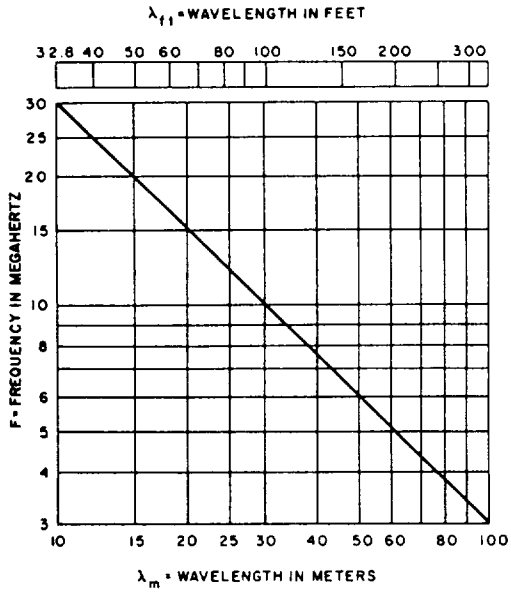


Figure 10-85.-Wavelength-frequency conversion chart.

WAVELENGTH FREQUENCY CONVERSION

The chart in figure 10-85 and table 10-1, when used in conjunction with each other, provide a ready means of converting frequency to wavelength and vice-versa.

Table 10-1.-Auxiliary Wavelength-Frequency Conversion Table

FOR FREQUENCIES IN MEGAHERTZ FROM	MULTIPLY f BY	MULTIPLY λ
0.03 - 0.3	0.01	100.000
0.30 - 3.0	0.10	10.000
3.00 - 30.0	1.00	1.000
30.00 - 300.0	10.00	0.100
300.0 - 3000.0	100.00	0.010
3000.00 - 30000.0	1000.00	0.001
The formula conversions are:	λ m - 300/f MHz	
	λ cm - 30/f GHz	
	λ FT - 984/f MHz	
	λ IN - 11.8/f GHz	

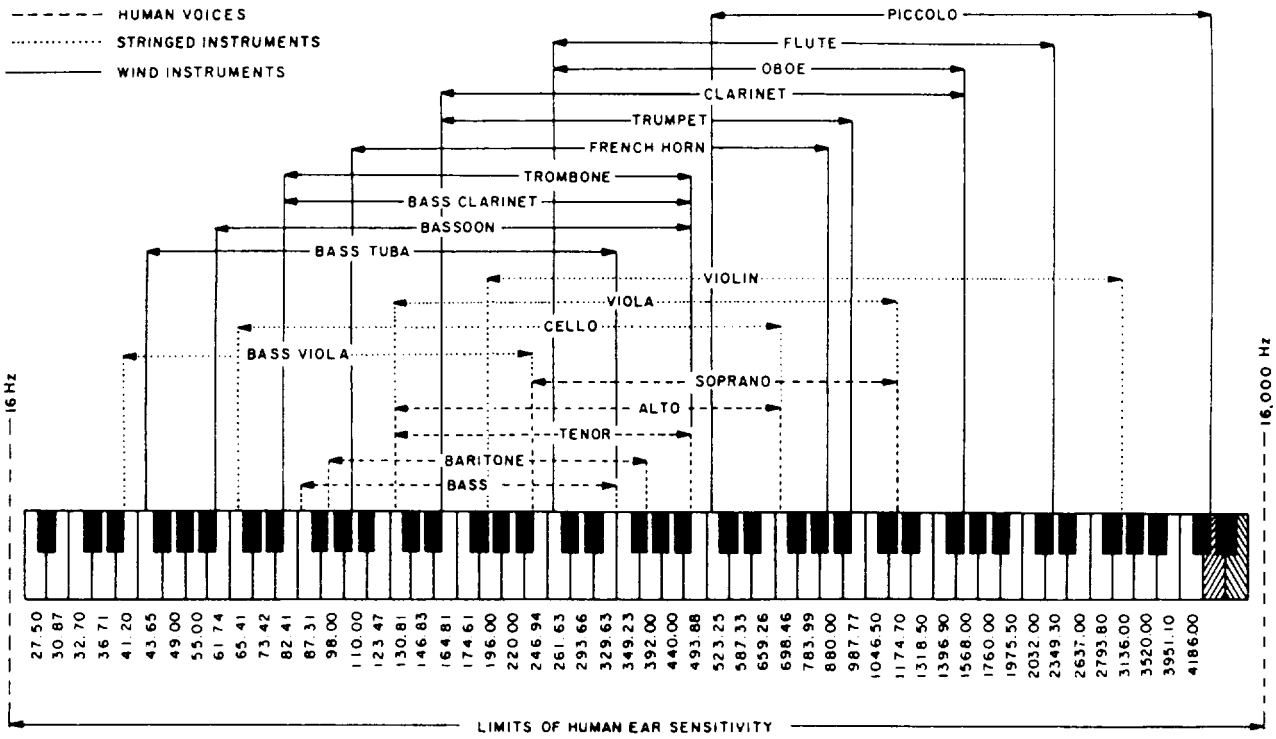


Figure 10-86.-Acoustic spectrum.

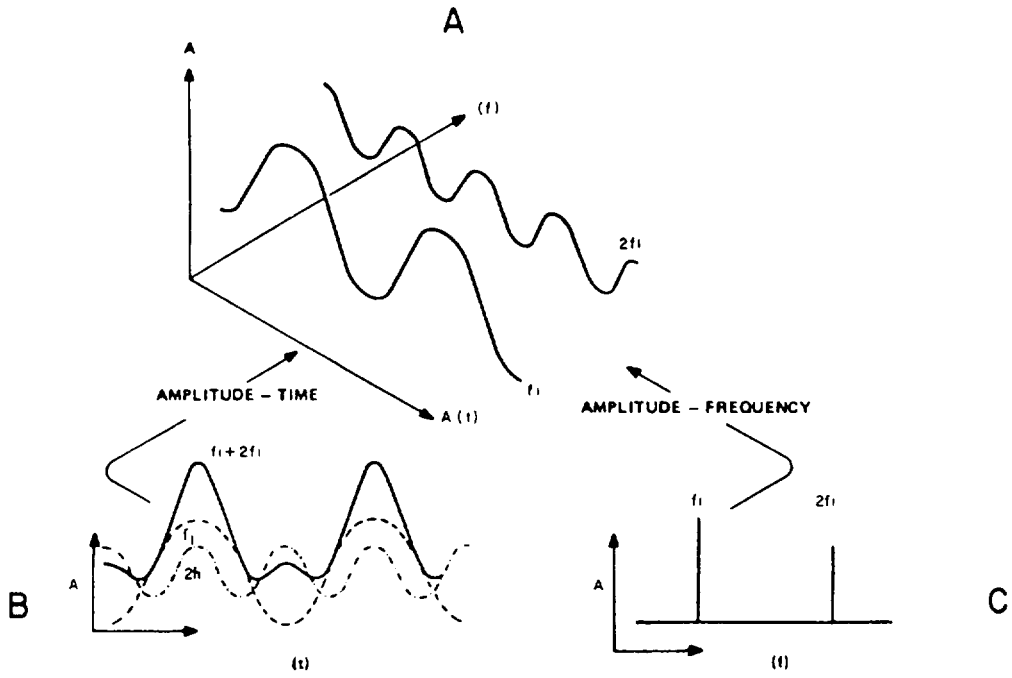


Figure 10-87.-Time vs frequencies.

For example, a frequency of 25 MHz has a wavelength of 12.5 meters, and a frequency of 250 MHz has a wavelength of 1.25 meters. Both frequencies enter the chart at the 25 level of f, but the λ multiple in table 10-1 for 25 MHz is 1.0 while it is .10 for 250 MHz.

ACOUSTIC SPECTRUM

The acoustic spectrum chart shown in figure 10-86 is provided to show the limits of the sound spectrum as set by human ear sensitivity. This chart is based on a musical pitch of 440, a physical pitch of 426.667, and an international pitch of 435.

SPECTRUM ANALYSIS

In the realm of varying frequency, three axes of degree exist: amplitude, time, and frequency. The time domain (amplitude vs time) plot is used to recover phase relationships and basic timing of the signal, and is normally observed with the aid of an oscilloscope. The frequency domain (amplitude vs frequency) plot is used to observe frequency response, employing the spectrum analyzer for this purpose. Figure 10-87 shows the difference between frequency and time domain plots. View A illustrates a three dimensional coordinate of a fundamental frequency and its second harmonic with respect to time frequency and amplitude. View B depicts the time domain display as it would be seen on an oscilloscope. The solid line, $f_1 + 2f_1$, is the actual display. The dashed lines, f, and $2f_1$, are drawn to illustrate the fundamental and second harmonic frequency relationship used to formulate the composite signal $f_1 + 2f_1$. View C is the frequency domain display of view B as it would be seen on a spectrum analyzer. Note in view C that the components of the composite signal are clearly seen.

FREQUENCY DOMAIN DISPLAY CAPABILITIES

The frequency domain contains information not found in the time domain. The spectrum analyzer can display signals composed of more than one frequency (complex signals); and it can discriminate between its components, while measuring the power level at each one. It is more sensitive to low-level distortion than an oscilloscope, and its sensitivity and wide dynamic range are also useful for measuring low-level modulation, as shown in views(a) and (b) of figure 10-88. The spectrum analyzer is useful in the measurement of long- and short-term stability such as noise sidebands on an oscillator, residual FM of a signal generator, or frequency drift of a device during warm-up. This is

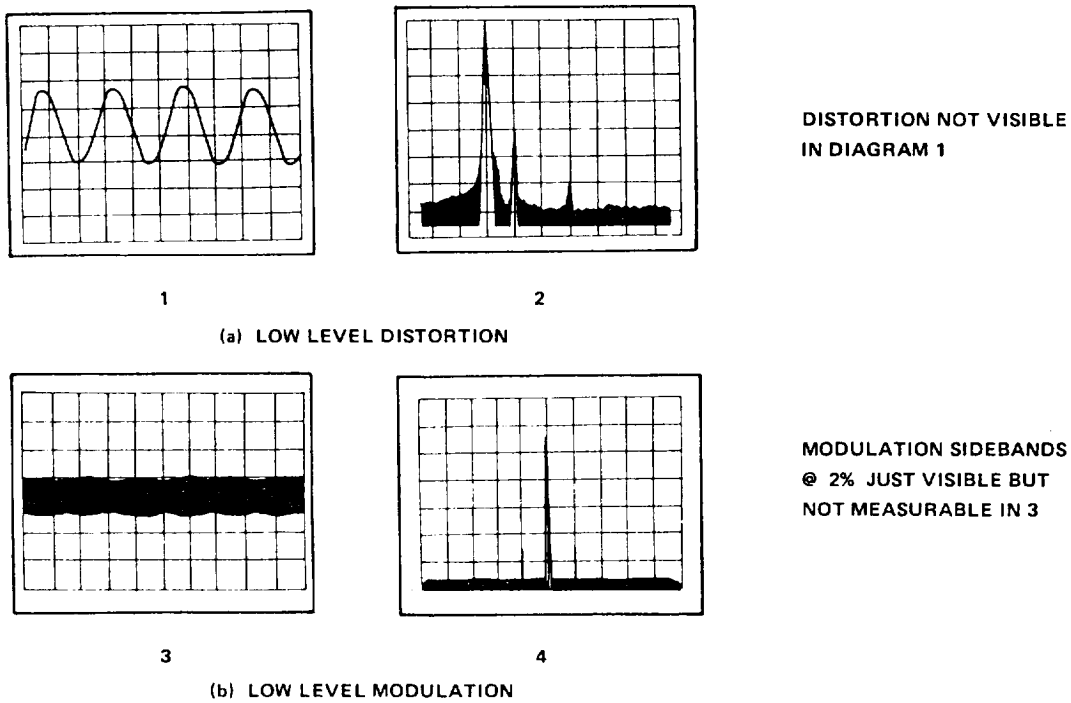


Figure 10-88.-Examples of time domain (left) and frequency domain (right) low-level signals.

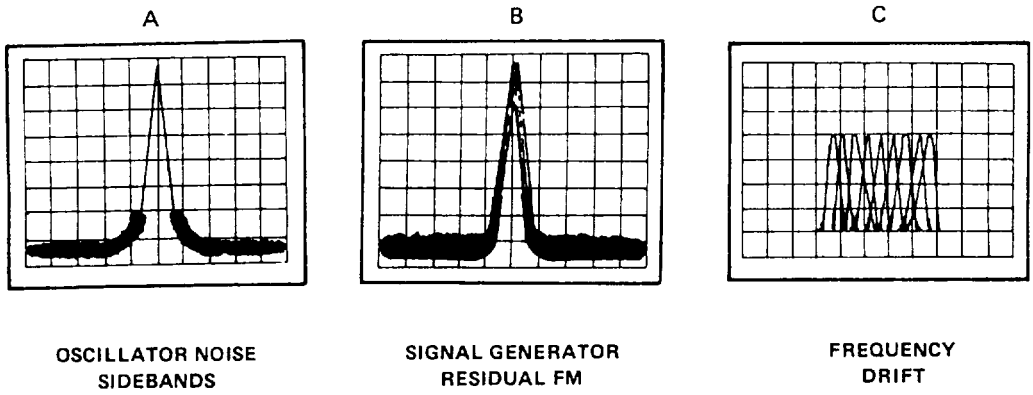


Figure 10-89.-Spectrum analyzer stability measurements.

shown in figure 10-89. The swept frequency response of a filter or amplifier and the swept distortion measurement of a tuned oscillator are also measurable with the aid of a spectrum analyzer. However, in the course of these measurements, a variable persistence display on an X-Y recorder should be employed for simplification of readability. Examples of tuned oscillator harmonics and falter response are shown in figure 10-90. Frequency conversion devices such as mixers, harmonic generators, etc., are easily characterized by such parameters as conversion loss, isolation, and distortion. These parameters can be displayed, as shown in figure 10-91, with the aid of a spectrum analyzer. Present-day spectrum analyzers can measure segments of the frequency spectra from Ø Hz to as high as 40 GHz.

SPECTRUM ANALYZER USAGES

Although the previously mentioned measurement capabilities are attainable with a spectrum analyzer, you will find that, in general, the spectrum analyzer is used

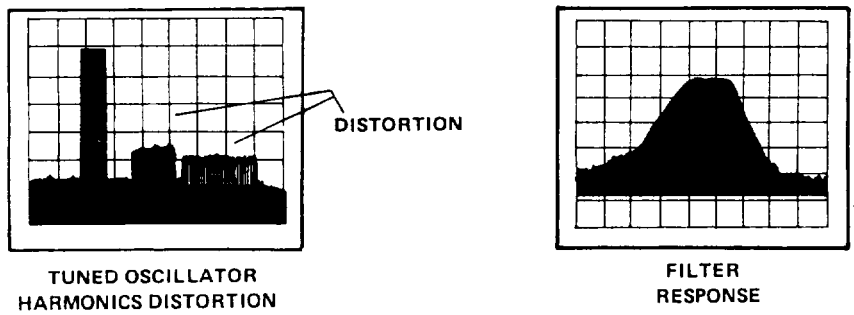


Figure 10-90.-Swept distortion and response characteristics.

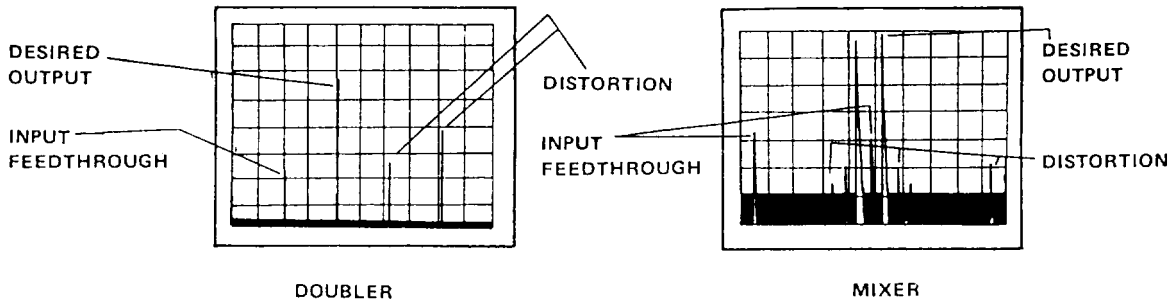


Figure 10-91.-Frequency conversion characteristics.

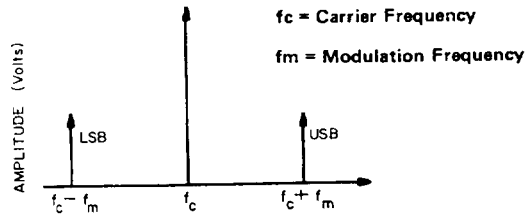


Figure 10-92.-Spectrum analyzer display of an AM signal.

to measure spectral purity of multiplex signals, percentage of modulation of AM signals, modulation characteristics of FM and PM signals, and in the interpretation of the displayed spectra of pulsed RF emitted from a radar transmitter. Occasionally, field strength measurements are required to determine RFI (i.e., radio frequency interference).

COMPLEX WAVEFORMS

Complex waveforms are divided into two groups–periodic waves, and nonperiodic waves. Periodic waves contain the fundamental frequency and its related harmonics. Nonperiodic waves contain a continuous band of frequencies, resulting from the repetition period of the fundamental frequency approaching infinity, and thereby creating a continuous frequency spectrum.

MODULATION MEASUREMENTS

In all types of modulation, the carrier is varied in proportion to the instantaneous variations of the modulating waveform. The two basic properties of the carrier available for modulation are the amplitude characteristics and the angular (frequency or phase) characteristic.

Amplitude Modulation

The modulation energy in an amplitude-modulated wave is contained entirely within the sidebands. Amplitude modulation of a sinusoidal carrier by another sinusoid would be displayed as shown in figure 10-92. For 100-percent modulation, total sideband power would be one-half of the carrier power; therefore, each sideband will be 6 dB less than the carrier, or one-fourth of the power of the carrier. Since the carrier component is not changed with AM transmission, the total power in the 100-percent modulated wave is 50-percent higher than in the unmodulated carrier.

The primary advantage of the log display of the spectrum analyzer over the linear display provided by the oscilloscopes for percentage of modulation measurements is that the high dynamic range of the spectrum analyzer (up to 70 dB) allows accurate measurements of values as low as 0.06-percent. It also allows the measurement of low-level distortion of AM signals. Both capabilities are shown in figure 10-93.

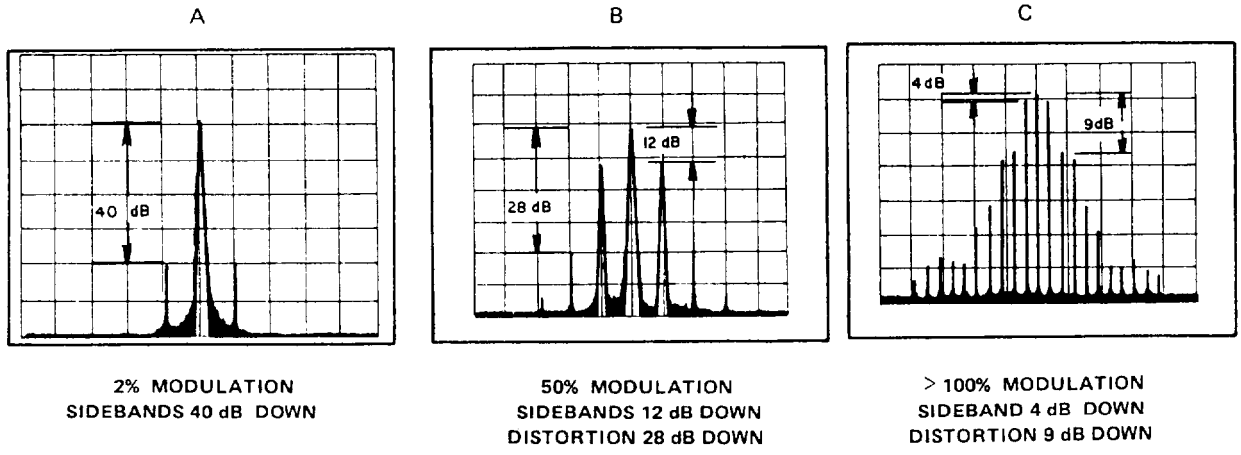


Figure 10-93.-AM display of spectrum analyzer.

The chart in figure 10-94 provides an easy conversion of dB down from carrier into percent of modulation. Anything greater than -6 dB is over 100-percent modulation; therefore, producing distortion as shown in view C of figure 10-93.

In modem long-range HF communication, the most important form of amplitude modulation is SSB (single sideband). Either the upper or lower sideband is transmitted and the carrier is suppressed. Single sideband requires only one-sixth of the output power required by AM to transmit an equal amount of intelligence power and less than half the bandwidth. Figure 10-95 shows the effects of balancing out the carrier of an AM signal. The most common distortion experience in SSB is intermodulation distortion, which is caused by nonlinear mixing of intelligence signals. The two-tone test is used to determine if any intermodulation distortion exists. Figure 10-96 shows the spectrum analyzer display of the two-tone test.

Frequency Modulation

Amplitude modulation contains the intelligence in the sideband current pairs spaced symmetrically about the carrier by an amount equal to each modulation frequency. Theoretically, frequency modulation can contain an infinite number of sideband current pairs per modulating frequency with the intelligence spread throughout them as well as along the carrier.

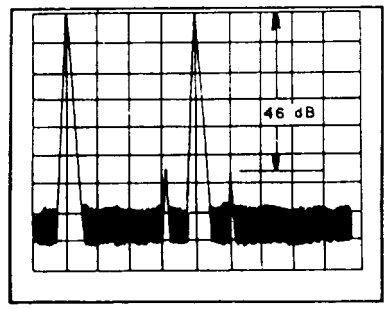


Figure 10-95.-Double sideband carrier suppressed.

The amplitude of a particular pair of side currents may be larger than the center frequency component. This fact also holds true for phase modulation, in that, with the same modulation index (ß), the same spectrum distribution is obtained. The number of important side currents is larger for low frequencies in the signal band than it is for high frequencies.

Table 10-2 provides selected amplitude factors used to multiply the maximum unmodulated carrier current level (I_m) to find the amplitude value of an individual side current pair in the frequency spectrum. For example, use a maximum center frequency swing (ΔF) = ±60 kHz and a 30 kHz signal frequency (f) to find the value of ß.

$$\beta = \frac{\Delta F}{f} = \frac{60}{30} = 2.$$

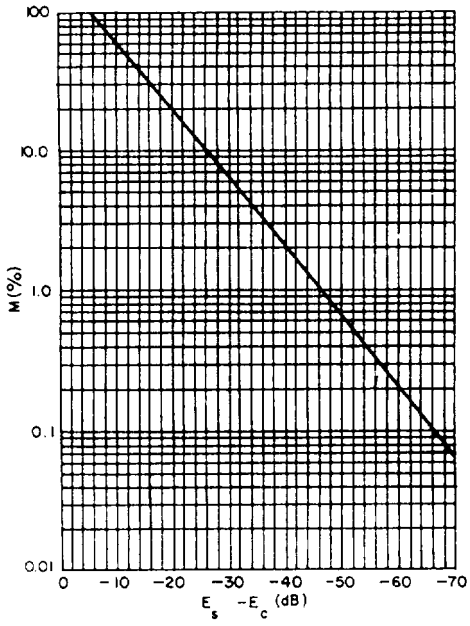


Figure 10-94.-Modulation percentage M vs sideband level (log display).

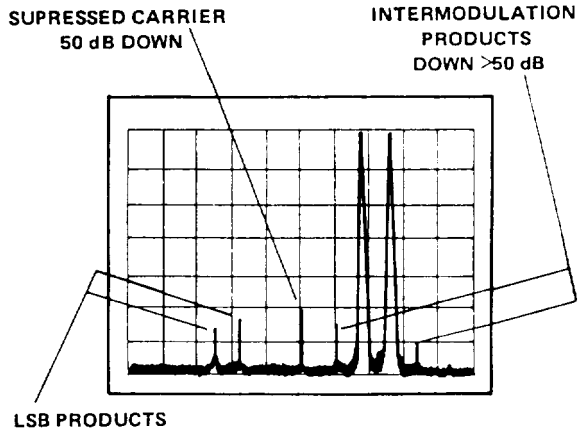


Figure 10-96.-Two-tone test.

Table 10-2.-Abbreviated Bessel Factor Table

β	$J_0(\beta)$	$J_1(\beta)$	$J_2(\beta)$	$J_3(\beta)$	$J_4(\beta)$	$J_{14}(\beta)$
0.0	1.0000	0.0000	0.0000	0.0000	0.0000	
0.5	0.9385	0.2423	0.0306			
1.0	0.7652	0.4401	0.1449	0.0196	0.0025	
2.0	0.2239	0.5767	0.3528	0.1289	0.0340	
5.0	–0.1776	–0.3276	0.0466	0.3648	0.3912	
10.0	–1.2459	0.0435	0.2546	0.0584	–0.2196	0.0120

Table 10-2 shows that $J_0(ß) = J_0(2) = 0.2239$, and this is the value multiplied times the maximum unmodulated current (I_m) to obtain the magnitude of the center frequency component (F). The value of $J_1(ß) = J_1(2) = 0.5767$ multiplied times I_m gives the amplitude of both the first upper side current at the frequency of (F + 30 kHz) and the first lower side current at the frequency of (F – 30 kHz).

The spectrum distribution for a modulation index of ß = 2 is shown in figure 10-97. This graph was prepared from data obtained from table 10-2. As all amplitudes are obtained by multiplying the Bessel factors obtained from table 10-2 times I_m, the magnitude of the Bessel factors directly determines the intensity of the sideband current pairs in the useful frequency spectrum. The sideband current pairs that are too far down in amplitude from the center frequency (F) are not significant; they are less than 1 percent of the unmodulated carrier (I_m). Thus, the bandwidth is determined by the number of significant sideband current pairs.

The bandwidth may also be calculated from table 10-2 for a specific frequency swing. For example, let ß = 10 and read the Bessel function values across the chart, from left to right. As you can see, the 14th sideband pair is 0.012 I_m, which is significant. Therefore, the maximum bandwidth for ß = 10 is 2 (both side currents) times 14 sidebands times the signal frequency deviation. For a 30-kHz signal deviation, the bandwidth would be 2 x 14 (30)= 840 kHz, and for 2 kHz it would be 56 kHz. This example shows why a higher-modulation signal requires more frequency

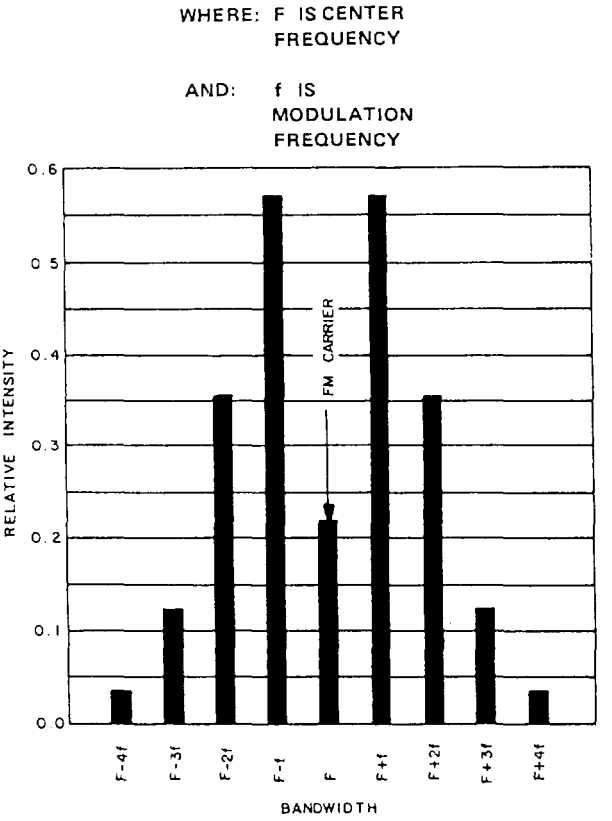
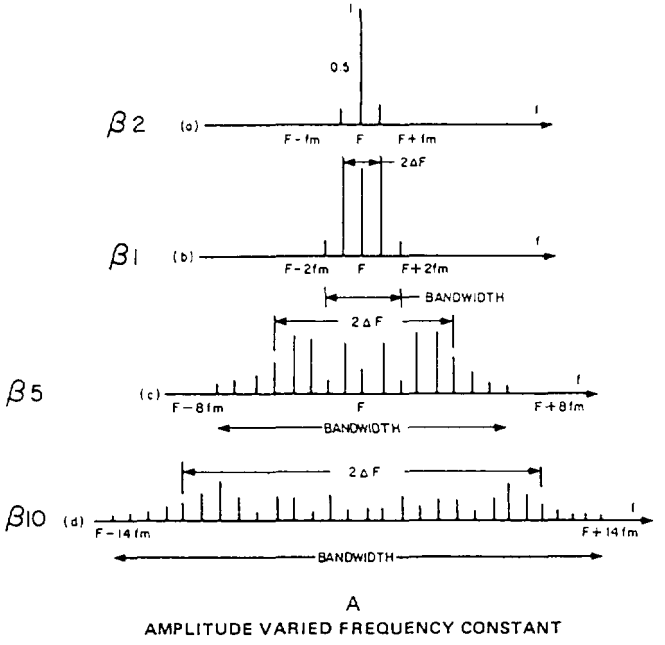


Figure 10-97.-Spectrum distribution for a modulation index of ß = 2.

space (that is, greater bandwidth) than does a lower modulation signal.

Figure 10-98 shows the effect of changing the amplitude or frequency of the modulating signal while holding the other constant. The Bessel factors given in table 10-2 and figure 10-99 also show that the greater the modulation index (ß), the greater the number of significant sidebands. As the modulation energy spreads out from the carrier frequency (F), it can be determined at what point to establish the bandwidth. For example, in figure 10-99, which was prepared for an index of ß = 24, the largest side currents occurred at the edges of the pass band; thereafter, continue to read off the Bessel factors until the side current pairs are less than 1 percent of the unmodulated carrier current level, I_m. Sideband currents beyond this point do not have significant amplitudes for practical consideration.



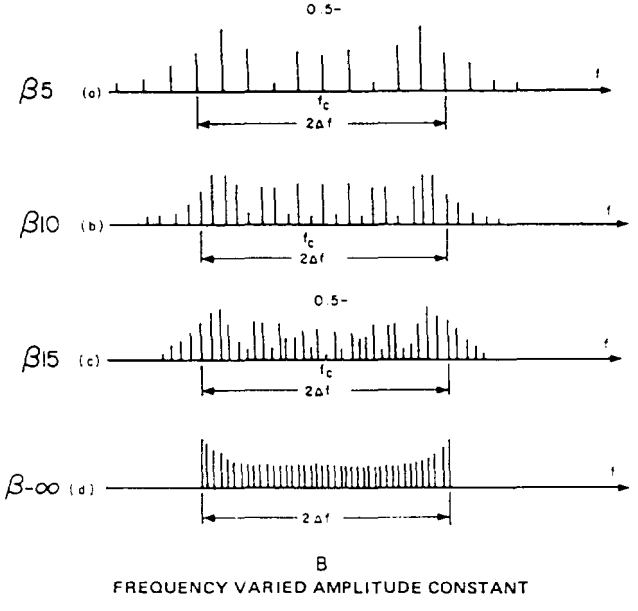


Figure 10-98.-Modulation frequency and amplitude effects on an FM carrier.

Frequency modulation, for modulation index values smaller than 0.2, is similar to amplitude modulation in that both types of modulation contain only one significant side-current pair. Therefore, for a value of

$$\frac{\Delta F}{f} = \beta = 0.2$$

or less, FM behaves exactly like AM with respect to spectrum distribution. However, unlike AM, no primary oscillator can be used in the FM transmitter because the carrier frequency must be varied during the modulation cycle to produce FM. The desired intelligence in FM or phase modulation (PM) will create more energy distribution, and thus a larger response in a receiver demodulator than will noise energy. This is the outstanding desirable feature of FM over AM spectrum distribution.

Phase Modulation

In PM, unlike FM or AM, the carrier current level (I_m), as well as the center carrier frequency (F), remains constant. Only the relative phase (θ) changes. The actual value of θ is not important. The deviation from this value is important, and it produces the desired PM. For example, if θ equals 30 degrees and the phase deviation is ±40 degrees, this will produce a certain modulation of the carrier. However, if θ had originally been 80 degrees rather than 30 degrees and was subjected to a ±40 degree deviation, the same output PM waveform, containing the same intelligence, would have been produced. The original value of θ indicates only the original amplitude at the start of the phase swing. Actually, the effect of a ±40-degree deviation would be to give the appearance of wobbling about the carrier frequency (F) as F goes through an angular distance of ±360 degrees, regardless of its beginning phase, θ. However, the equivalent instantaneous frequency of the modulated earner would remain the same. Thus, PM

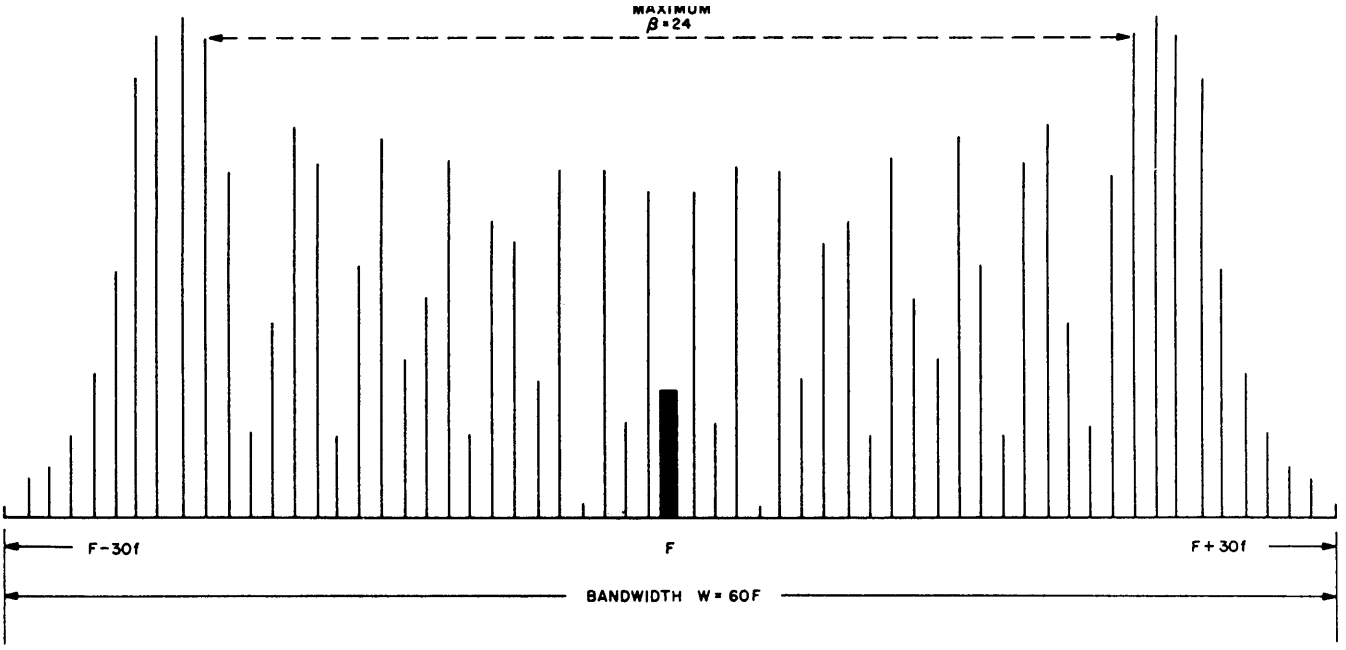


Figure 10-99.—Spectrum distribution for an index of 24.

entails the important fact that it is created not only by the maximum phase deviation ($\Delta\beta$), but also by the applied signal frequency (f). Therefore, the frequency shift is greater for higher frequencies than it is for lower frequencies.

Sidebands

The sidebands for PM are similar to the sidebands for FM, and the same general formula for the modulation index holds true. That is

$$\beta = (\frac{\Delta\theta}{f}).$$

As with FM, the side currents contain a symmetrical frequency distribution around the carrier. In other words, the first upper sideband and the frost lower sideband have the same numerical value, amplitude, and frequency difference from the carrier; the value of the modulation index (ß) is proportioned to the phase deviation ($\Delta\theta$); and for a fixed maximum phase swing, it does not matter whether you use a 15 hertz or 15 kHz signal to modulate the phase of the carrier. In either case, you will obtain the same number of important side-current pairs. However, for the 10th upper and lower sideband in the 15-hertz modulation case, the side-current pair is 10 x 15 hertz or 150 hertz above and below the carrier. For the 15 kHz case, the 10th side-current pair is 10 x 15 kHz or 150 kHz above and below the carrier, and thus requires a much broader bandwidth. Therefore, PM and FM, unlike AM, have a spectrum distribution of the modulation energy, which is proportional to the square of the spectrum amplitudes, and does affect the carrier frequency amplitude. However, it does not matter whether the technician is dealing with a modulation index of 10 for FM or maximum phase deviation of 10 radians (573 degrees) for PM. This is true for PM as long as the maximum phase swing ($\Delta\theta$), which causes the PM, is fixed. The same spectrum distribution is equally true for FM if the maximum deviation (Δ F), which causes the FM, changes directly with the signal frequency (f).

As the bandwidth of PM remains the same regardless of signal frequency changes, the PM bandwidth can be calculated directly from the modulation index (ß). For small values of modulation index (0.3 or less), PM will contain only one pair of significant sidebands, the same as for FM. Actually, this permits you to amplitude-modulate a carrier, suppress the carrier, and shift the modulation product by 90 degrees to provide narrow-band FM or PM. The FM or PM effect can be obtained by applying a signal voltage having a magnitude that is inversely proportional to the signal frequency (f). The phase-shifted product is then combined with the unmodulated carrier.

CARRIER FREQUENCY

Both FM and PM contain a true instantaneous and an equivalent instantaneous carrier frequency.

Therefore, the number of hertz must remain constant to prevent the center or mean frequency (F) from drifting to some other nearby frequency not originally assigned to the carrier. If this occurred, the entire spectrum centered around the carrier would drift and infringe on the other nearby FM channels.

PULSED WAVES

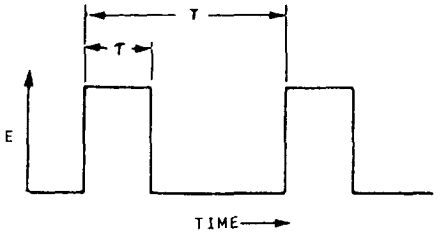
An ideal pulsed radar signal is comprised of a train of RF pulses with a constant repetition rate, constant pulsewidth and shape, and constant amplitude. To receive the energy reflected from a target, the radar receiver requires close to ideal pulse radar emission characteristics. By observing the spectra of a pulsed radar signal, such characteristics as pulsewidth, duty cycle, peak and average power can be measured easily and accurately.

RECTANGULAR PULSE

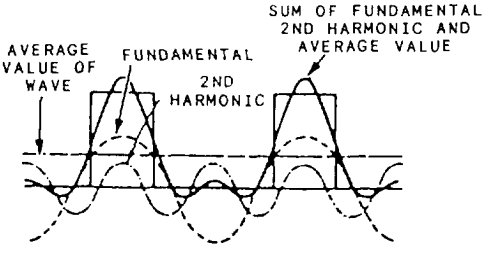
A rectangular wave is used to pulse-modulate the constant frequency RF carrier to produce the pulse radar output. The rectangular wave is comprised of a fundamental frequency and its combined odd and even harmonics. Both in and out of phase harmonics relationships are utilized, depending on the desired pulsewidth or pulse interval. Figure 10-100 shows the development of a rectangular wave and its spectral content.

PULSED WAVE ANALYSIS

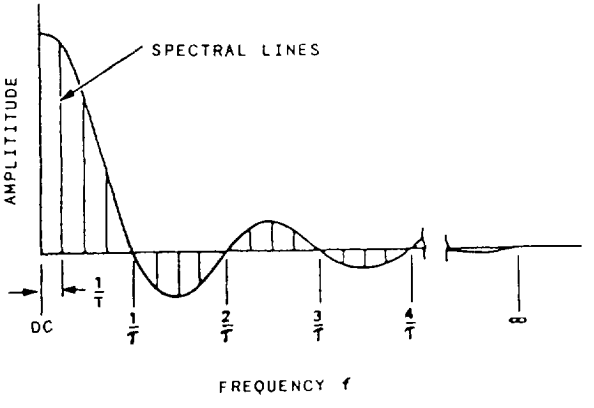
In AM, the sidebands are produced above and below the carrier frequency. The principle also applies for a pulse, except that the pulse is comprised of many tones. These tones produce multiple sidebands, which are commonly referred to as spectral lines or "rails" on the analyzer display. There will be twice as many rails in the pulse radar's modulated output as there are harmonics



RECTANGULAR PULSE
TIME DOMAIN



RECTANGULAR PULSE
HARMONIC DEVELOPMENT



RECTANGULAR PULSE
SPECTRA (FREQUENCY DOMAIN)

Figure 10-100.-Rectangular pulse.

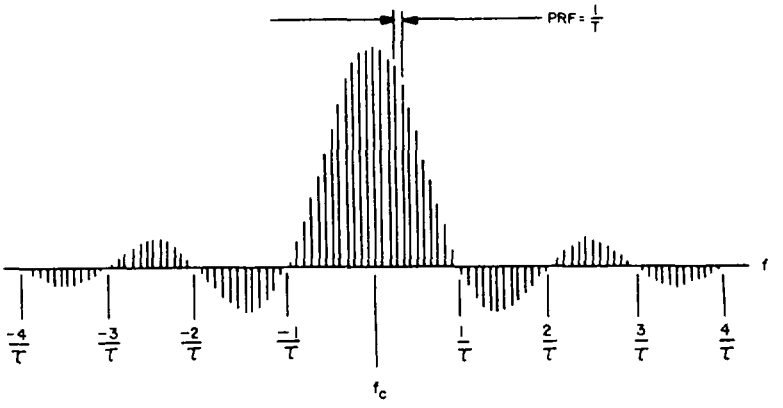


Figure 10-101.-Pulsed radar output.

contained in the modulating pulse (upper and lower sidebands), as shown in figure 10101.

In figure 10-101, the PRF is equal to the pulse interval of 1/T. The actual spectrum analyzer display would show the lower lobes (shown below the reference line) on top because the spectrum analyzer does not retain any phase information. Changing the pulse interval or pulsewidth of the modulation signal will change the amount of rails (PRF) or number of lobe minima, as shown in figure 10-102.

ANALYZING THE SPECTRUM PATTERN

The leading and trailing edges of the radiated pulse-modulated signal must be extremely steep, with a

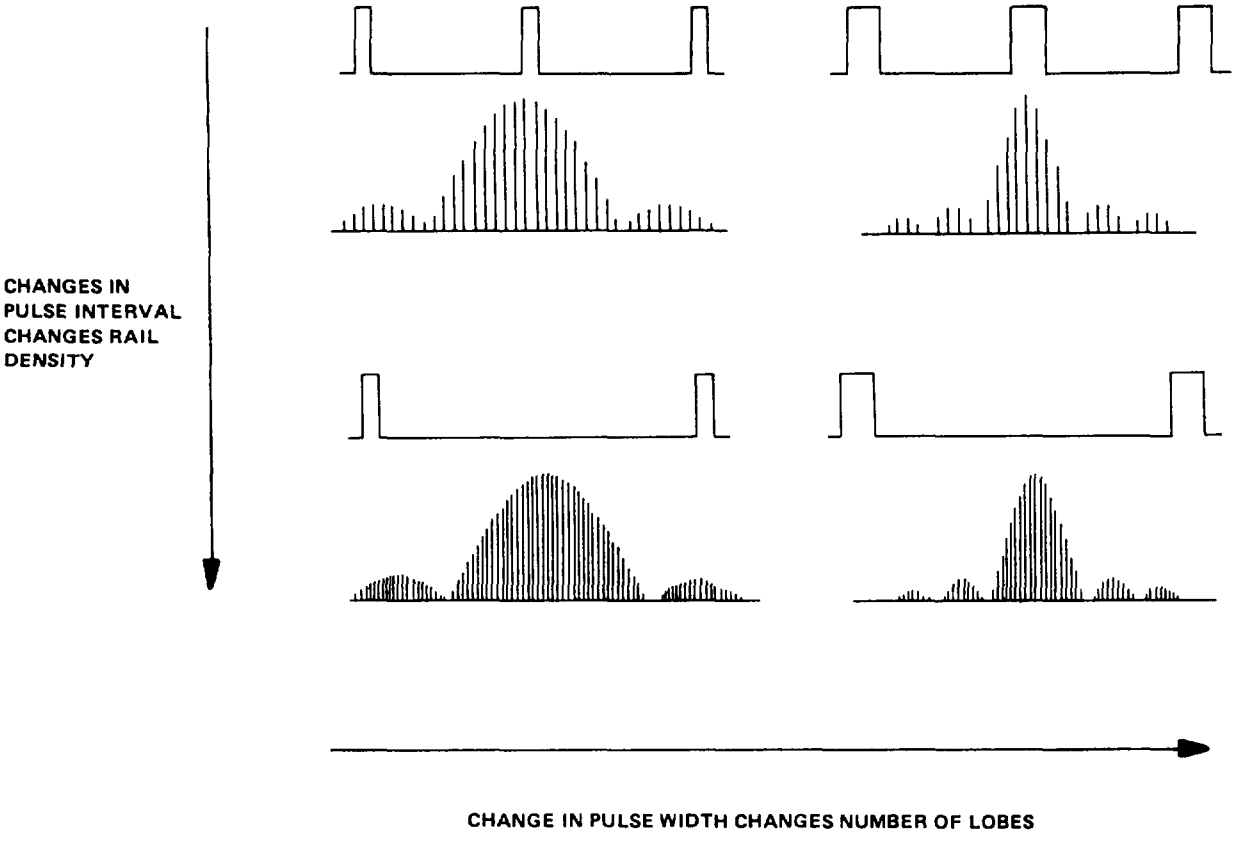


Figure 10-102.-Pulsed radar changes affected by modulating signal changes.

constant amplitude between them. Incorrect pulse shape will cause frequency spread and pulling, which results in less available energy at the frequency to which the receiver is tuned.

The primary reason for analyzing the spectrum is to determine the exact amount of amplitude and frequency modulation present. The amount of amplitude modulation determines the increase in the number of sidebands within the applied pulse spectrum, whereas an increase in frequency modulation increases the amplitude of the side lobe frequencies. In either case, the energy available to the main spectrum lobe is decreased.

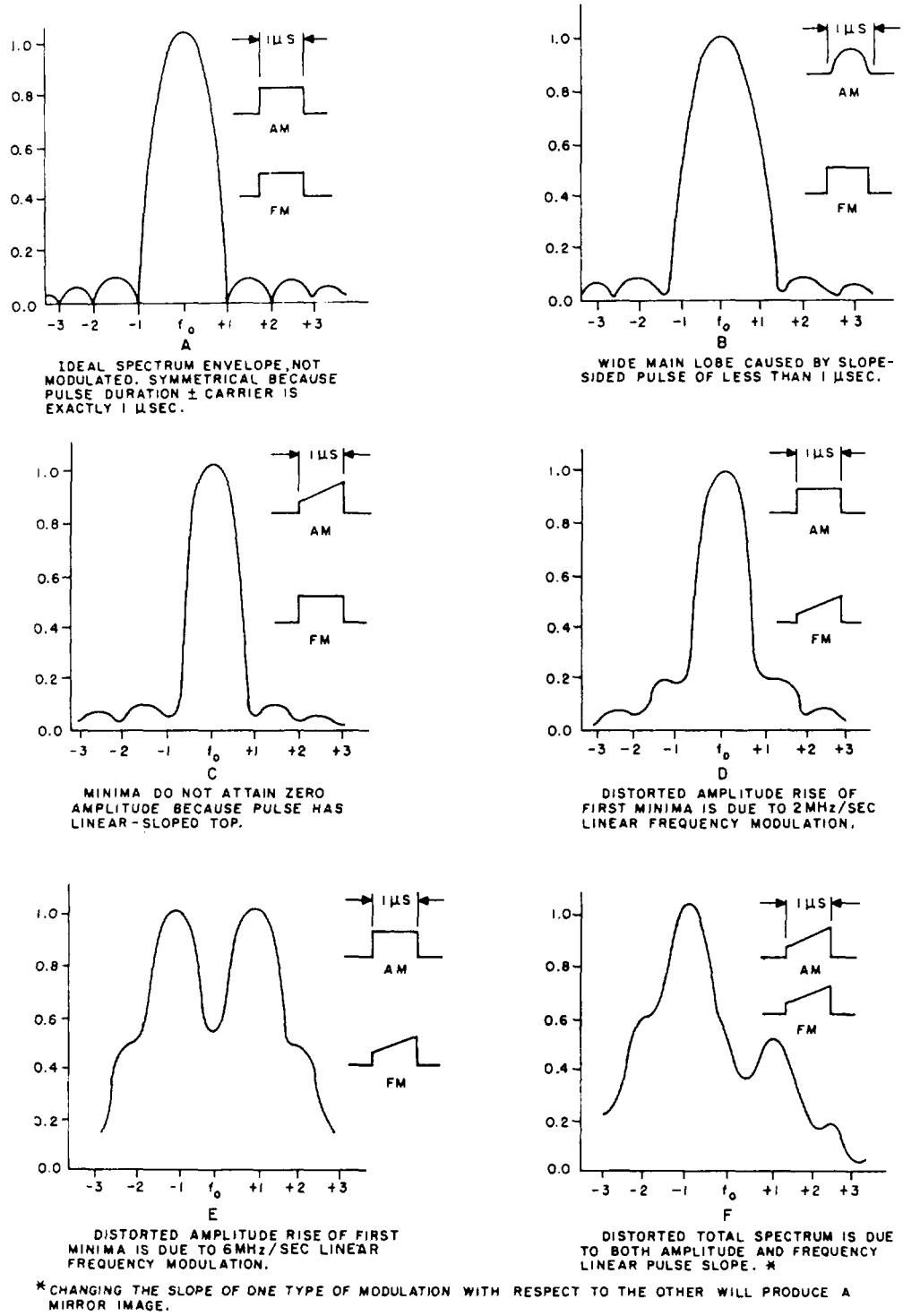


Figure 10-103.-Spectrum patterns.

Typical Spectrum Patterns

Figure 10-103 contains several illustrations of commonly obtained patterns. These spectra are the result of pulse-modulated waves, which are special types of RF carrier amplitude modulation. As can be seen, amplitude modulation alone does not seriously affect the frequency spectrum on an RF pulse. The type of modulation present can be easily determined because amplitude modulation primarily affects the amplitude of the side lobes and does not affect the shape of the main lobe. Frequency modulation affects the main lobe bandwidth. Spectrum asymmetry, as shown in view F of figure 10-103, occurs only when both amplitude and frequency modulation occur simultaneously.

Spectrum Analyzer Interpretations

A pulsed RF signal has unique properties; therefore, you must be careful to correctly interpret the display on a spectrum analyzer. Spectrum analyzer response to a pulsed radar signal can be of two kinds, resulting in displays that seem similar but are of completely different significance. One response is called a "line spectrum," and the other is called a "pulse spectrum." Both are responses to the same pulsed radar signal, and the *line* and *pulse* spectrum terms refer solely to the response of the display on the spectrum analyzer.

LINE SPECTRUM.– A line spectrum occurs when the spectrum analyzer's 3 dB bandwidth (ß) is narrow compared to the frequency spacing of the input signal components, as shown in figure 10-104.

PULSE SPECTRUM.– A pulse spectrum occurs when the spectrum analyzer's bandwidth (ß) is equal to or greater than PRF. The spectrum analyzer, in this case, cannot resolve the actual individual frequency components because several rails are within its bandwidth. However, if the spectrum analyzer's bandwidth (ß) is narrow, as compared to the spectrum envelope, the envelope can then be resolved, as shown in figure 10-105. Figure 10-106 shows the effect of varying the scan width bandwidth of the spectrum analyzer in the line analysis interpretation. Figure 10-107 shows the effect of the same variations in the pulse analysis interpretation.

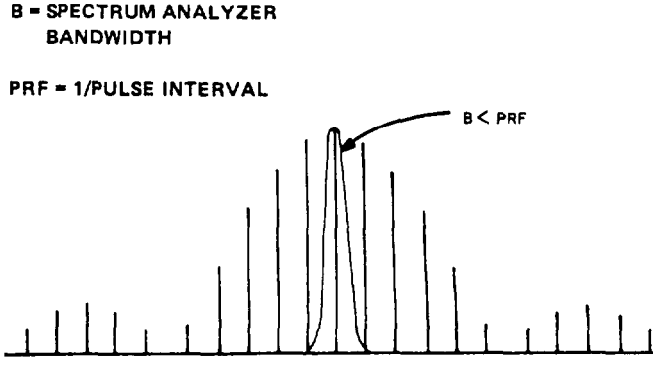


Figure 10-104.-Line spectrum B<PRF.

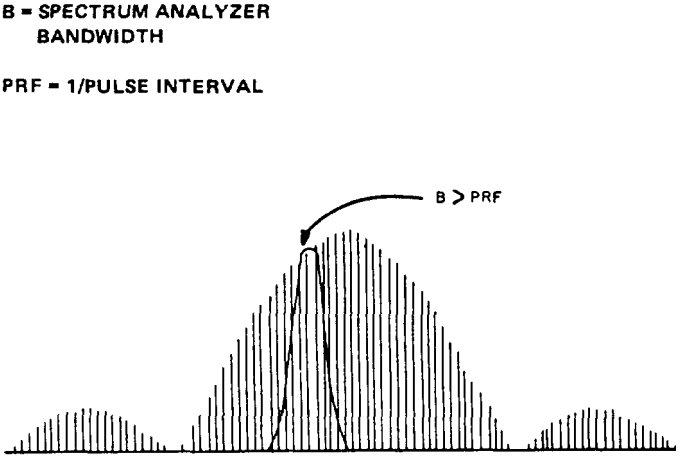
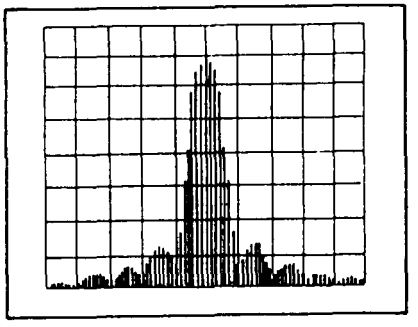


Figure 10-105.-Pulse spectrum B>PRF.

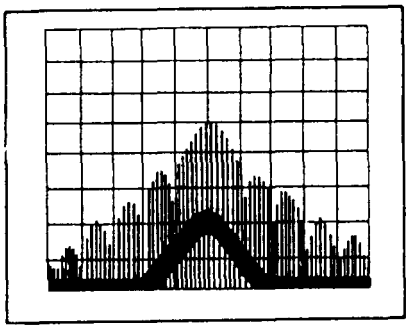
SPECTRUM ANALYZER OPERATION

The information desired from the spectra to be analyzed determines the spectrum analyzer requirements. Real-time analysis is used if a particular point in the frequency spectrum is to be analyzed, such as line spectra displays. Continuous or swept frequency analysis, which is the most common mode of observation, is used to display a wider portion of the frequency spectrum or (in some cases) the entire range of the spectrum analyzer in use. Changing the spectrum analyzer setting from one mode to another is accomplished by varying the scan time and/or the spectrum analyzer's bandwidth.

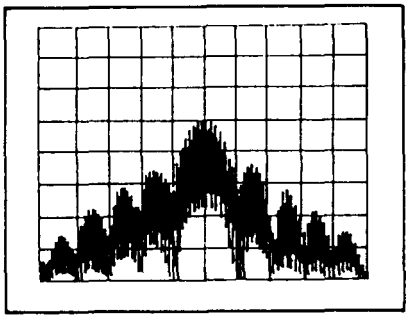
Most real-time spectrum analyzers, however, are preceded by mechanical filters, which limit the input bandwidth of the spectrum analyzer to the desired spectra to be analyzed. Tunable or swept spectrum analyzers function basically the same as heterodyne receivers; the difference being that the local oscillator is not used but is replaced by a voltage control oscillator (VCO). The VCO is swept electronically by a ramp input from a sawtooth generator. The output of the receiver is applied to a CRT, which has its horizontal sweep in synchronization with the VCO. The lower



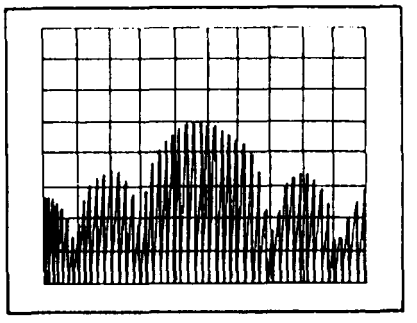
A LINE SPECTRUM OF THE PULSED 50 MHz SIGNAL. LINEAR DISPLAY 100μV/DIV, SCAN 10 kHz/DIV.



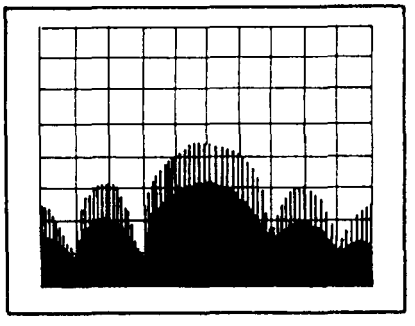
B SAME SPECTRUM IN LOGARITHMIC DISPLAY SCAN WIDTH 10 kHZ/DIV, BANDWIDTH 100 Hz, LOG REF. –20 dBm, 10 dB/DIV.



C SAME SPECTRUM WITH 300 Hz ANALYZER BANDWIDTH. SCAN WIDTH 10 kHz/DIV, LOG REF. –20 dBm, 10 dB/DIV.

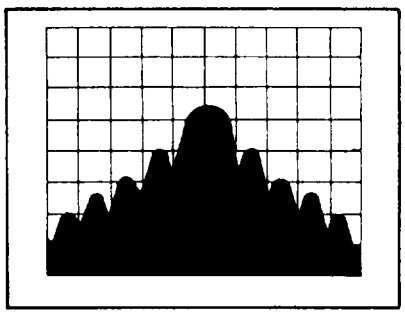


D SAME SIGNAL BUT SCAN WIDTH CHANGED TO 5 kHz/DIV. BANDWIDTH 100 Hz, LOG REF. –20 dBm, 10 dB/DIV.

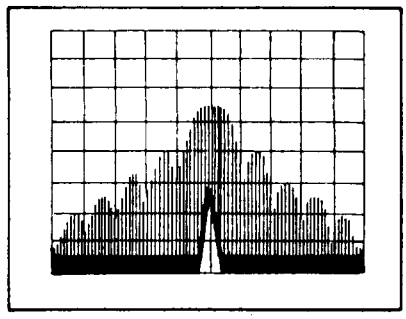


E CARRIER NOW MODULATED WITH A PULSE WIDTH OF τ_{eff} = 0.05 ms (PRF = 1 kHz) SCAN WIDTH 10 kHz/DIV, BANDWIDTH 100 Hz, LOG REF. –20 dBm, 10 dB/DIV.

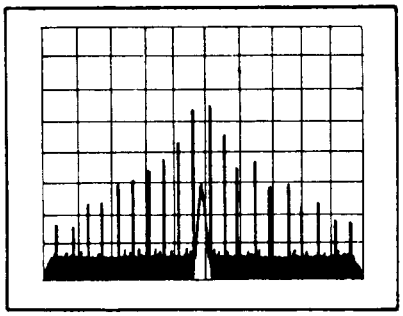
Figure 10-106.-Line spectra of a pulse-modulated 50 MHz carrier.



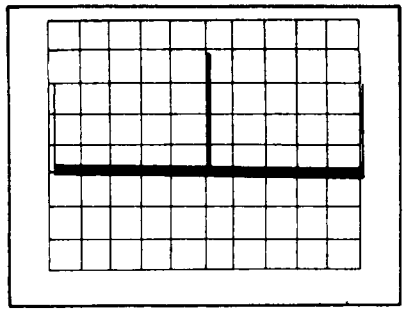
A SIGNAL (PEAK AMPLITUDE –30 dBm) PULSED WITH PRF = 100 Hz, T_{eff} = 1/10 kHz = 100 µs. SCAN WIDTH 10 kHz/DIV, B = 1 kHz SCAN TIME 0.5 s/DIV.



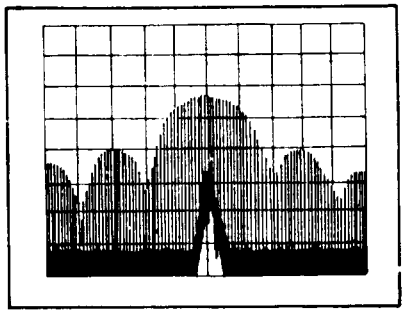
B SAME SIGNAL, BUT SCAN TIME CHANGED TO 0.1 s/DIV.



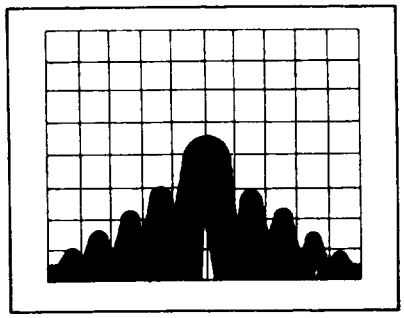
C SAME SIGNAL WITH A SCAN TIME OF 20 ms/DIV.



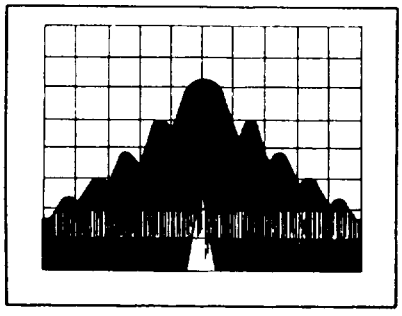
D SAME SIGNAL, BUT B = 300 kHz AND SCAN TIME 2 ms/DIV. THE PRF CAN BE MEASURED TO 1/10 ms = 100 Hz



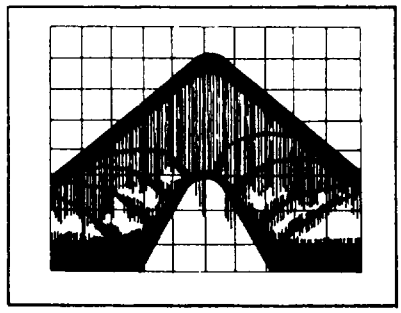
E SAME SIGNAL WITH SCAN WIDTH 5 kHz/DIV, B = 1 kHz, SCAN TIME 0.1 s/DIV.



F SAME SIGNAL WITH B= 300 Hz SCAN WIDTH 10 kHz/DIV, SCAN TIME 0.2 s/DIV



G SAME SIGNAL WITH B = 3 kHz.



H SAME SIGNAL WITH B = 10 kHz.

Figure 10-107.-Pulsed RF signal in "pulsed" spectrum display.

frequency appears at the left of the display. As the trace sweeps to the right, the oscillator increases in frequency. Figure 10-108 is a block diagram of a heterodyne spectrum analyzer.

Resolution

Before the frequency of a signal can be measured on a spectrum analyzer, it must be resolved. Resolving a signal means distinguishing it from other signals near it. Resolution is limited by the narrowest bandwidth of the spectrum analyzer because the analyzer traces out its own IF bandwidth shape as it sweeps through a signal. Thus, if the narrowest bandwidth is 1 kHz, then the nearest any two signals can be, yet still be resolved, is 1 kHz. Reducing the IF bandwidth indefinitely would obtain infinite resolution were it not that the usable IF bandwidth is limited by the stability (residual FM) of the spectrum analyzer. The smaller the shape factor of the IF bandwidth, the greater the analyzer's capability to resolve closely-spaced signals of unequal amplitude. Signals of equal amplitude can be resolved only when they are separated by the 3 dB bandwidth. Unequal signals can be resolved if they are separated by greater than half the bandwidth at the amplitude difference between them.

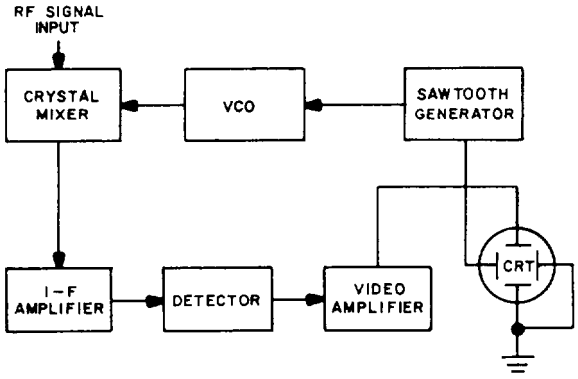


Figure 10-108.-Block diagram of a heterodyne spectrum analyzer.

Other Spectrum Analyzer Considerations

It is important that the spectrum analyzer be more stable infrequency than the signals being measured. The stability of the analyzer depends on the frequency stability of its VCO. Scan time of the spectrum analyzer must be long enough, with respect to the amplitude of the signal to be measured, to allow the spectrum analyzer's IF circuitry to charge and recover. This will prevent amplitude and frequency distortion, as shown in figure 10-109.

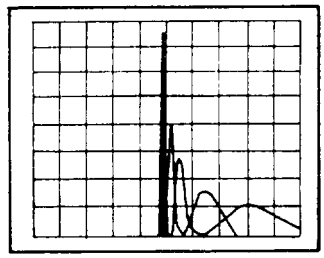


Figure 10-109.-Effects of decreased scan time.

CHAPTER 11

AUTOMATIC TEST EQUIPMENT

Chapter Objective: *Understand the operation and maintenance principles of various types of automatic test equipment to include a radar test station and a more general radio and communication test station.*

Automatic test equipment (ATE), such as the Test Station AN/USM-467 (RADCOM) or the Radar Set Test Station AN/APM-446 (RSTS), is used aboard U.S. Navy aircraft carriers and shore installations. RADCOM and RSTS are designed to deal with the continually changing field of avionics testing. Computerized ATE significantly reduces the space required for testing as compared to space required when equivalent special- and manual-support test equipment is used. This chapter discusses both RADCOM and RSTS equipment.

RADAR AND COMMUNICATIONS (RADCOM) EQUIPMENT

Learning Objective: *Identify the operation and maintenance procedures for the RADCOM* test *bench.*

The station is used to perform diagnostic testing and troubleshooting of radar and communications equipment. The station is capable of generating complex digital and RF signals, ac and dc voltages, and diagnostics to units under test (UUT).

The station contains an automatic self-test system that verifies its operational integrity and automatic self-certification to verify the system parameters.

GENERAL DESCRIPTION

The station consists of four equipment racks separated into two two-rack sections (fig. 11-1) that contain the following sections: section 1, which is the computer section; section 2, which is the display monitor section; section 3, which is the digital input/output (I/O) and miscellaneous test equipment section; and section 4, which is the RF I/O and test equipment rack. Each of the stations contain rack-mounted functional subassemblies. These subassemblies are discussed in the following paragraphs.

The power control electrical assembly is a slide-rack mounted assembly that provides control and routing of the ac power input. An external connection to the thermostat assembly removes ac and dc outputs during an overtemperature condition. An elapsed time meter is mounted on the front panel.

The ac power electrical assembly is a slide-rack mounted assembly. It provides 400 hertz of ac power and 28 volts of dc power to the UUT and detects UUT generated interrupt signals.

The dual disk drive is a slide-rack mounted assembly. It is a dual disk, random access, mass storage device. The dual disk drive contains a fixed disk and a removable disk. It reads information from or writes information to the disks. The disks store the stations operating system and test programs used for testing UUTs.

The filter duct is a prefilter assembly. It provides clean ambient air to the absolute air filter system in the dual disk drive. The filter duct contains a removable polyurethane foam filter element.

The tape reader interface panel provides access to external system diagnostics. System diagnostics from a punched tape reader can be loaded into the computer system via this panel.

The computer controls and supervises station operation. The computer controls the application of stimuli to UUTs and evaluates UUT responses during test programs. The computer contains a control processing unit (CPU) and interfaces with other station components through nine I/O channels.

The thermal graphics printer is the station's data output device and is controlled by the computer. The thermal graphics printer is used to print program listings and the test results of UUTs. A self-contained, automatic self-test is performed at each initialization of power. The thermal graphics printer mounts on a slide-out utility shelf.

The programmable power supply provides constant dc voltages to the UUT and the station. The programmable power supply is a slide-rack mounted unit containing four individual power supplies. The power supplies are controlled by the dc power supply programmer.

The station has four identical blower assemblies located at the bottom of each equipment bay. Each blower assembly provides ambient cooling air for the bay in which it is located. Each blower assembly contains a removable, permanent aluminum falter.

The bus test unit is a slide-rack mounted assembly. It provides electrical interface and programmable protocol for testing UUTs that require communication through the data bus.

The disk controller is a slide-rack mounted unit. The disk controller provides interfacing between the computer and the dual disk drive. It translates command words into usable data for the dual disk drive.

The digital multimeter is a slide-rack mounted programmable digital volt/ohmmeter. The meter measures response voltages and resistances from the UUT. Internal test circuits verify the proper operation of analog/logic circuits and the numeric display.

The color graphics generator is a special-purpose digital processor that operates in two modes. The edit

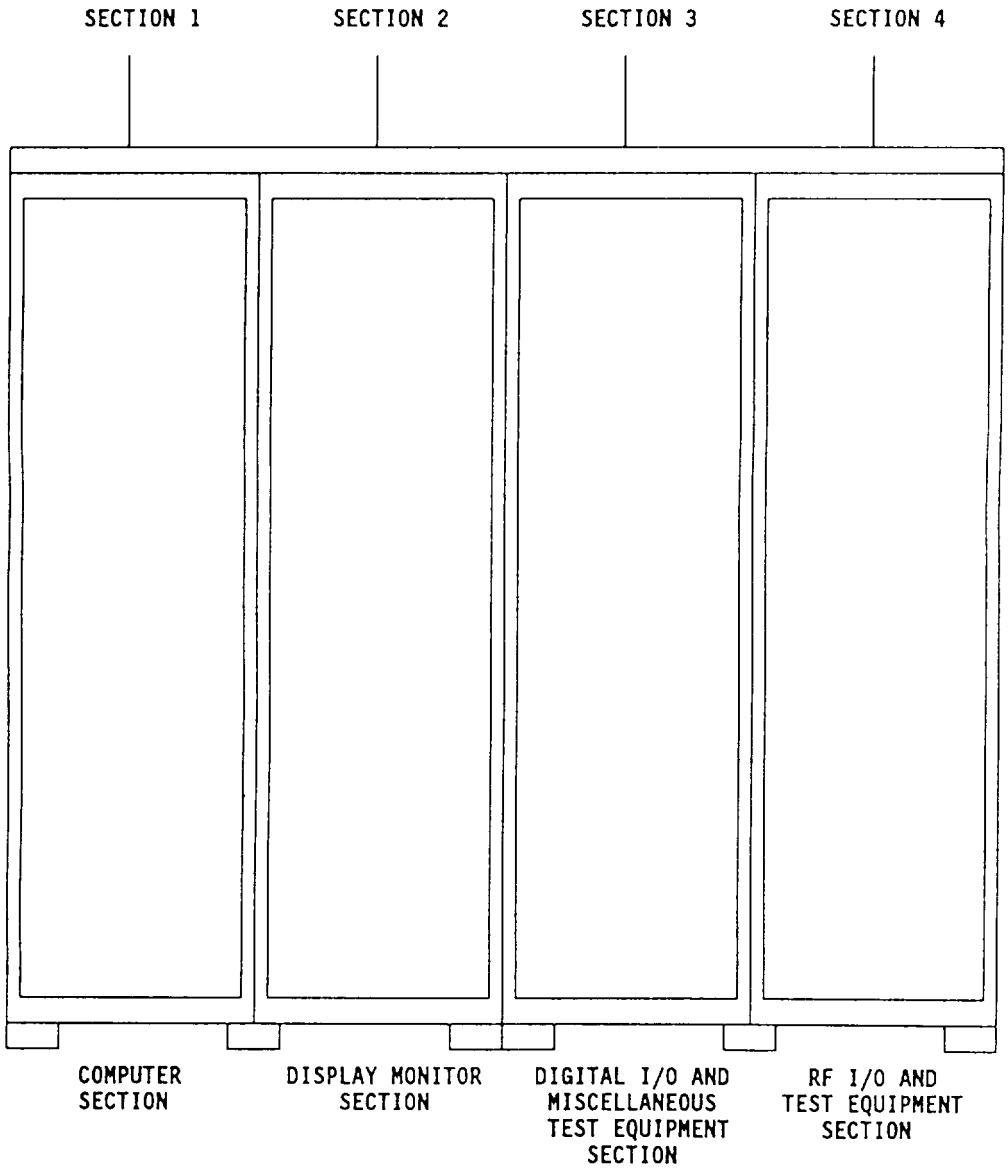


Figure 11-1.-RADCOM station.

mode produces a monochrome alphanumeric display, and the graphic mode produces red/green/blue video signals. The color graphics generator is a slide-rack mounted assembly.

The color graphics monitor is a slide-rack mounted unit containing a commercial color monitor. The color graphics monitor displays alphanumeric and graphic data generated in the color graphics generator.

The keyboard is used to manually input data into the station. It receives operator-entered alphanumeric characters and control codes and provides outputs to the color graphics generator. The keyboard mounts on a slide-out utility shelf.

The dc power supply programmer receives digital data from the computer. The data is converted to analog voltage outputs to control the programmable power supplies. The dc power supply programmer is a slide-rack mounted unit.

The programmable power supply No. 2 provides constant dc voltages to the UUT and the station. The power supply is a slide-rack mounted unit that contains three individual power supplies. The power supplies are controlled by the dc power supply programmer.

The multimode storage oscilloscope is a commercial-type storage oscilloscope with three display modes and four storage modes of operation. The oscilloscope is a slide-rack mounted assembly.

The vector voltmeter is a dual channel voltmeter capable of measuring complex quantities according to magnitude and phase. The vector voltmeter is a slide-rack mounted assembly.

The digital word generator is a high-speed, real-time, automatic test device. It generates stored digital I/O test patterns for the testing of digital devices. The digital word generator is a slide-rack mounted assembly.

The UUT prime power interface panel provides power interface between the UUT and the system. It provides interfacing for programmable 400-Hz voltages, 28 volts of dc power, and UUT interrupts. The interface panel provides power and data interface between the UUT and station assemblies.

The multiple matrix switch performs switching and routing of all analog signals and functions available at the test interface with the UUT. The multiple matrix switch is a slide-rack mounted unit.

The synchro/resolver simulator and angle indicator assembly is comprised of two separate units. The simulator unit is a precision digital-to-synchro/resolver angle signal source, and the angle indicator unit is a precision analog-to-digital converter. The synchro/resolver simulator and angle indicator assembly is a slide-rack mounted assembly.

The arbitrary function generator is designed to provide three independent signal sources to test UUTs and station circuits. The arbitrary function generator is a slide-rack mounted assembly.

The programmable power supply provides constant dc voltage to the UUT and the station. The power supply is a slide-rack mounted unit that contains two individual power supplies. The power supplies are controlled by the dc power supply programmer.

The automatic spectrum analyzer is a high-performance, two-band analyzer. It is capable of operating at low frequencies or in the microwave range. The analyzer is a slide-rack mounted assembly.

RF generators No. 1 and No. 2 are identical, Only RF generator No. 1 is discussed here. RF generator No. 1 is a broadband, analog sweep, frequency synthesizer. It has a versatile control for both modulation and output. The generator is a slide-rack mounted assembly.

The RF interface unit provides complete RF signal switching and distribution, video detection, input attenuation, preamplification, and RF power sensor capabilities. The interface unit is a slide-rack mounted assembly.

The RF generator No. 3 is a precision synthesized signal generator with excellent stability and spectral purity. The RF generator No. 3 is a slide-rack mounted assembly.

The auxiliary equipment interface panel provides ac and dc power and data interface between the UUT and station assemblies.

The signal analyzer provides measurement capabilities for use in computer-controlled analog test systems that test UUTs and analog circuits requiring pulse, square, sine, ramp, or triangular waves. The signal analyzer is a slide-rack mounted unit.

The power meter and frequency oscillator is a slide-rack mounted unit. The power meter provides automatic and manual measurement of RF and microwave power levels. The frequency oscillator provides a frequency stable 10-MHZ output signal from a voltage-controlled crystal oscillator.

PRINCIPLES OF OPERATION

The station is divided into five subsystems. See figure 11-2. The subsystems are the

- station power and power distribution subsystem, the
- display and control subsystem, the
- analog video subsystem, the
- RF subsystem, and the
- digital subsystem.

Station Power and Power Distribution Subsystem

The power control unit provides control and routing of the three-phase, 120-Vac, 50/60 Hz to the four power strips. It also supplies a single-phase, 120-Vac, 50/60 Hz slave output for use in auxiliary equipment. The power control unit contains two regulated power supplies that generate -5, +5, -12, +12, -24, and +24 dc voltage outputs to the multiple matrix switch. An internal power monitor is used to monitor the ac input voltage, and it provides for the removal of ac and dc output voltages when an overvoltage, undervoltage, or an incorrect phase rotation condition occurs. An external connection to the thermostat assembly removes ac and dc outputs during an overtemperature condition. The elapsed time meter, which is mounted on the front panel, records system operating time.

Display and Control Subsystem

During normal testing the operating system is resident in the computer memory. Test languages or programs are read into the computer from the dual disk drive via I/O port 11. The computer contains 11 hardware registers, and in conjunction with individual UUT test programs, it controls and integrates all of the stations components. Operator data is supplied through I/O port 14 to the color graphics generator and monitor. Diagnostic routines are permanently stored in a read-only memory (ROM). An automatic test checks the central processing unit (CPU) and the memory to ensure and verify the operational status.

The dual disk drive has a total memory capacity of 20 megabytes. Data is accessed through three surfaces by three read/write heads. The dual disk drive contains a direct drive motor to drive the disks at a constant 3600 RPM. A position and velocity servo system positions the heads on the surface of the disks. Transfer of information to and from the computer is made via the disk controller.

The disk controller provides interface between the computer and the dual disk drive through the use of a microprocessor card, a CPU card, and a device controller card. An error correct card determines if the data supplied to the computer is free of errors and stores operating instructions for the disk controller.

Keyboard data from the keyboard is fed directly into the color graphics generator. The generator contains four refresh memory cards, a microprocessor memory card, microprocessor card, microprocessor I/O card, and a video card. Video and graphic signals are generated and controlled for use in the color graphics monitor.

The monitor displays video data in two separate modes of operation as controlled by the color graphics generator. The monitor consists of eight modules. The monitor is removed from the station, and repairs are accomplished at depot. Details for the principles of operation are available only at that level of maintenance.

Operator data is manually entered into the station via the keyboard. The keyboard contains 82 keys, and is capable of performing 140 functions. Each key stroke is converted into a +5 Vdc signal, which is encoded into a 9-bit binary-coded signal. The binary-coded signal is then sent to the color graphics generator.

The thermal graphics printer provides output data information. A printer control assembly, in conjunction with the front panel, power supply, and stepper motor assembly, controls the print head. Print head control line data and serial dot patterns are fed into the print head assembly, and results in printed data output. I/O port 13 provides input data.

Analog and Video Subsystem

The multiple matrix switch provides all of the switching and distribution for the analog and video subsystem. The signals available at the switch interface panel include power, stimuli, response lines, and all of the analog signals and functions. The switch contains 12 circuit cards and 20 cable assemblies. Interface with the computer is I/O port 15. All of the other units, except for the oscilloscope in the analog and video subsystem, are cable connected to the multiple matrix switch. The switch provides interface with all of these devices. The switch also receives data from the RF subsystem as well as the wizard digital logic probe through the interface panel.

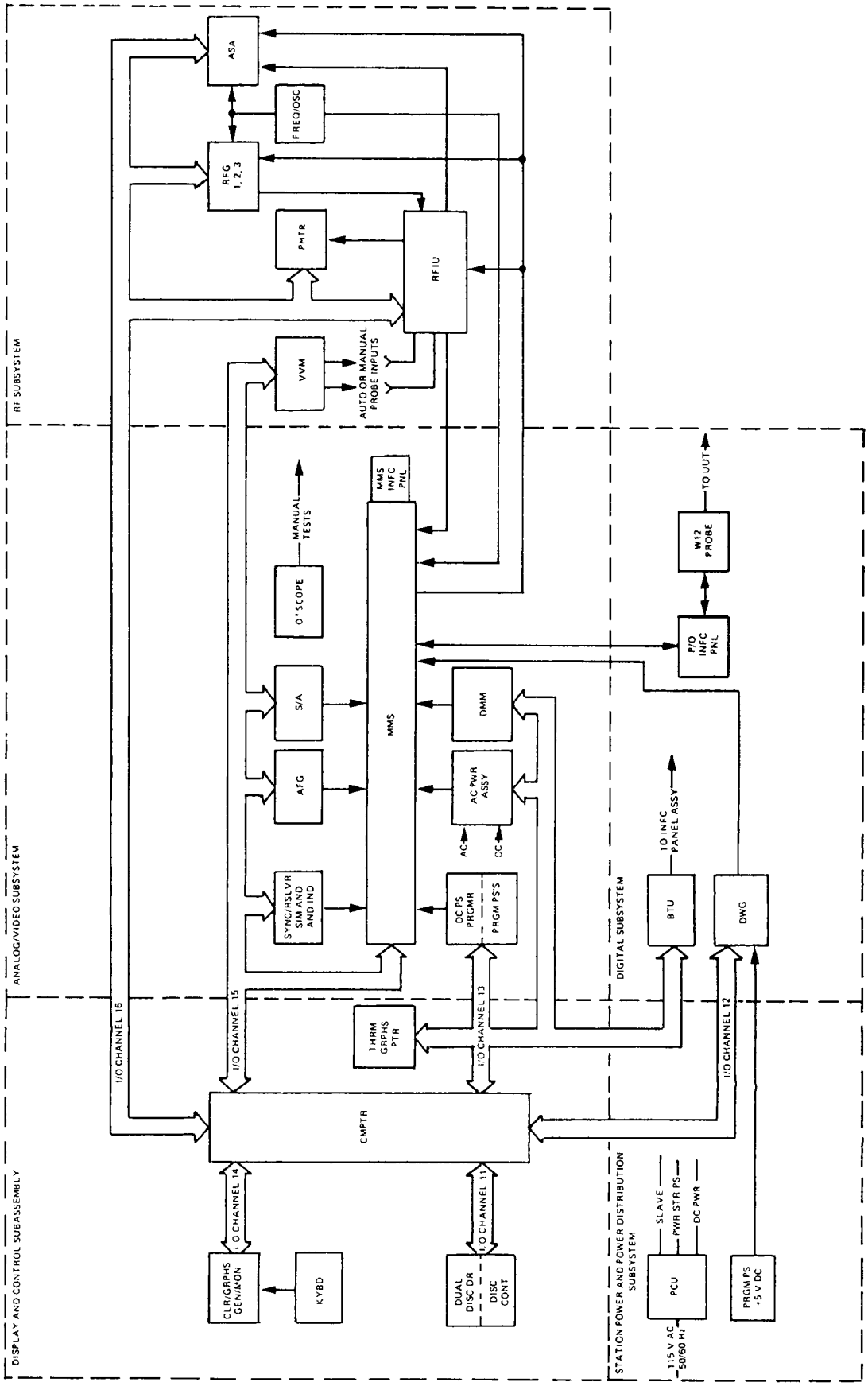


Figure 11-2.–Test station AN/USM-467.

As mentioned earlier, the synchro/resolver simulator and angle indicator is comprised of two separate shop replaceable assembly (SRA) units. The simulator is a precision digital-to-synchro/resolver angle indicator source. The four circuit cards provide three-wire synchro or four-wire resolver output voltage source signals. The angle indicator is a precision analog-to-digital converter. The indicator contains three circuit cards, which continuously digitizes and displays synchro or resolver data in real time.

The arbitrary function generator provides for three independent signal sources. Each of the three outputs can be programmed as a standard wave generator, an arbitrary function generator, or as a pulse generator. The signals available are pulse, square wave, sine, ramp, triangular, or those that are user defined. The arbitrary function generator consists of eight circuit cards.

The signal analyzer functions as three independent instruments–a digital multimeter, a counter-timer, and a digital sampler. Frequency measurements, period measurements, and time-interval measurements can all be made. The analyzer consists of six unique circuit cards for these functions.

The station consists of three completely separate, individual, programmable power supply assemblies. The power supply assemblies are each controlled by the dc power supply programmer. Each of the power supplies are discussed separately in the following paragraphs.

Programmable power supply No. 1 is located in equipment rack No. 1. This power supply contains four individual SRA power supply assemblies. Each is controlled by the dc power supply programmer and provides constant dc voltages for the UUT. The two power supplies located on the left side of the power supply drawer are identical and provide 0 to 55 Vdc rated at 0 to 2 amperes. The two power supplies located on the right side of the power supply drawer are identical and provide 0 to 36 Vdc rated at 0 to 3 amperes.

Programmable power supply No. 2 is located in equipment rack No. 2. This power supply contains three individual SRA power supply assemblies and a dc power supply programmer assembly.

NOTE: The dc power supply programmer is not used in this station.

Each of the three power supplies is controlled by the dc power supply programmer and provides constant dc voltages for the UUT. The power supply located on the left side of the power supply drawer provides 0 to 15 Vdc rated at 0 to 6 amperes. The two power supplies located in the middle of the power supply drawer are identical and provide 0 to 36 Vdc rated at 0 to 3 amperes.

Programmable power supply No. 3 is located in equipment rack No. 3. This power supply contains two individual SRA power supply assemblies. Each is controlled by the dc power supply programmer and provides constant dc voltages for the UUT and the digital word generator located in the digital subsystem.

The power supply located on the left side of the power supply drawer provides 0 to 25 Vdc rated at 0 to 10 amperes. The power supply located on the right side of the power supply drawer provides +5 Vdc 3 Vdc rated at 0 to 55 amperes only to the digital word generator.

The dc power supply programmer controls the three individual power supplies of the station. Power supply outputs are received by the programmer and applied to the multiple matrix switch for use in UUTs. The programmer receives digital data from the computer via the I/O port 13. This digital data is converted into an analog voltage, which is used to control the outputs of each power supply. The dc power supply programmer contains eight identical programming circuit cards.

The ac power assembly contains a bus interface card, a motor control card, two motorized variable transformers, and a dc power supply. The ac power assembly receives 115 Vac, 400 Hz, and 28 Vdc from the facility power source. It provides 400-Hz ac power and 28 volts of dc power to the UUT through the multiple matrix switch. It detects interrupt signals generated by UUTs and provided through the switch. The ac power assembly provides the following output voltages:

- 115 Vac, 400 Hz, three-phase (WYE)
- 0 to 130.0 Vac, 400 Hz, three-phase (WYE)
- 0 to 130.0 Vac, 400 Hz, phase A
- 28 Vac, 400 Hz, phase A
- 28 Vdc

The digital volt/ohmmeter is a programmable digital volt/ohmmeter. The meter is capable of measuring ac voltages, dc voltages, and resistances from a UUT as provided by the multiple matrix switch. Interface with the computer for control is through I/O port 13. The switch contains an internal test function to

verify proper operation of its analog/logic circuits and the numeric display.

The oscilloscope is a storage-type scope made up of three individual plug-in assemblies. The scope has three separate display modes of operation. It also provides the operator with four selectable modes for storage, allowing for the retention and viewing of all fast rising, low repetition rate, single shot, or any slow moving waveforms.

RF Subsystem

The RF interface unit provides complete RF signal switching and distribution of the assets of the RF subsystem to the UUT and other station assemblies through the multiple matrix switch. Computer control is through I/O port 16. The interface unit receives, controls, amplifies, and measures the phase of signals from RF generators 1, 2, and 3. The signal analyzer and power meter receive RF output signals from the interface unit. The unit contains three circuit cards, a power sensor, reference attenuator, and other electronic circuits.

The RF generator No. 1 is a versatile synthesized sweeper that generates frequencies from 10 MHz to 26.5 GHz. It also provides for a linear sweep of either frequency or power across the entire range of outputs. Computer control of RF generator No. 1 is through I/O port 16. The generator provides the UUT with pulse modulated or AM modulated synthesized frequencies through the RF interface unit. The generator receives a 10-MHZ frequency reference input from the frequency oscillator. The generator contains 61 SRAs.

The RF generator No. 3 is a highly accurate low noise RF sweeper that generates frequencies from 10 KHz to 1280 MHz. The generator is capable of delivering continuous wave (CW) frequencies that have been AM or FM modulated. It also provides for a linear sweep of either frequency or power across the entire range of outputs. Computer control of RF generator No. 3 is through I/O port 16. RF generator No. 1 provides the UUT with CW AM or FM modulated synthesized frequencies through the RF interface unit. The generator receives a 10-MHZ frequency reference input from the frequency oscillator. The RF generator No. 3 contains 61 SRAs.

The automatic spectrum analyzer provides for a broadband frequency and time analysis. It operates at 110 Hz to 2.5 GHz and 2 GHz to 22 GHz using two bands. UUT frequencies to be measured and analyzed are routed through the multiple matrix switch to the automatic spectrum analyzer. Frequency outputs from the RF interface unit, as well as the frequency standard of the frequency oscillator, are also sent to the analyzer. Computer control of the analyzer is through I/O port 16.

> **NOTE:** The power meter/frequency oscillator are actually two SRA units housed in a single drawer. They are separated in the following paragraphs only for ease in the explanations of their individual functions.

The frequency oscillator generates a highly stable 10 MHz signal for use in standardizing the units of the RF subsystem. The signal analyzer output of the oscillator is used to phase lock the signal analyzer. Outputs to RF generators 1, 2, and 3 are used as reference and for phase locking. The 10-MHZ frequency is also used as a reference source for UUTs and other station functions through the multiple matrix switch.

The power meter is the stations standard for the measurement of absolute power. It is also a transfer standard for the measurement of relative power by remotely using the automatic spectrum analyzer. Input frequencies to be measured are applied from the RF interface unit. Computer control of the power meter is through I/O port 16. The power meter sensor unit is remotely located in the RF interface unit.

The vector voltmeter is a dual channel voltmeter. It measures signals in magnitude and in phase, and is capable of performing a ratio measurement between the two channels. The voltmeter has a manual capability where the operator may select to make measurement using the two attached probes. For an automated test measurement, the probes are connected to the RF interface unit, and computer control of the voltmeter is through I/O port 15.

Digital Subsystem

The digital word generator provides the station with the ability to perform real-time, high-speed testing of digital devices. It is capable of dynamic testing of UUTs as well as the units of the station. The generator receives +5 Vdc logic operating power from an individual power supply (dc programmable power supply located in the station power and power distribution subsystem). Communication with the computer is through I/O port 12. Inputs to the station and outputs from UUTs are

through the multiple matrix switch. The generator contains 12 printed circuit cards.

The bus test unit is a programmable communications controller that links the station to a UUT. The bus test unit can be programmed to provide the following to UUTs:

1. Manchester 1-0) or binary encoding
2. Parity
3. Message and sync signature
4. Number of data bits/word
5. Internal or external triggering
6. Data rate (source and range)
7. Response or response delay time
8. Intermessage
9. Trigger delay time

The test unit communicates with the computer through I/O port 13. The programmed outputs of the test unit are applied to the panel assembly.

The wizard digital logic probe is a hand-held probe used to fault isolate failed components on card assemblies. A probe logic circuit card is located in the multiple matrix switch. The probe can communicate with the computer and the digital word generator through part of the panel assembly to the multiple matrix switch.

RADAR SET TEST STATION (RSTS)

Learning Objective: *Identify radar set test station principles of operation* to *include station power and an overall block diagram description.*

The station is used to perform diagnostic testing arid troubleshooting of radar equipment. The primary test station that we will discuss is the AN/APM-446.

GENERAL DESCRIPTION

The Radar Set Test Station (RSTS) AN/APM-446 is comprised of three units with four and one-half bays of equipment, as shown in figure 11-3. Bay 1 contains a holding fixture for large weapon replaceable assemblies (WRA). Bay 2A1 also contains a unit under test (UUT) support surface for holding smaller WRA and shop replaceable assemblies (SRA) during testing.

The system is fully maintainable from the front of the bays and top of bay 1. All elements requiring periodic maintenance are accessible from the front of the cabinet. Multiple assemblies are mounted, using slides that provide for access to all interfacing cables.

The RSTS and associated test program sets provide automatic test capabilities for fault isolation of 6 WRA and 21 SRA of the F/A-18 radar system (AN/APG-65). The station is capable of implementing performance and diagnostic tests, WRA/SRA alignments/adjustments, and physical and electrical requirements.

PRINCIPLES OF OPERATION

The RSTS can functionally be divided into the following subsystems:

- Station power
- Station control
- Stimulus and loads
- Measurement
- Switching
- Cooling

A general description of each subsystem unit and how each unit integrates into the system as a whole is presented in this section of the chapter.

Station Power

The station power distribution unit, in conjunction with the control panel assembly and high current power distribution unit, distributes power to the various RSTS test equipment. These three units also distribute prime and programmable power to the UUT via the high current power distribution unit. The station power distribution unit supplies overall power for the entire station via the buses.

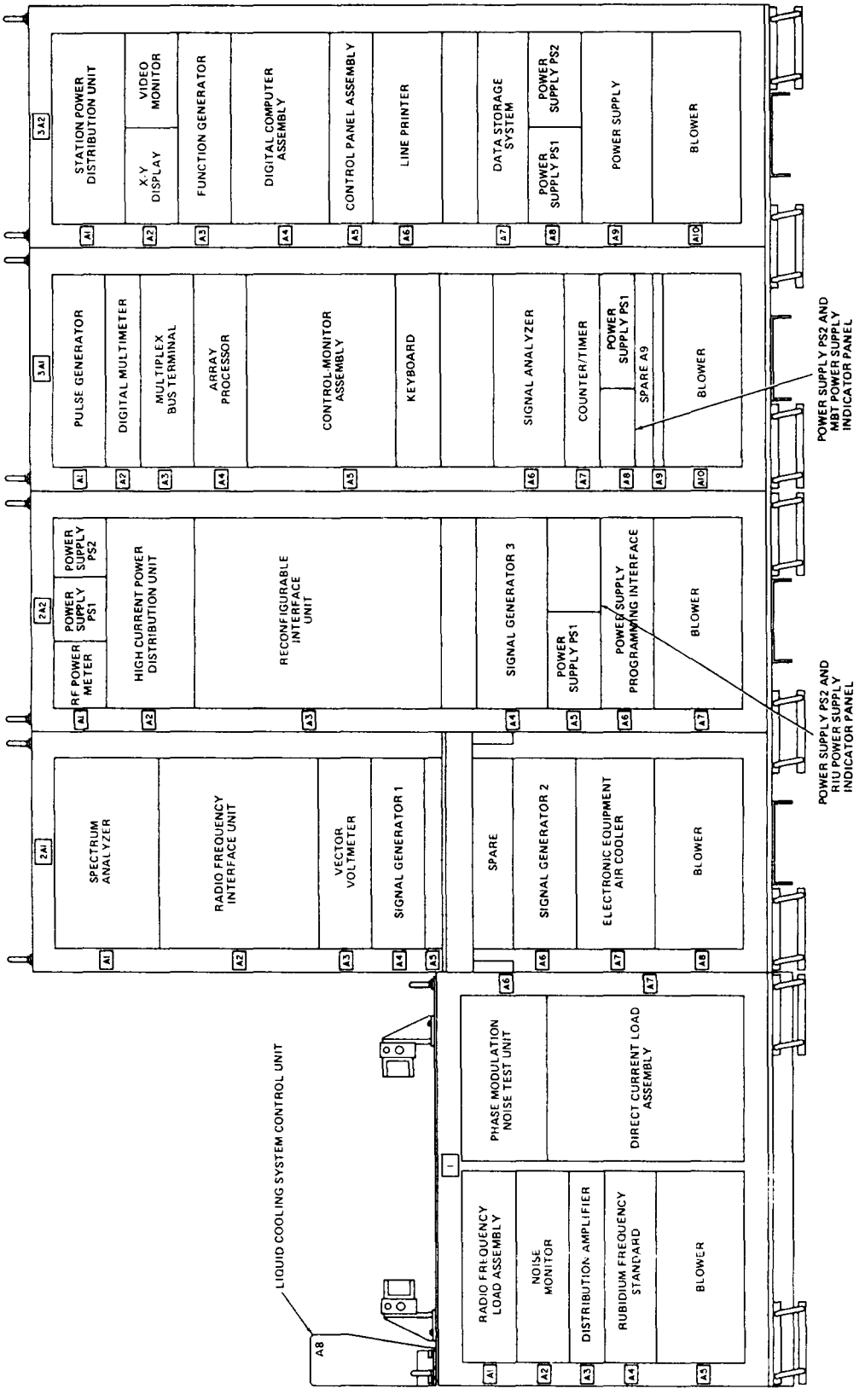


Figure 11-3.–Radar Set Test Station AN/APM-446.

There are four functional type buses used for routing ac power within the station. Refer to figure 11-4. These are the computer bus, the equipment bus, the stand-by bus, and the blower bus. The computer bus routes ac power to the computer and those instruments and assemblies that interface directly with the computer data bus. These include the line printer, disk drive, array processor, multiplex bus terminal, and the color monitor/keyboard. The equipment bus is used to route ac power to the instrumentation and assemblies within the system. The standby bus provides a continuous uninterruptible source of power for the rubidium frequency standard. The blower bus provides a means of separately switching power on and off for the blowers located at the bottom of each bay.

The Station Power Distribution Unit 3A2A1 monitors supplied 120-Vac, 60-Hz, three-phase input, and 115-Vac, 400-Hz, three-phase input power for overvoltage/undervoltage conditions, proper frequency, and phase rotation. Refer to figure 11-5. The station power distribution unit will disconnect input power from all associated circuitry when any nonconforming conditions are detected. The station power distribution unit also provides monitoring of overvoltage/ undervoltage conditions for the supplied 28 Vdc.

The Control Panel Assembly 3A2A5 contains the master power on and off switches and separate equipment on and off switches that allow the equipment bus to be powered down when only the computer and peripherals are being used. It also contains blower 400 Hz and 28 Vdc operation indicators.

The high-current Power Distribution Unit 2A2A2 routes 115 volts ac, 400 Hz, and 28 volts dc and adjustable three-phase 0 to 140 volts ac, 400 Hz to the UUT. The unit contains the emergency off switch, located on the front panel. The unit also performs switching of five high-current power supplies.

The Power Supply 2A2A5 PS2 and reconfigurable interface unit power supply indicator panel provides 5 Vdc and 24 Vdc to the interface unit and panel lamp indication of power to the reconfigured interface unit.

The Power Supply 3A1A8 PS2 and multiplex bus terminal power supply indicator panel provide 5 Vdc 30 amperes, 15 Vdc 4 amperes, and 15 Vdc 2 amperes to the MBT and panel lamp indication of power to the bus terminal.

The station power distribution unit requires 120-Vac, 60-Hz, three-phase power. This power input is routed through RFI line filters FL1, FL2, and FL3 to main circuit breaker CB1, computer circuit breaker CB2, and to the equipment circuit breaker CB7. These circuit breakers provide overcurrent protection.

The power is routed from CB1 to voltage monitor VM1, power supply PS1, circuit breaker CB4, and the CPA. Phase C is fed back to one side of the auxiliary trip coil of CB1. When the emergency off switch on the high current power distribution unit is depressed, the trip coil

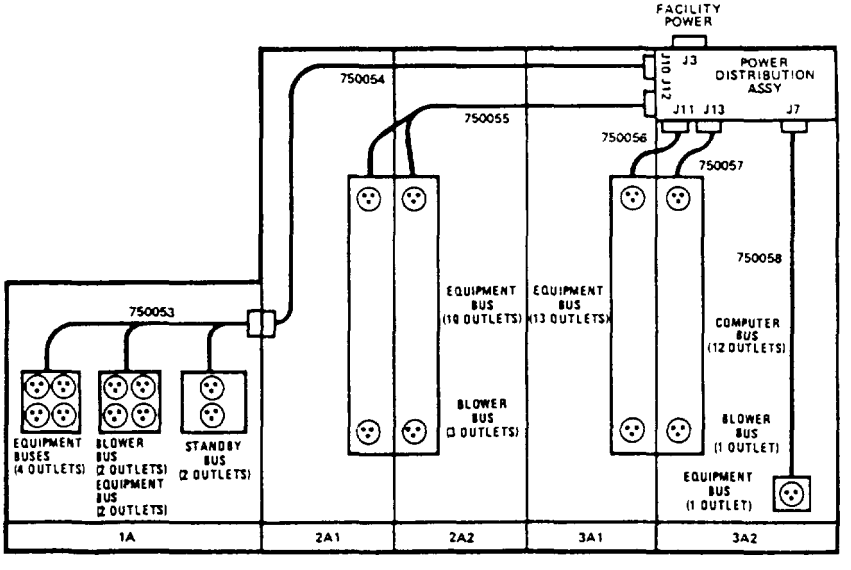


Figure 11-4.-RSTS ac power distribution.

circuit is completed, which de-energizes CB1, and via interlocked relays, removes all power sources from the RSTS and UUT. Monitor VM1 contains sensors that monitor the voltage, frequency, and phase rotation of the power. When voltage and frequency of the input power are within prescribed high and low limits and phase rotation is A-B-C, the output relay of VM1 will energize after a 2-second time delay. When energized, the output relay permits additional routing of phase C and 24-Vdc power.

Phase C power is routed from the VM1 output relay through standby circuit breaker CB3 to the RSTS rubidium frequency standard 1A4. This circuit breaker provides overcurrent protection for 1A4.

The VM1 output relay also routes the MASTER ON interlock signal from the control panel assembly MASTER ON switch through relay K4 and the circuit breaker CB2 auxiliary contact to the 24 Vdc return of PS1. In the depressed ON position, the control panel assembly MASTER ON switch routes the 24 Vdc return to the station PDU on the MASTER ON bus, energizing dc relays K2 and K3. Relay K2 receives 24 Vdc coil power from PS1 and 120-vac, 60-Hz, three-phase power through circuit breaker CB2. Phase A power is routed to the time totalizing meter M1, and three-phase power is routed to the station computer through the K2 relay contacts. Overcurrent protection for the computer is provided by CB2. Relay K3 receives 24 Vdc coil power from PS1 and 120 vac, 60-Hz, three-phase power through circuit breaker CB4. Three-phase power is routed to the station blowers from K3. Overcurrent protection for the blowers is provided by CB4. lime-delay relay dropout K4 receives 24 Vdc from station interrupts via the computer. If an overtemperature condition occurs, station interrupt temperature sensor contacts will open and the 24 Vdc will be removed from K4. After a specified time, delay K4 will drop out, removing power from the RSTS.

Phase B power from CB1 is applied to the primary of PS1. The secondary 24-Vdc output of PS1 provides the following:

- Power to indicator lights and switches on the control panel assembly
- Power to station interrupts (temperature sensors)
- Control power to dc relay K4 via station interrupts
- Control power to dc relays K1 through K3 with power returns via the control panel assembly

Phase B is also routed via control panel assembly circuit breaker CB1 to two electrical outlets located on the control panel assembly front panel.

Power is routed through circuit breaker CB7 and ac relay K7 to various RSTS equipment. Overcurrent protection for the station equipment is provided by CB7. Relay K7 is controlled by dc relay K1, which has its 24-Vdc coil return routed through the NC contacts in the control panel assembly EQPT OFF switch. When K1 is energized, the return for phase C power routed to the K7 ac relay coil is completed, and three-phase power is routed to the station equipment. Phase C power is also routed from CB7 to the high current power distribution unit to operate an ac relay coil. CB7 via its auxiliary contact also connects either the control power assembly EQPT ON or EQPT OFF light via the equipment power control 24 Vdc return, depending on the switch and circuit breaker positions.

The station power distribution unit requires 115-Vac, 400-Hz, three-phase power. Power input is routed through circuit breaker CB5, RFI line filters FL5, FL6, and FL7 to voltage monitor VM2 and dc relay K5.

The operation of VM2 is identical to that of VM1, as discussed earlier. When energized, the output relay of VM2 switches the following signals:

1. The VM2 output relay connects the 400-Hz light on (24 Vdc) signal from the control power assembly to the 24 Vdc return of PS1 to light the CPA 400-Hz indicator light. When the light is on, the presence of power is verified to be within prescribed limits and phase rotation.

2. The VM2 output relay also switches on a 115-Vac, 400-Hz ON signal (24 Vdc) to the computer. This signal is routed from the reconfigurable interface unit via the high current power distribution unit to VM2 of the station power distribution unit out to the computer. The signal informs the computer that power is available and is within prescribed limits and phase rotation.

3. The VM2 output relay also completes the 115-Vac, 400-Hz on control signal (24 Vdc) from the high current power distribution unit through the relay K5 coil and VM2 back to the high current power unit and reconfigurable interface unit. When energized, K5 routes 400 Hz, three-phase power from the line filters through the high current power distribution unit to the UUT. Relay K5 also switches the self-test signal for the 115-Vac, 400-Hz power. This signal is routed from the high current power distribution unit through K5 to the interface unit via the high current power distribution unit

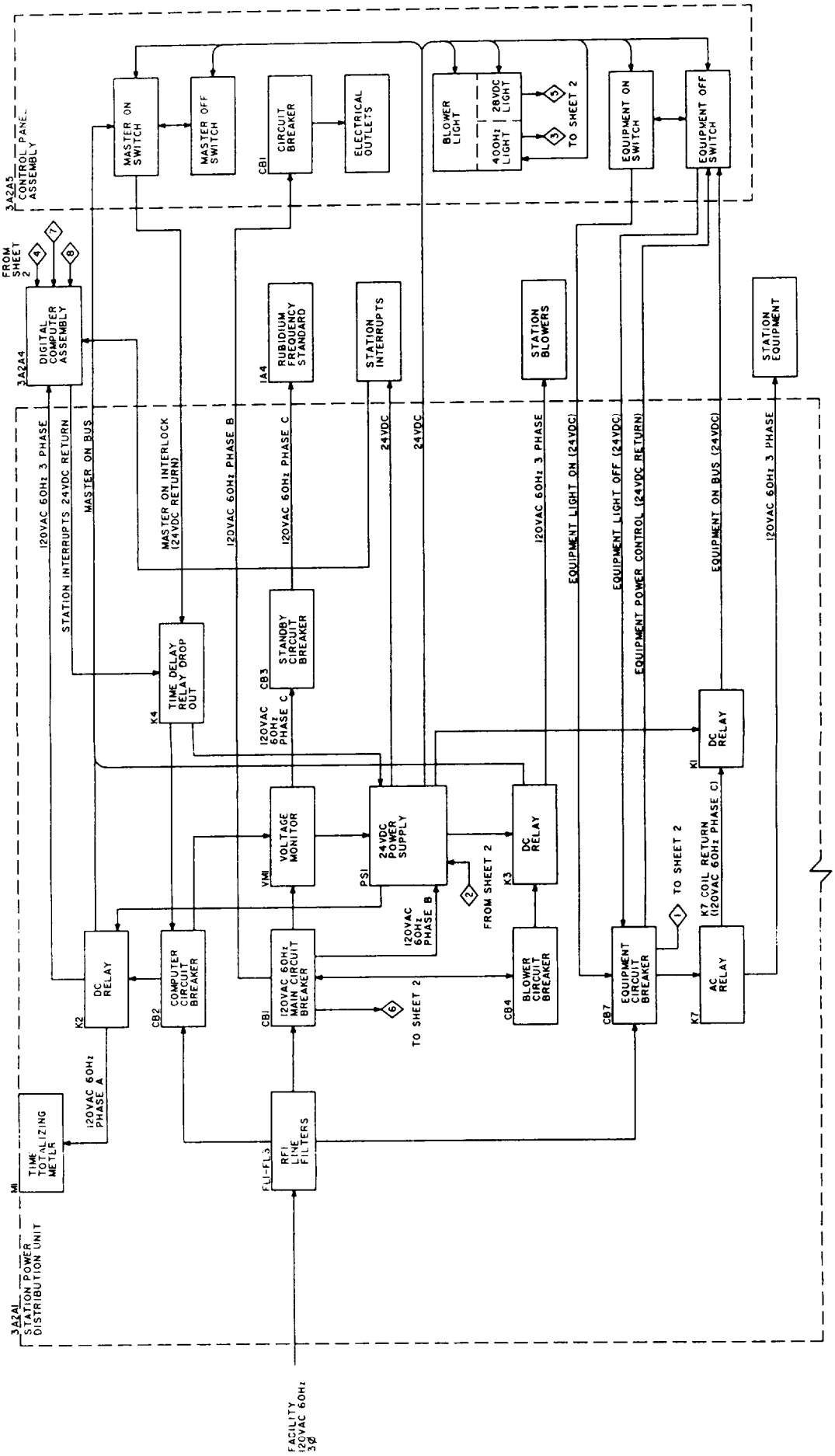


Figure 11-5.-RSTS power distribution block diagram.

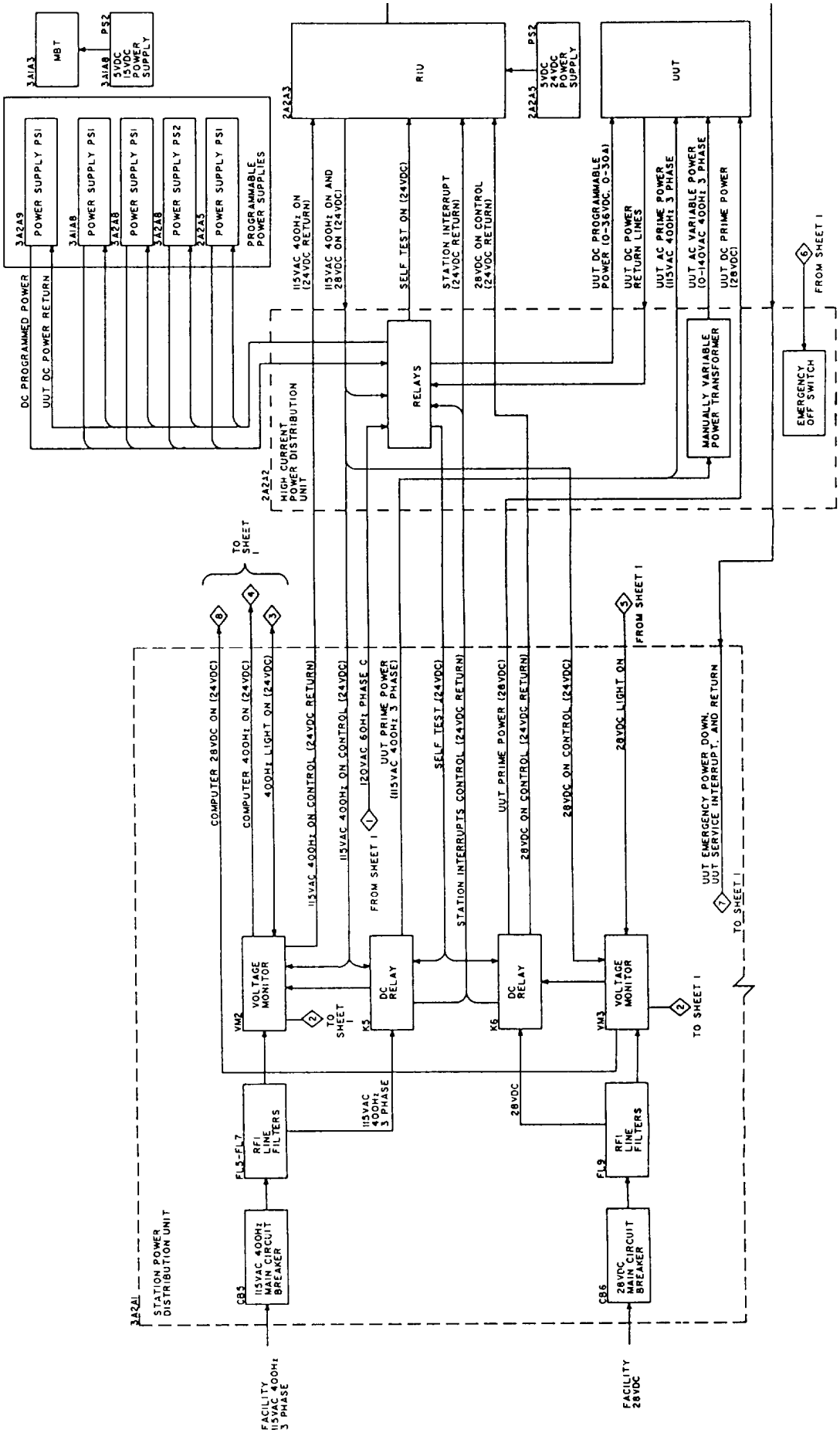


Figure 11-5.–RSTS power distribution block diagram–Continued..

on the station interrupts control line (24 Vdc RTN) during self-test.

The station power distribution unit requires 28 Vdc. Power is applied to main circuit breaker CB6 and routed through RFI line filter FL9 to voltage monitor VM3 and dc relay K6. Filtered facility power is now referred to as prime power.

Monitor VM3 operates as a power disconnect via relay K6 when the dc line voltage is beyond prescribed limits from 24 Vdc to 29 Vdc. When power is within these limits, the output relay of VM3 is energized and engages the following signals:

1. The VM3 output relay connects the 28-Vdc light on (24 Vdc) signal from the control panel assembly to the 24 Vdc return of PSI to light the control panel assembly 28 Vdc indicator light. Power is verified to be within prescribed limits when this light is illuminated.

2. The VM3 output relay also switches on relay K6 by connecting the 28-Vdc ON control line (24 Vdc) from the interface unit via the high current power distribution unit through VM3 and the relay K6 coil back to the high current power distribution unit. Relay K6 switches the 28 Vdc prime power to the UUT via the high current power distribution unit.

Relay K6 also switches on the self-test signal for the 28 Vdc power. This signal is routed from the high current power distribution through K6 to the interface unit on the station interrupt control line (24 Vdc RTN) during self-test.

Station Control

The station control subsystem controls all RSTS functions and monitors RSTS status over the appropriate buses and interfaces. Refer to figure 11-6 for the station control subsystem block diagram. The RSTS is controlled by a digital computer assembly located in bay 3A2. The main operator interface is provided by the color monitor and keyboard located in bay 3A1. A data storage system is included for storage of all system software, including the station executive program as well as UUT test programs. A line printer is available to provide printouts of test results as well as program listings and other information, which must be retained for future reference. All station assets are controlled via four independent IEEE-488 buses and three RS-232C buses. High-speed communication channels are used to transfer information between the station computer and WRA. The UNIBUS provides the internal address and data bus for communication

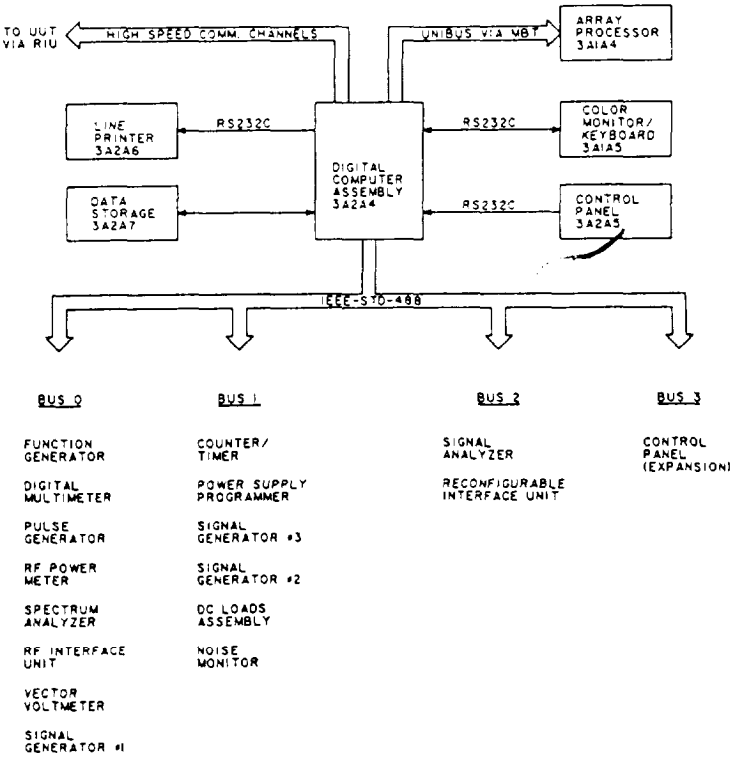


Figure 11-6.-Control subsystem block diagram.

between the central processor, its modules, the multiplex bus terminal, and the array processor.

The RSTS hardware uses interrupt signals to monitor certain components throughout the station. During normal system operation, each bay is monitored by a separate thermal switch. These switches are mounted at the top of each bay near the air outlet. These switches have NC contacts, which open at 140$F4.9$F, and are wired to provide separate interrupt signals to the central processor. In the event of a thermal interrupt, the location is identified and printed on the printer and station power is removed.

Other interrupts that are continuously monitored by the central processor include liquid coolant interrupts, which indicate abnormally high coolant temperature or pressure, coolant level in the reservoir high or low, low nitrogen reservoir; an RF loads interrupt, which indicates no coolant flow through the RF loads; a UUT air-cooling unit interrupt; a UUT emergency power down interrupt; a UUT service interrupt; and power distribution interrupts, which indicates either the 28-Vdc or the 400-Hz power is not present.

The PDP 11/44 computer 3A2A4 controls all station instrumentation via four IEEE-488 general-purpose interface buses and executes all system software. The computer interfaces with all station equipment directly or indirectly.

The Data Storage System 3A2A7 contains an input tape drive for the loading of all station programs. It also contains an 8-inch Winchester disk drive, a proprietary controller for each, and a host adapter for each controller to run the program. The data storage system interfaces with the high-speed communication channels of the computer.

The Color Monitor/Keyboard 3A1A5 provides information and real-time displayed messages to the operator concerning system status, operational test status, appropriate control functions, and/or required operator actions. The keyboard allows the operator to control operation of the test station by entering control statements, numerical information, test information, and test executive commands.

The Line Printer 3A2A6 provides the station with line printing communication capabilities for the UUT testing and self-testing.

The Multiplex Bus Terminal 3A1A3 is a programmable device that simulates a MIL-STD-1553 data bus system for use in testing the various WRA. It has a high-speed parallel interface for real-time simulation. The terminal interfaces with the reconfigurable interface unit and computer.

The Array Processor 3A1A4 provides the station with high-speed, floating-point mathematical computation capabilities. It is capable of handling large amounts of data in a short period of time. It interfaces with the reconfigurable interface unit, multiplex bus terminal, and the station computer. The station computer UNIBUS terminator module is normally contained in this unit.

RSTS instruments are controlled and accessed through three IEEE-488 general-purpose interface buses. A fourth IEEE-488 bus is only used during self-certification. Each bus is interfaced to the computer through a GPIB 11-2 IEEE direct memory access interface card. Bus instruments are interconnected in a parallel arrangement (16 signal lines and 8 ground lines) to provide an orderly flow of data to and from the instruments. Three handshake lines provide the means to asynchronously transfer data between instruments. Each instrument is identified on its bus by a unique bus address number.

The RS-232C buses interface the computer to the color monitor and line printer, and provide an external interface located at the control panel assembly.

High-speed communication channels transfer large amounts of data from the RSTS computer memory directly to the radar target data processor and computer/power supply WRA. Circuit cards high-speed comm channel (HSCC) No. 1 and HSCC No. 2 work as a pair within the RSTS computer to accomplish the data transfer.

Refer to figure 11-7 for the following functional theory. The HSCC No. 1 card takes control of the unibus when large data transfers occur. The HCSS No. 1 card

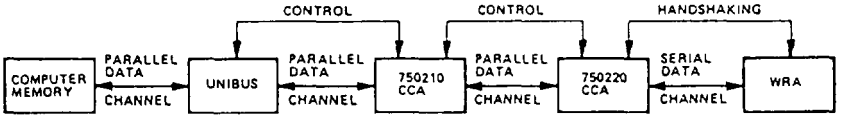


Figure 11-7.-High-speed communication channels block diagram.

provides a direct parallel data channel between the computer memory and HSCC No. 2 card. The HSCC No. 2 card converts the parallel data to a special serial format and generates all handshaking signals to the WRA. Data transfers may take place in either direction as a unibus read or write.

Stimulus and Loads

The stimulus and loads subsystem provides stimulus and loads to the UUT necessary to simulate actual operating conditions. Operating power for UUT operation is provided via programmable power supplies, which are accessible at the high current power distribution unit. The following units make up the stimulus and loads subsystem. Refer to figure 11-8 for the RSTS block diagram.

Signal Generators No. 1, 2A1A4, and No. 2, 2A1A6, receive a 10-MHz signal from the distribution amplifier. The generators provide the station, via the radio frequency interface unit, with a programmable signal source with a range of 50 MHz to 18 GHz.

Signal Generator 2A2A4, No. 3, receives a 10-MHZ signal from the distribution amplifier, then provides the station, via the RF interface unit, with a programmable signal source with a range of 10 kHz to 1280 MHz.

The programmable Vector Voltmeter 2A1A3 measures both phase angle and magnitude of complex ac signals and vector components with respect to a reference input. The unit can be operated in an automatic or manual mode. In the automatic mode, signals are routed to the vector voltmeter via the RF interface unit. In the manual mode, signal inputs are obtained by manually probing.

The programmable Digital Multimeter 3A1A2 provides resistance and voltage measurement capabilities to the station during UUT testing and maintenance self-testing via the RIU.

The programmable Signal Analyzer 3A1A6 is a microprocessor-based, dual-channel waveform, digitizing instrument that samples, digitizes, and stores waveforms of frequencies up to 200 MHz. The signal analyzer interfaces with the reconfigured interface unit and X-Y display.

The X-Y Display 3A2A2A1 is a solid-state directed-beam monitor that provides a video display of X, Y, and Z input information. It works in conjunction with the signal analyzer.

Measurement

The programmable Counter/Timer 3A1A7 provides the following measurement capabilities:

- Time interval averaging
- Period
- Period average
- Frequency (0 -100 MHz)
- Frequency ratio
- Frequency totalize
- Arithmetic functions
- Two channels are available at the RIU.

The Video Monitor 3A2A2A2 provides a high resolution display of video input signals received from the RIU during UUT testing. The programmable noise monitor 1A2 provides the station with noise (figure and gain) measurement and monitoring capabilities during UUT testing via the RF interface unit.

The Phase Modulation Noise Test Unit 1A6, in conjunction with the spectrum analyzer and via the RF interface unit, measures self-generated or inherent noise level signals from the UUT.

The dual channel RF Power Meter 2A2A1M1 measures RF power levels. All measurement modes are programmable, including zeroing, calibration, and cal factor. The RF power meter interfaces with the UUT through the reconfigurable interface unit or the RF interface unit.

Switching

To be able to test a variety of different types of UUT, the test station must be capable of switching various stimulus signals to the appropriate UUT input pins and switching various signals available at the UUT output pins to any of a variety of measurement devices for analysis. The following units describe how the test station accomplishes the switching of signals.

The High-Current Power Distribution Unit 2A2A2 (in addition to routing facility power to the UUT) provides switching for five high-current power supplies.

The Radio Frequency Interface Unit 2A1A2 is divided into four functional sections: signal generation, signal routing, signal attenuation/termination, and microprocessor control. The signal sections generate, attenuate, terminate, and route signals for measurement

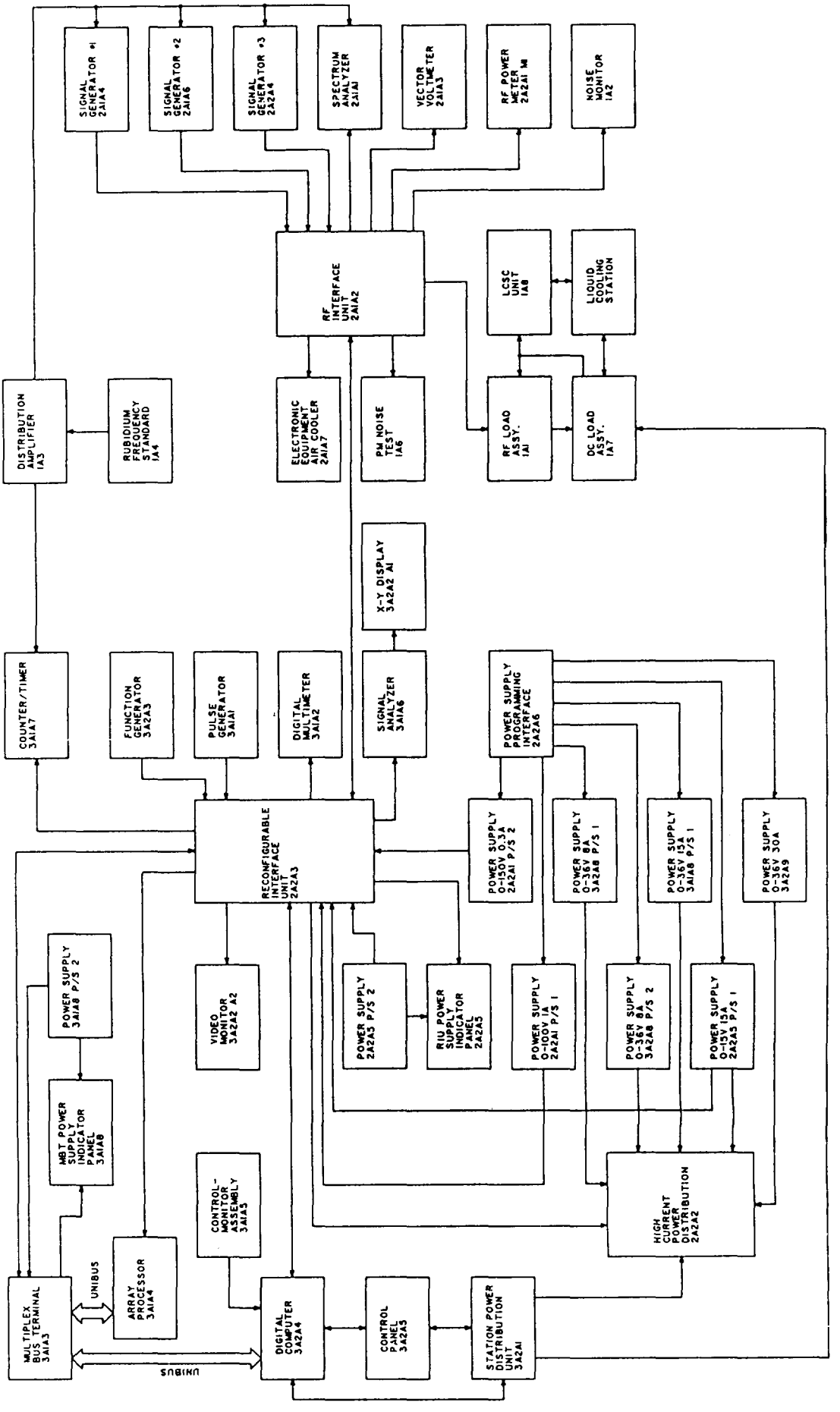


Figure 11-8.-Radar Set Test Station AN/APM-446 block diagram.

and excitation. The microprocessor section provides interface to the IEEE-488 bus and accepts commands to control switching paths in the three functional signal sections, as well as for phase modulator test drawer and electronic equipment air cooler.

The Recofigurable Interface Unit 2A2A3 is the primary interface between the UUT and test system input/output signal functions. It provides all of the test station analog, digital, and MIL-STD-1553B test capabilities required for testing WRA and SRA. A UUT is connected to the station using an interface device. This device is designed to mate with the patch panel, which is located on the front of the reconfigurable interface unit assembly. The interface's specific configuration can be altered by reassigning the card slots within the assembly. Under control of a dedicated processor, the LSI 11/02, the interface unit interfaces the patch panel pins with the station instrumentation, providing a very flexible I/O port. The patch panel has 1,008 single conductor pins. In addition, the interface unit provides 54 coax patch pins for routing of higher frequency signals that might be degraded by passing through the single conductor pins. The reconfigurable interface unit interfaces with nearly all station equipment concerned with analysis and measurement.

Cooling

The Electronic Equipment Air Cooler 2A1A7 supplies blower air to the UUT. It is program activated, via the RF interface unit, to provide cooling air at a rate of up to 150 cfm. A computer “UUT air cooling failure” interrupt is provided when an over/under airflow or overtemperature is sensed. Failure will cause stopping of UUT diagnostics, UUT power-down, and a “UUT air cooling failure” message to be displayed.

The liquid cooling system control unit is the interface between the liquid cooling station and radar transmitter WT. Liquid coolant is also supplied by the liquid cooling system control unit to the dc load and RF load assemblies.

APPENDIX I

GLOSSARY

ABSORPTION–Loss of energy that is turned into heat.

ABSORPTION FREQUENCY METER (WAVEMETER)–A frequency-measuring device incorporating a variable-tuned circuit that absorbs a small portion of the radiated energy under measurement.

ACCESS TIME–In computers, the time interval between the calling for information from a computer unit and the instant that such information is delivered.

ACCUMULATOR–A computer unit wherein numbers are accumulated. Usually an accumulator holds one number in storage; when a second number is entered, the accumulator adds the two numbers and retains the sum in storage.

ACOUSTIC–Pertaining to sound or the study of sound.

ACTIVE SONAR–An apparatus that radiates and receives information from returning echoes.

ADDER–An electronic circuit capable of providing the sum of two numbers entered therein.

ADDRESS–In computers, an identifying number or numbers or a particular group of symbols that identifies a particular storage location.

ADF–Automatic direction finding. An automatic radio compass that automatically aims a directional antenna to show the direction of the location of a transmitter. The ADF is normally used for homing purposes, but it can be used in conjunction with the magnetic compass to provide line-of-position information.

ADP–Acoustic data processor.

AGM–Air-launched, surface attact, guided missile.

AIM–Air-launched, aerial intercept, guided missile.

AMBIENT CONDITIONS–Physical conditions of the immediate environment; may pertain to temperature, humidity, pressure, etc.

AMBIENT NOISE–The naturally occurring noise in the sea and the noise resulting from man's activity, but excluding self-noise and reverberation.

ANALOG COMPUTER–A type of computer that provides a continuous solution of a mathematical problem with continuously changing inputs. Inputs and outputs are represented by physical quantities that may be easily generated or controlled.

AND GATE–A logic circuit having multiple inputs and a single output, so designed that the output is energized when (and only when) every input is in the prescribed signal state.

ANTENNA–Also aerial. A conductor or system of conductors that radiates or intercepts energy in the form of electromagnetic waves.

ANTIJAMMING–A function of a radar set to reduce or eliminate enemy jamming of electromagnetic waves, which hinder the usefulness of specific segments of the radio spectrum.

A-SCAN (A-DISPLAY)–In radar, a display in which targets appear as vertical displacements from a line representing the time base. Target distance is represented by the horizontal distance from one end of the time base. Amplitude of the vertical deflection is a function of the signal intensity.

ASW–Antisubmarine warfare. Operations conducted against submarines, their supporting forces, and bases.

ASWOC–ASW operations center.

ASYMMETRIC–Not symmetrical; without symmetry.

AVB–Avionic Bulletin.

AVC–Avionic Change.

AZIMUTH–Angular position or bearing in a horizontal plane, usually measured clockwise from true north. Azimuth and bearing are often used synonymously.

BALLISTICS–The term that refers to the science of the motion of projectiles or bombs.

BAND–The radio frequencies existing between two definite limits and used for a definite purpose; for example, standard broadcast band extending from 550 to 1600 kHz.

BANDWIDTH–The total frequency width of a channel or band of frequencies.

BATHYTHERMOGRAPH–A recording thermometer for obtaining a permanent graphical record of water temperature in degrees Fahrenheit at different water depths, in feet, as it is lowered or dropped into the ocean.

BEACON–Compared to a lighthouse. A radio or radar signal station that provides navigation and interrogation information for ships and aircraft.

BEAMWIDTH–The width of an electromagnetic beam, measured in degrees on an arc that lies in a plane along the axis of propagation, between points of equal field strength. It may be measured in the horizontal or vertical plane.

BEARING–The angular position of an object with respect to a reference point or line. If the reference point is true north, the bearing is the true bearing; if the reference is NOT true north, then the bearing is a relative bearing. If magnetic north (vice true north) is used as the reference, the bearing then becomes a magnetic bearing. Also, the direction of the line of sight, from a radar antenna to a target, measured in degrees. *See also* AZIMUTH.

BIAS–In vacuum tubes, the difference of potential between the control grid and the cathode; in transistors, the difference of potential between the base and emitter and between the base and collector; in magnetic amplifiers, the level of flux density in the core under no-signal conditions.

BIDIRECTIONAL COUPLER–A waveguide device having two outputs, which sample and present a signal atone output that is largely a function of the wave traveling in one direction, while the signal at the other output is largely a function of the wave traveling in the opposite direction.

BLACKBODY–An ideal body that absorbs all incident light, and therefore appears perfectly black at all wavelengths. The radiation emitted from such a body when it is hot is called "blackbody" radiation. The spectral energy density of blackbody radiation is the theoretical maximum for a body in thermal equilibrium.

BLANKING–The process of applying negative voltage to the control grid of the cathode-ray tube to cut off the electron beam during the retrace or flyback period.

BOLOMETER–A small resistive element used in the measurement of low and medium RF power. It is characterized by a large temperature coefficient of resistance that is capable of being properly matched to a transmission line.

BOTTOM BOUNCE–That form of sonar sound transmission in which sound rays strike the ocean bottom in deep water at steep angles and are reflected back to the surface and returned, which allows the obtaining of target information at long distances.

BRIDGE CIRCUIT–The electrical bridge circuit is a term referring to any one of a variety of electric circuit networks, one branch of which, the "bridge" proper, connects two points of equal potential, and hence carries no current when the circuit is properly adjusted or balanced.

B-SCAN (B-DISPLAY)–In radar, a rectangular display in which targets appear as illuminated areas, with bearing indicated by the horizontal coordinate and distance by the vertical coordinate.

CAGING (GYRO)–The act of holding a gyro so that it cannot precess and change its attitude with respect to the body containing it.

CAVITATION–The formation of local cavities (bubbles) in a liquid as a result of the reduction of total pressure. This pressure reduction may result from a negative pressure produced by rarefaction or from the reduction of pressure by hydrodynamic flow, such as is produced by high-speed movement of an underwater propeller.

CAVITY RESONATOR–A hollow, metallic cavity in which electromagnetic oscillation can exist when the cavity is properly excited.

CCTV–Closed circuit television. The application of television where reception is limited by broadcasting on specific frequencies and/or by connecting the receivers directly to the television camera via coaxial cables.

CHARACTERISTIC (ITERATIVE) IMPEDANCE–The apparent load presented to a source; in electronics, the characteristic impedance at any frequency range is approximately equal to the ratio of the inductance to the capacitance.

CLEARING PULSE–In computers, a pulse that is employed for clearing or resetting a circuit to its predetermined initial state.

COMPARATOR–A circuit that compares two signals or values, and indicates agreement or variance between them.

COMPOSITE VIDEO–The total video signal, consisting of picture information, blanking pulses, and sync pulses.

COMPRESSION–In wave motion, the forcing together of the medium's molecules. *See also* RAREFACTION.

COMPUTER–A mechanism or device that performs mathematical operations. *See also* ANALOG COMPUTER *and* DIGITAL COMPUTER.

COMPUTER CODE (ALSO CALLED A COMPUTER LANGUAGE)–The code by which data are represented within a computer system; for example, binary coded decimal.

COMPUTER PROGRAM–A series of instructions or statements prepared in a form acceptable to the computer.

CONTROL CIRCUITS–In computers, those circuits involved in the carrying out of the program instructions.

COUNTERMEASURES–Devices and/or techniques intended to impair the operational effectiveness of enemy activity.

COUNTING CIRCUIT–A circuit that receives uniform pulses representing units to be counted and produces a voltage in proportion to their frequency.

CRT–Cathode-ray tube.

DC RESTORER–A circuit used to reinsert the dc component of the video signal lost during amplification.

DEGREES OF FREEDOM (GYRO)–A term applied to gyros to describe the number of variable angles required to specify the position of the rotor spin axis relative to the case.

DETECTORS, INFRARED–Thermal devices for observing and measuring infrared radiation, such as the bolometer, radiomicrometer, thermopile, pneumatic cell, photocell, photographic plate, and photoconductive cell.

DIFAR–Directional frequency analyzing and recording. An ASW technique used in pinpointing submerged contacts.

DIFFERENTIAL–A mechanical computing device used to add or subtract two quantities.

DIFFUSION–The spreading out of energy or particles from a high concentration to a low concentration, due to random velocity and scattering.

DIGITAL COMPUTER–A type of computer in which quantities are represented in numerical form and that is generally made to solve complex mathematical problems by use of the fundamental processes of addition, subtraction, multiplication, and division. Its accuracy is limited only by the number of significant figures provided.

DIPPING SONAR–Used by helicopters. Lowered from the helicopter for searching and retracted for flight.

DIRECTIONAL COUPLER–A device used to extract a portion of the RF energy moving in a given direction in a transmission line or waveguide. Energy moving in the opposite direction is rejected. *See also* BIDIRECTIONAL COUPLER.

DIRECTION FINDER (DF)–VHF/UHF navigation aid operated by personnel on the ground to furnish azimuth information to aircraft.

DISCRIMINATOR–A dual-input circuit in which the output is dependent on the variation of one input from the other input or from an applied standard.

DISTORTION–The production of an output waveform that is not a true reproduction of the input waveform. Distortion may consist of irregularities in amplitude, frequency, phase, etc.

DIURNAL–Having a recurring daily cycle.

DIVERGENCE–Energy loss caused by spreading in all directions.

DOPPLER EFFECT–An apparent change in the frequency of a sound wave or electromagnetic wave reaching a receiver when there is relative motion between the source and the receiver.

DRIFT–Net change in characteristics of electronic components or parameters, resulting from external or incidental conditions.

DRUM–In computers, a cylinder coated with a material capable of being magnetized so that it can be employed for the retention of information in storage functions.

DUPLEXER–A switch or tube that permits the use of a single antenna for both transmission and reception. The dual function of the duplexer is to prevent absorption of transmitter energy by the receiver system (thereby protecting the receiver) and to prevent absorption of any appreciable portion of the received echo signal by the transmitter.

ECHO–That portion of the energy reflected to the receiver from the target.

ECHO BOX–A high-Q resonant cavity used with microwave radar sets to provide artificial targets for radar testing and for tuning the receiver to the transmitter. The echo box stores RF energy during the transmitted-pulse interval, and reradiates it through the same antenna for a short time following the pulse.

ECM–Electronic countermeasures. The means by which enemy electronic devices are nullified and, at the same time, intelligence is gathered concerning the nature of the enemy radiations. ACTIVE ECM implies jamming/deceptive techniques to degrade enemy equipment or operator functions. PASSIVE ECM entails the use of receiving (only) equipment to detect, locate, analyze, and evaluate enemy radiations and radio emissions.

ELECTRONIC SWITCH–A circuit that causes a start and stop action or a switching action by electronic means.

ELECTROSTRICTION–That property of certain ceramic materials that, after having a permanent operating bias established, causes these materials to vary slightly in length when they are placed in an electric field.

EQUIVALENT CIRCUIT–A diagrammatic arrangement of component parts, representing in simplified form the effects of a more complicated circuit, to permit easier analysis.

ERASING HEAD–A device that removes stored data from the surface of a magnetic storage material.

ESM–Electronic warfare support measures. Concerns electronic emissions and countermeasures.

E-TRANSFORMER–A magnetic device with an E configuration, used as an error detector.

EW–Electronic warfare. Tactical use of electronics to prevent or reduce the enemy's effective use of radiated electromagnetic energy, and the actions taken to assure the effective use of ours. *See also* ECM.

FEEDBACK–The return of a portion of the output of a circuit stage to the input of that stage or a preceding stage, such that there is either an increase (regeneration) or a reduction (degeneration) in amplification, depending on the relative phase of the returned signal with the input.

FERRITE–A hard and brittle crystalline substance made from a mixture of powdered materials, including iron oxides; it has special magnetic properties of particular value in computers and in many other applications.

FIDELITY–The extent to which a system, or a portion of a system, accurately reproduces at its output the essential characteristics of the signal that is impressed upon its input.

FLIR–Foward Looking InfraRed system.

FREE GYRO–A gyro so gimbaled that it can assume and maintain any attitude in space. A free gyro has two degrees of freedom; torque cannot be applied to the rotor of a truly free gyro.

FREQUENCY–The number of hertz (cycles per second) of an alternating current.

FULL ADDER–An adder circuit that can complete the adding procedure involving the carry process, as distinguished from the half adder, which is not capable of accepting a previous carry.

GATING CIRCUIT (GATE)–A circuit used to activate (or deactivate) another circuit by permitting (or prohibiting) operation during selected periods of time.

GIMBAL–A frame in which the gyro wheel spins and that allows the gyro wheel to have certain freedom of movement. It permits the gyro rotor to incline freely and retain that position when the support is tipped or repositioned.

GRADIENT–The nature of the Sound-transmission curve (negative, positive, isothermal, etc.) as used in sonar applications. *See also* ISOTHERM *and* THERMOCLINE.

GRADIENT, NEGATIVE–When the temperature of the water decreases with depth, it has a negative temperature gradient.

GRADIENT, POSITIVE–When the temperature of the water increases with depth, it has a positive temperature gradient.

GYROSCOPES–A wheel or disk so mounted as to spin rapidly about one axis and be free to move about one or both of the two axes mutually perpendicular to the axis of spin.

HALF ADDER–A partial adding circuit that is not capable of accepting a previous carry. It must be combined with another half adder and a circuit

capable of performing the carry function to form a full adder.

HERO–Hazardous electromagnetic radiation to ordnance.

HERTZ–A unit of frequency equal to 1 cycle per second.

HETERODYNE–To mix two alternating currents of different frequencies in the same circuit; they are alternately additive and subtractive, thus producing two beat frequencies, which are the sum of, and difference between, the two original frequencies.

HORIZONTAL PLANE–A horizontal plane is tangent to the surface of the earth. Visualize this condition by laying a playing card on an orange. The card represents the horizontal plane; the orange symbolizes the earth; and the point of contact between the two is the point of tangency. Every plane parallel to the horizontal plane is likewise a horizontal plane.

HYDROPHONE–An acoustic device that receives and converts underwater sound energy into electrical energy.

HYSTERESIS–A lagging of the magnetic flux in a magnetic material behind the magnetizing force that is producing it.

INFRARED–Invisible waves in that portion of the electromagnetic spectrum lying between visible light and radio frequencies, and having a penetrating heating effect.

INHIBITORY PULSE– pulse that acts to inhibit or suppress another signal from going through a logic circuit and appearing at the output.

INPUT-OUTPUT EQUIPMENT–A device that provides the means of communication between the computer and external equipment. The device accepts new data, sends it into the computer for processing, receives the results, and transforms the data into usable form. In many cases it is also referred to as peripheral equipment.

INSTRUCTION–In computer programming, a set of identifying characters or a computer "word" that is designed to cause the computer to perform specific operations.

INTEGRATING CIRCUIT–A circuit whose output voltage is proportional to the product of the instantaneous applied input voltages and their durations.

INTEGRATOR–A computing device used for summing up an infinite number of minute quantities.

INTELLIGENCE–The message or information conveyed, as by a modulated radio wave.

INTERFACE–A concept involving the specification of the interconnection between two equipments or systems. The specifications include the type, quantity, and function of signals to be interchanged via those circuits. A device that converts or translates any type of information from one given medium into signals of another given medium; for example, electrical signals to fluidic signals, fluidic signals to electronic signals, etc.

IR–InfraRed

ISOTHERM–A line connecting points of equal temperature.

ISOTHERMAL LAYER–A layer of water in which there is no appreciable change of temperature with depth.

ISOVELOCITY LAYER–A layer of water in which there is no appreciable change of sound velocity with depth.

KINEMATIC LEAD–The lead required to score a hit on a specified target due to relative motion between target and gun platform.

KNEE (OF A CURVE)–An abrupt change in direction between two fairly straight segments of a curve.

LAYER DEPTH–The depth from the surface of the sea to the top of the first significant negative thermocline.

LAYER EFFECT–Partial protection from echo ranging and listening detection when below layer depth.

LOGIC CIRCUITS–Digital computer circuits used to store information signals and/or to perform logical operations on those signals.

LOOP ANTENNA–One or more complete turns of wire used with a radio receiver. Also used with direction-finding equipment.

LOS–Line of sight. The straight-line distance from ship to horizon. Represents radio and radar VHF and UHF transmission range limits under normal conditions.

MAD–Magnetic anomaly detection. The detection of slight distortions in the earth's magnetic field. In the U.S. Navy, it is used exclusively by aircraft.

MAGNETIC FIELD–The region in space in which a magnetic force exists, caused by a permanent magnet or as a result of current flowing in a conductor.

MAGNETOSTRICTION–That property of certain ferro-type materials that causes them to vary slightly in length when they are in an alternating magnetic field.

MAGNETRON–A microwave oscillator that uses an electron tube (consisting of a cathode and an anode), a strong axial magnetic field, and resonant cavities.

MAGNETRON ARCING–Internal breakdown between cathode and anode of a magnetron, usually resulting from presence of gas. Occurs during the breaking-in or "seasoning" period and again at the end of the useful life. Occasional arcing is common, especially in high-power magnetrons.

MAGNETRON PULLING–The frequency shift of a magnetron resulting from a mismatch at the output. It is caused by such factors as faulty rotating joints, reflections from objects near the antenna, etc.

MAGNETRON PUSHING–The frequency shift of a magnetron resulting from faulty operation of the modulator. It may result from an improperly shaped pulse or from interaction of the pulse with the magnetic field.

MASTER CLOCK–The timed and synchronized generators that comprise the source and time reference for computer signals.

MEMORY UNIT–In computers, a device used for storing data for possible use in computation.

MICROFICHE–A film negative card (fiche) developed for many purposes throughout the Navy wherever microfilming is used to reduce amounts of paper documents.

MICROMETER–A unit of length equal to 10^{-6} meter. Formerly a micron.

MICRON– *See* MICROMETER.

MICROWAVES–A term commonly used to indicate electromagnetic waves in the frequency range between 1,000 and 300,000 megahertz (30 cm to 1 mm).

MILLIAMMETER–An ammeter that measures current in thousandths of an ampere.

MODULATION–The process of varying the amplitude or frequency of a carrier wave in accordance with other signals to convey intelligence. The modulating signal may bean audiofrequency signal, a video signal (as in television), or even electrical pulses or tones to operate relays.

MODULE–In electronic terminology, a group or cluster of circuits/components usually mounted together on a "board" or "potted" together in a lump.

MONOPULSE–A method of antenna lobing that permits information to be obtained on target range, bearing, and elevation from a single pulse (as distinguished from sequential lobing).

NOISE–Any undesired disturbance within the useful frequency band; also, that part of the modulation of a received signal (or an electrical or electronic signal within a circuit) representing an undesirable effect of transient conditions.

NOT CIRCUIT–In computers, a circuit in which the output signal does not have the same polarity as the input signal. A phase inverter.

NULL–A point or position where a variable-strength signal is at its minimum value (or zero).

OFF-LINE EQUIPMENT–Peripheral computer equipment that can operate independently of the main computer for such operations as transcribing punch card information to magnetic tape, or magnetic tape to printed form.

OMNIDIRECTIONAL–Going out in all directions, as the radiation pattern of a single dipole antenna.

ON-LINE EQUIPMENT–Computer equipment, due to configuration or design, that requires the use of the central processing unit of the computer.

OR GATE–A logic circuit having multiple inputs and a single output, so designed that the output is energized when any one or more of the inputs are in the prescribed signal state.

PARALLEL MODE–In computer operation, the handling of a group of numbers or other symbols simultaneously.

PARAMETERS–In electronics, the design or operating characteristics of a circuit or device.

PASSIVE SONAR–An apparatus that receives energy generated from another source.

PERIPHERAL EQUIPMENT–Either on-line or off-line auxiliary equipment supporting the operations, but is not a part of the computer itself.

These machines may consist of card readers, card punches, magnetic tape feeds, and high-speed printers.

PHOTON–A quantum of electromagnetic energy. The equation hv, where h is Plank's constant and v is the frequency associated with the photon,

PICKOFF–In gyros, a sensing device that measures the angle of the spin axis with respect to its reference, and provides an error signal that indicates the direction and (in most cases) the magnitude of the displacement.

PIEZOELECTRIC EFFECT–Effect of producing a voltage by placing stress, either by compression, expansion, or twisting, on a crystal and conversely, producing a stress in a crystal by applying a voltage to it.

PIPS–Popular term for bright spots on a CRT display such as a radar or sonar screen.

POLARIZATION–In electronics, a term used in specifying the direction of the electric vector in a linearly polarized electromagnetic wave as radiated from a transmitting antenna, or as picked up by a receiving antenna.

POTENTIOMETER–A variable voltage divider; a resistor that has a variable contact arm so that any portion of the potential applied between its ends may be selected

PPI SCAN (PPI DISPLAY)–A cathode-ray tube presentation in which the signal appears on a rotating radial line. Distance is indicated radially, and bearing as an angle.

PRECESSION–The reaction of a gyro to an applied torque, which causes the gyro to tilt itself at right angles to the direction of the applied torque in such a manner that the direction of spin of the gyro rotor will be in the same direction as the applied torque.

PROGRAM–A complete plan for the solution of a problem, including the complete sequence of machine instructions and routines necessary to solve the problem by an electronic computer.

PROPAGATION–Extending the action of, transmitting, carrying forward as in space or time or through a medium (as the propagation of sound, light, or radio waves).

PSEUDO–Term meaning false or fake.

PULSE–A momentary sharp surge of electrical voltage or current.

PULSE DURATION–The time interval between the leading and trailing edges of each of a particular group of pulses; the instantaneous values of these are often used in a specific relation to the peak pulse amplitude to determine power output.

PULSE INTERVAL–The time interval between the leading edges of successive pulses in a sequence.

PULSE SEPARATION–The time interval between the trailing edge of one pulse and the leading edge of the next pulse.

PULSE TRAIN–A series of pulses passed through a circuit as control or information signals.

RADIAN–In a circle, the angle included within an arc equal to the radius of the circle. A complete circle contains 2! radians. One radian equals 57.3 degrees and 1 degree equals 0.01745 radian.

RANGE–The distance of an object from an observer.

RAREFACTION–In wave motion, when the vibration is inward, a rarefaction or region of reduced pressure is produced.

RASTER–The illuminated rectangular area scanned by the electron beam in a picture tube/CRT.

RATE GYRO–A gyro with one-degree-of-freedom, which has an elastic restraint, with or without a damper, and whose output will be proportional to the rate of the applied torque.

REFLECTION, SOUND–Sound rays transmitted in the sea eventually reach either the surface or the bottom. Since these boundaries are abrupt and very different in sound transmitting properties from the water, sound energy along a ray path striking these boundaries will be returned (reflected) to the water.

REFRACTION, SOUND–The bending or curving of a sound ray that results when the ray passes from a region of one sound velocity to a region of a different velocity. The amount of ray bending depends on the amount of difference between sound velocities.

REGISTER–A specific computer unit that stores a single computer word.

RELATIVE BEARING–A bearing taken when the heading of a ship serves as the reference line. See also BEARING.

RELATIVE MOTION–The apparent movement of an object in relation to another object.

RESONANT CAVITY–A space, normally enclosed by an electrically conductive surface, in which oscillatory electromagnetic energy is stored, and whose resonant frequency is determined primarily by the geometry of the enclosure.

REVERBERATION–A succession of echoes caused by reflections of sounds. In the ocean it is caused by irregularities in the ocean bottom, surface, and suspended matter (as fish). Under these conditions, an emitted pulse maybe received as a muffed echo due to sound interference.

RHEOSTAT–A variable resistor that has one fixed terminal and a moveable contact. Potentiometers may be used as rheostats, but a rheostat cannot be used as a potentiometer because connections cannot be made to both ends of the resistance element.

RIGIDITY–In gyros, the characteristics of a spinning body that causes it to oppose all attempts to tilt it away from the axis in which it is spinning.

RING TIME–In radar, the time during which the output of an echo box remains above a predetermined level; used in measuring the performance of radar equipment.

SAR–Search and rescue.

SCALE FACTOR–A quantity used to introduce a change according to a fixed ratio or scale; a proportionality constant.

SCANNING SONAR–Sonar that transmits sound pulses in all directions simultaneously.

SCATTERING–Reflection losses from particles suspended in the water.

SENSO–Sensor operator (SO). Operates the ASW platforms acoustic and nonacoustic sensor systems.

SENSOR–A component that senses variables and produces a signal therefrom. Temperature, sound, heat, and light sensors are some examples.

SEQUENTIAL LOBING–Successively shifting the radar beam about the scanner centerline through a particular pattern; differs from monopulse.

SERIAL OPERATION–In computers, the sequential handling of a group of numbers or symbols.

SHIFT REGISTER–In computers, a circuit that will shift a digit or a group of digits either to the left or to the right; it is of particular importance in some multiplication and division processes, and in sequential storage of pulse trains.

SHOT EFFECT–Noise voltages developed as a result of the random nature of electron flow in vacuum tubes, or the random flow of either primary or secondary carriers in transistors.

SLEW–To change the position of an indicator mark on a CRT display by varying the time relationship of the mark with respect to the start of the sweep.

SOFTWARE–Pertains to the programs and routines used with computers. The totality of programs and routines used to extend the capabilities of computers. In contrast to HARDWARE, which is the construction parts (mechanical, electrical, and electronic elements) of the computer.

SONAR–Acronym for SOund Navigation And Ranging. Apparatus or technique of obtaining information regarding objects or events underwater.

SONIC–Within the audible range of the human ear.

SONOBUOY–Small floating buoy with an attached hydrophone and a radio transmitter that relays underwater sounds picked up by the hydrophone to ASW units.

SONOBUOY RECEIVER SYSTEM (SRX)–An FM radio receiver system used exclusively for sonobuoy RF signal reception and processing.

SONOBUOY REFERENCE SYSTEM (SRS)–The system used to determine the position of deployed sonobuoys relative to aircraft position.

SOUND CHANNEL–Condition when two layers of water with near equal temperatures produce a sound channel. Sound between the two layers is refracted by the layers, stays between them, and travels for great distances.

SYNC–A short form of the word synchronizing, which means to cause two elements of a system to coincide in speed, frequency, relative position, or time.

TACCO–Tactical coordinator.

THERMAL NOISE–A very low-level noise produced by molecular movement in the sea.

THERMISTOR–A solid-state, semiconducting device whose resistance varies with temperature.

THERMISTOR–A bolometer characterized by a decrease of resistance as the temperature rises. *See also* BOLOMETER.

THERMOCLINE–The layer in the sea where the temperature decreases continuously with depth.

Usually the decrease (gradient) is greater than 2.7°F per 165 feet in depth.

TORQUE–A force tending to cause rotational motion; the product of the force applied times the distance from the force to the axis of rotation.

TRANSDUCER–A device that converts signals received in one medium into outputs in some other medium; for example, electrical inputs to fluidic outputs, or mechanical motion into electrical quantities.

TRIGGERING–Starting an action in another circuit, which then operates for a time under its own control.

TRUE BEARING–A bearing given in relation to true geographic north. *See also* BEARING.

TUMBLE (GYRO)–To subject a gyro to a torque so that it presents a precession violent enough to cause the gyro rotor to spin end over end.

VELOCITY–A vector quantity that includes both magnitude (speed) and direction in relation to a given frame of reference.

VERTICAL PLANE–A vertical plane is perpendicular to the horizontal plane, and is the reference from which bearings are measured. Relative bearing, for example, is measured in the horizontal plane clockwise from the vertical plane through own ship's centerline to the vertical plane through the line of sight. The system of planes makes possible the design and construction of mechanical and electronic equipment to solve the fire control problem. These lines and planes are imaginary extensions of some characteristic of the ship or target, or of the relation in space between them.

WAVEGUIDE–Metal tubes or dielectric cylinders capable of propagating electromagnetic waves through their interiors. The dimensions of these devices are determined by the frequency to be propagated. Metal guides are usually rectangular or circular in cross section; they maybe evacuated, air filled, or gas filled, and may or may not be pressurized. Dielectric guides consist of solid dielectric cylinders surrounded by air.

WAVELENGTH–The distance traveled by a wave during the time interval of one complete cycle. It is equal to the velocity divided by the frequency.

WAVE PROPAGATION–The radiation, as from an antenna, of RF energy into space, or of sound energy into a conducting medium.

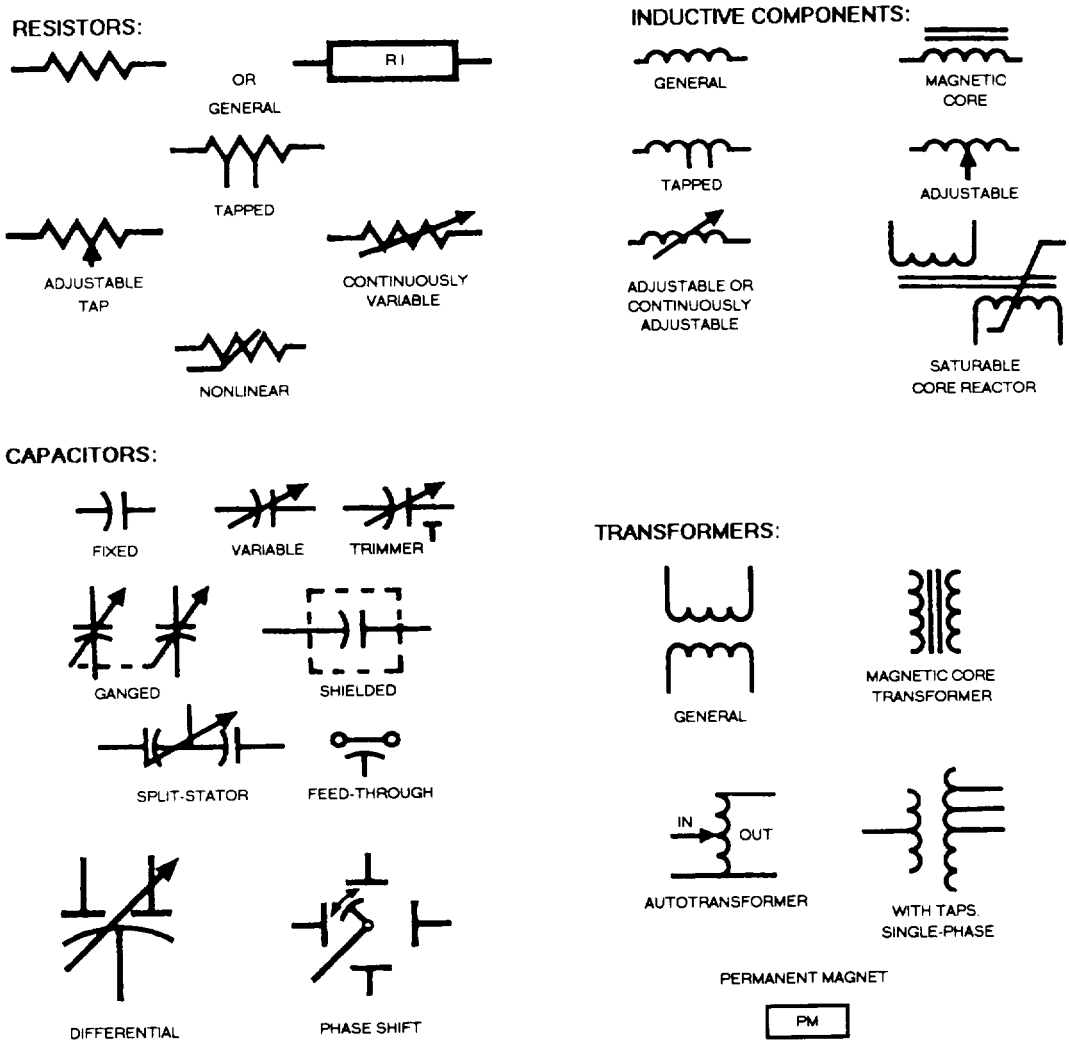
WORD–In computers, a particular number of characters handled as a unit by the computer and having a specific meaning with respect to the computation process.

APPENDIX II

SYMBOLS, FORMULAS, AND MEASUREMENTS

SYMBOLS

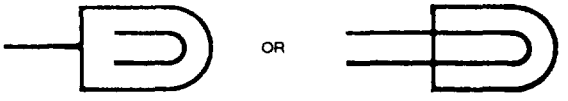
(SEE ANSI/IEEE STD Y32.2-1975 AND 315A-1986)



(WHEN CAPACITOR ELECTRODE INENTIFICATION IS NECESSARY, THE CURVED ELEMENT SHALL REPRESENT THE OUTSIDE ELECTRODE IN FIXED PAPER-DIELECTRIC AND CERAMIC-DI-ELECTRIC, THE NEGATIVE ELECTRODE IN ELECTROLYTIC CAPACITORS, THE MOVING ELEMENT IN VARIABLE AND ADJUSTABLE CAPACITORS, AND THE LOW POTENTIAL ELEMENT IN FEED-THROUGH CAPACITORS.)

NOTE: FOR FURTHER INFORMATION CONCERNING SYMBOLS REFER TO IEEE STANDARDS AND AMERICAN NATIONAL STANDARD GRAPHICS SYMBOLS FOR ELECTRICAL AND ELECTRONICS DIAGRAMS, ANSI Y32 2/IEEE NO. 315, WHICH HAS BEEN ADOPTED FOR MANDATORY USE BY THE DEPARTMENT OF DEFENSE

INDICATOR LAMP:



MICROPHONE:



CRYSTAL:

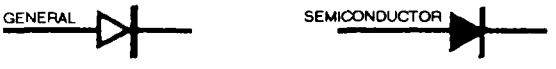


QUARTZ CRYSTAL;
PIEZOELECTRIC CRYSTAL
UNIT.

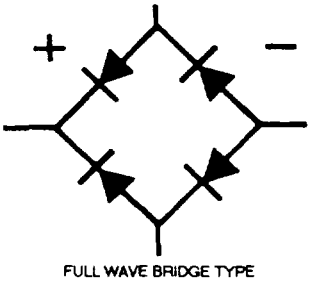
KEY:



RECTIFIER:

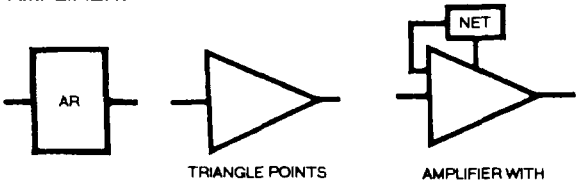


NORMAL CURRENT FLOW IS AGAINST THE ARROW



FULL WAVE BRIDGE TYPE

AMPLIFIER:

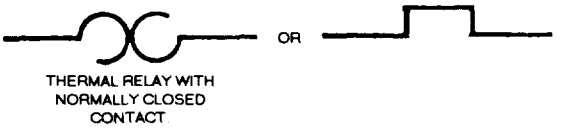


TRIANGLE POINTS
IN DIRECTION OF
TRANSMISSION
(SIGNAL FLOW)

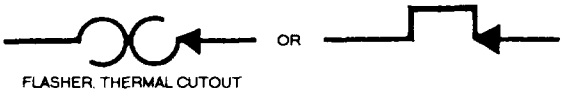
AMPLIFIER WITH
EXTERNAL
FEEDBACK PATH

BASIC SYMBOL INDICATES ANY METHOD OF
AMPLIFICATION EXCEPT THAT OPERATING ON
THE PRINCIPLE OF ROTATING MACHINERY.

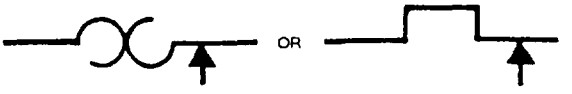
THERMAL ELEMENTS:

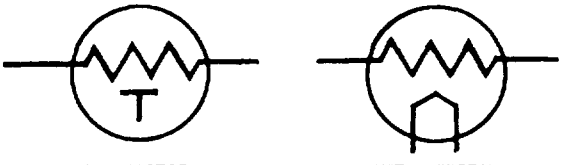


THERMAL RELAY WITH
NORMALLY CLOSED
CONTACT.



FLASHER, THERMAL CUTOUT





THERMISTOR

WITH INTEGRAL
HEATING ELEMENT



TEMPERATURE-MEASURING THERMOCOUPLE
(DISSIMILAR METAL DEVICE)

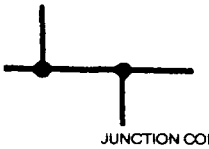
INPUTS (NONSTANDARD):



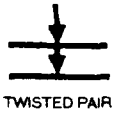
PATH, TRANSMISSION:



CROSSING NOT
CONNECTED



JUNCTION CONNECTED



TWISTED PAIR

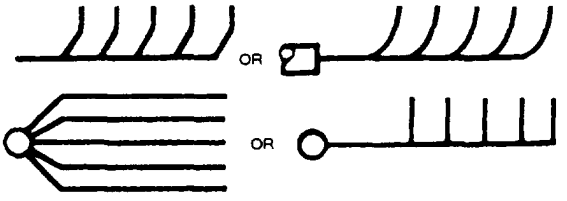


AIR OR SPACE PATH

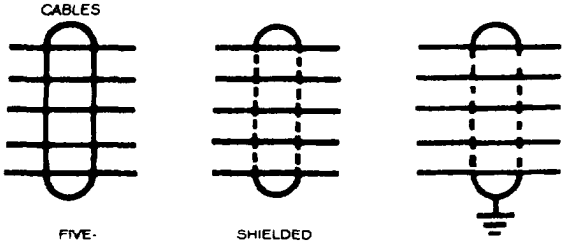


COAXIAL

GROUPING OF WIRES IN BUNDLES

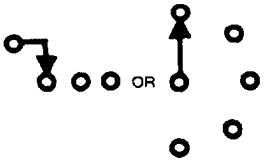


GROUPING OF WIRES IN CABLES



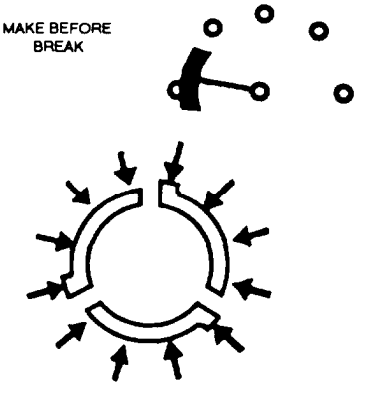
NUMBER OF CONDUCTORS MAY BE ONE OR MORE AS NECESSARY

SELECTOR SWITCHES



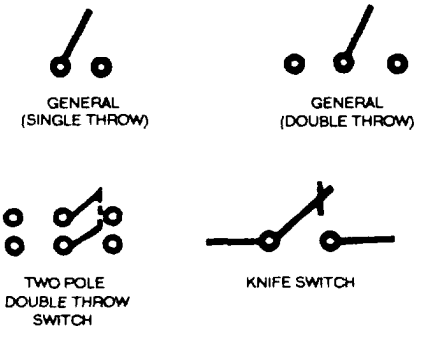
GENERAL

ANY NUMBER OF TRANSMISSION PATHS MAY BE SHOWN. ALSO BREAK BEFORE MAKE SWITCH

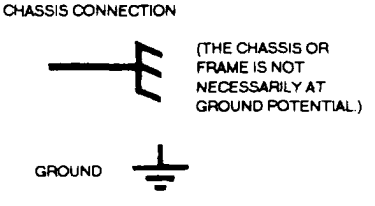


WAFER, TYPICAL 3-POLE, 3-CIRCUIT SWITCH. VIEWED FROM END OPPOSITE CONTROL KNOB. FOR MORE THAN ONE SECTION, #1 IS NEAREST CONTROL KNOB.

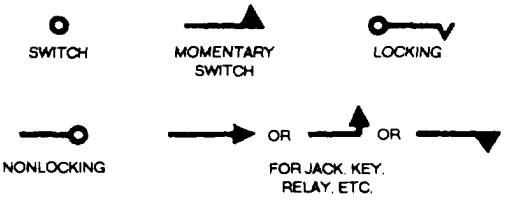
SWITCHES:



CIRCUIT RETURNS:



CONTACTS (ELECTRICAL):

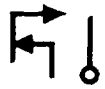


CONTACTS (ELECTRICAL) (CONTINUED):

CONTACT ASSEMBLIES



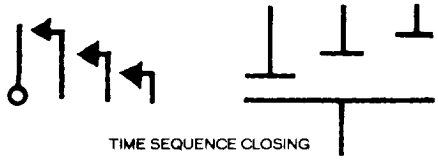
CLOSED CONTACT
(BREAK)



MAKE BEFORE
BREAK

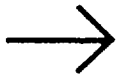


OPEN CONTACT
(MAKE)

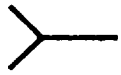


TIME SEQUENCE CLOSING

DISCONNECTING DEVICES:



MALE
(PIN CONTACT)



FEMALE
(SOCKET CONTACT)



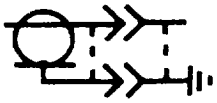
ENGAGED
(PIN-TO-SOCKET)



COAXIAL
(MALE)



COAXIAL CONNECTORS
MATED

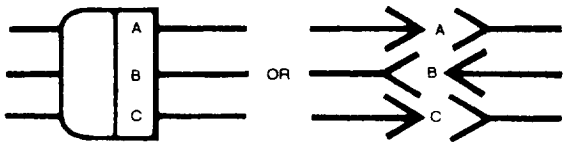


COAXIAL CONNECTED
TO SINGLE CONDUCTOR

THE CONNECTOR SYMBOL IS NOT AN ARROWHEAD. IT IS LARGER AND THE LINES ARE DRAWN AT A 90° ANGLE.



SPLICE



CONNECTOR ASSEMBLY (GENERAL)

ELECTRON TUBES :

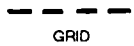
COMPONENT TUBE SYMBOLS



DIRECTLY-HEATED
(FILAMENTARY)
CATHODE



INDIRECTLY-HEATED
CATHODE



GRID



POOL CATHODE



COLD
CATHODE



ANODE OR
PLATE



PHOTOCATHODE



ENVELOPE (SHELL)



GAS FILLED
ENVELOPE

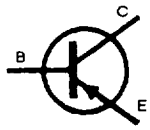


SPLIT ENVELOPE

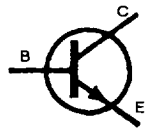
SEMICONDUCTOR DEVICES :



DIODE

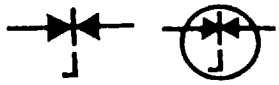


PNP

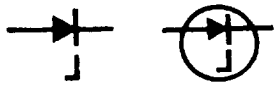


NPN

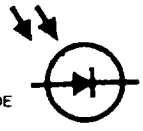
TRANSISTORS



BREAKDOWN DIODE,
BIDIRECTIONAL



BREAKDOWN DIODE,
UNIDIRECTIONAL
(ALSO BACKWARD DIODE)



PHOTODIODE

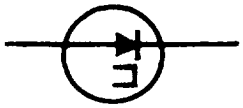
SEMICONDUCTOR DEVICES (CONTINUED):



TEMPERATURE DEPENDENT DIODE



PNPN SWITCH



TUNNEL DIODE

TYPICAL ELECTRON TUBES:



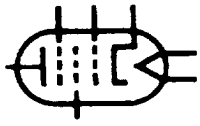
COLD CATHODE GAS TUBE



PHOTOTUBE SINGLE UNIT, VACUUM



DIODE



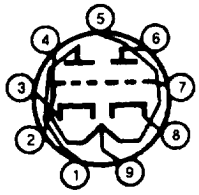
PENTODE



TWIN TRIODE ILLUSTRATING ELONGATED ENVELOPE

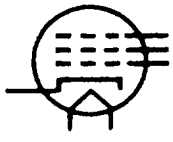


DIODE SHOWING BASE CONNECTIONS



TWIN TRIODE WITH TAPPED HEATER

TYPICAL CATHODE RAY TUBES:



MAGNETIC DEFLECTION

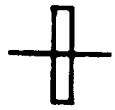


ELECTROSTATIC DEFLECTION

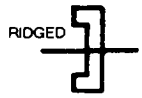
WAVEGUIDES:



CIRCULAR



RECTANGULAR



RIDGED

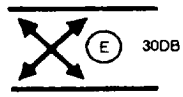


ROTARY JOINT

DIRECTIONAL COUPLERS



GENERAL



E PLANE APERTURE COUPLING, 30 DB TRANSMISSION LOSS

COUPLING METHODS:

GENERALLY USED FOR COAXIAL AND WAVEGUIDE TRANSMISSION.



COUPLING BY APERTURE WITH AN OPENING OF LESS THAN FULL WAVEGUIDE SIZE. TYPE OF COUPLING WILL BE INDICATED WITHIN CIRCLE (E, H, OR HE).



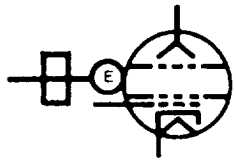
COUPLING BY LOOP TO SPACE

COUPLING BY LOOP TO GUIDED TRANSMISSION PATH

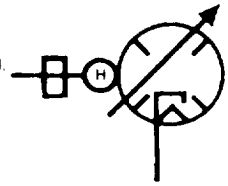


COUPLING BY PROBE FROM COAXIAL TO RECTANGULAR WAVEGUIDE WITH DIRECT-CURRENT GROUNDS CONNECTED

TYPICAL MAGNETRONS AND KLYSTRONS:

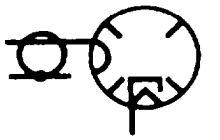


REFLEX KLYSTRON, APERATURE COUPLED

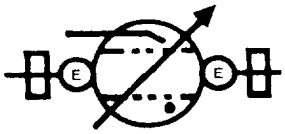


TUNABLE MAGNETRON, APERTURE COUPLED

TYPICAL MAGNETRONS AND KLYSTRONS (CONTINUED):



REASONANT TYPE WITH COAXIAL OUTPUT



TRANSMIT-RECEIVE (TR) TUBE GAS FILLED, TUNABLE INTEGRAL CAVITY, APERTURE COUPLED, WITH STARTER

ROTATING MACHINES:

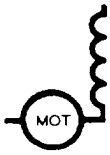


MOTOR

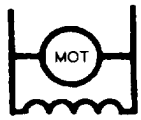


GENERATOR

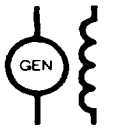
TYPES OF WINDINGS



SERIES



SHUNT



SEPARATELY EXCITED



DYNAMOTOR

WINDING SYMBOLS



SINGLE-PHASE



TWO-PHASE



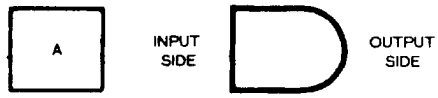
THREE-PHASE (WYE)

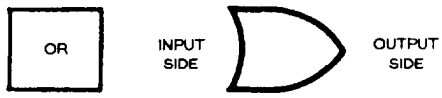


THREE-PHASE (DELTA)

LOGIC FUNCTION:

AND FUNCTION

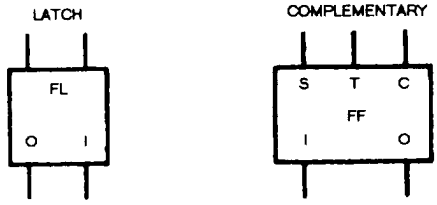




EXCLUSIVE OR FUNCTION



FLIP-FLOPS

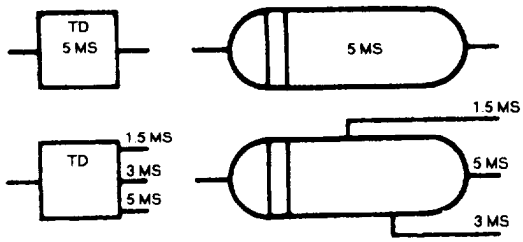


NEGATION

ELECTRIC INVERTER



TIME DELAY

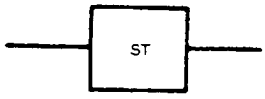


LOGIC FUNCTIONS (CONTINUED):

SINGLE SHOT



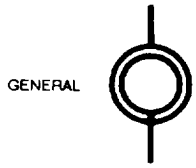
SCHMITT TRIGGER



OSCILLATOR



SYNCHROS:



A LETTER COMBINATION FROM THE FOLLOWING LIST MAY BE PLACED ADJACENT TO THE SYMBOL TO INDICATE THE TYPE OF SYNCHRO

- TX - TORQUE TRANSMITTER
- TDX - TORQUE DIFFERENTIAL TRANSMITTER
- CX - CONTROL TRANSMITTER
- CDX - CONTROL DIFFERENTIAL TRANSMITTER
- TR - TORQUE RECEIVER
- CT - CONTROL TRANSFORMER

SYNCHROS (CONTINUED):



TRANSMITTER, RECEIVER, OR CONTROL TRANSFORMER

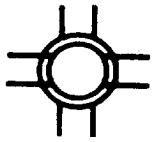


DIFFERENTIAL TRANSMITTER OR RECEIVER

RESOLVER (SYNCHRO)

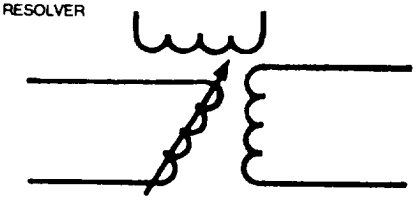


SINGLY-WOUND ROTOR

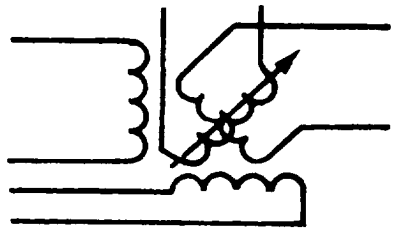


DOUBLY-WOUND ROTOR

RESOLVER



SINGLY-WOUND ROTOR



DOUBLY-WOUND ROTOR

PICKUP HEADS:



GENERAL



WRITING, RECORDING, HEAD, SOUND RECORDER



READING, PLAYBACK, HEAD, SOUND REPRODUCER



APPLICATION, WRITING, READING, AND ERASING



ERASING, ERASER, MAGNETIC

BATTERIES:



ONE CELL MULTICELL TAPPED MULTICELL

(LONG LINE IS ALWAYS POSITIVE)

CIRCUIT PROTECTORS:

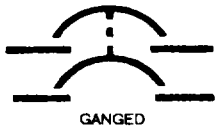
FUSE



CIRCUIT BREAKERS



SWITCH

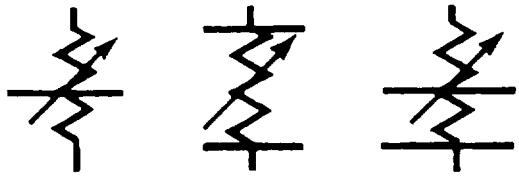


GANGED



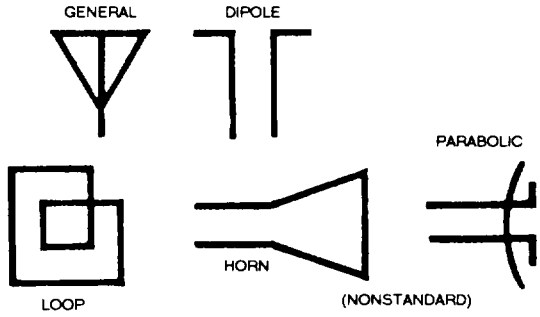
PUSH PULL OR PUSH

ATTENUATORS:



GENERAL BALANCED UNBALANCED

ANTENNAS:



METERS:



A - AMMETER

CRO - OSCILLOSCOPE

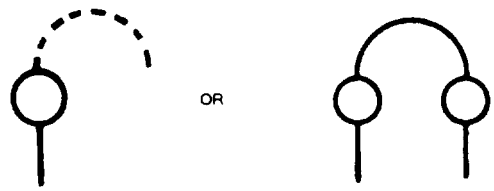
G - GALVANOMETER

MA - MILLIAMMETER

OHM - OHMMETER

V - VOLTMETER

HEADSET:



OR

FORMULAS

Ohm's Law for dc Circuits

$$I = \frac{E}{R} = \frac{P}{E} = \sqrt{\frac{P}{R}}$$

$$R = \frac{E}{I} = \frac{P}{I^2} = \frac{E^2}{P}$$

$$E = IR = \frac{P}{I} = \sqrt{PR}$$

$$P = EI = \frac{E^2}{R} = I^2R$$

Resistors in Series

$$R_T = R_1 + R_2 + \ldots$$

Resistors in Parallel

Two resistors

$$R_T = \frac{R_1R_2}{R_1 + R_2}$$

More than two

$$\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \ldots$$

RL Circuit Time Constant

$\frac{\text{L (in henrys)}}{\text{R (in ohms)}}$ = t (in seconds), or

$\frac{\text{L (in microhenrys)}}{\text{R (in ohms)}}$ = t (in microseconds)

RC Circuit Time Constant

R (ohms) × C (farads) = t (seconds)

R (megohms) × C (microfarads) = t (seconds)

R (ohms) × C (microfarads) = t (microseconds)

R (megohms) × C (micromicrofarads) = t (microseconds)

Capacitors in Series

Two capacitors

$$C_T = \frac{C_1C_2}{C_1 + C_2}$$

More than two

$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} + \ldots$$

Capacitors in Parallel

$$C_T = C_1 + C_2 + \ldots$$

Capacitive Reactance

$$X_C = \frac{1}{2\pi fC}$$

Impedance in an RC Circuit (Series)

$$Z = \sqrt{R^2 + (X_C)^2}$$

Inductors in Series

$L_T = L_1 + L_2 + \ldots$ (No coupling between coils)

Inductors in Parallel

Two inductors

$L_T = \frac{L_1L_2}{L_1 + L_2}$ (No coupling between coils)

More than two

$\frac{1}{L_T} = \frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} + \ldots$ (No coupling between coils)

Inductive Reactance

$$X_L = 2\pi fL$$

Q of a Coil

$$Q = \frac{X_L}{R}$$

Impedance of an RL Circuit (Series)

$$Z = \sqrt{R^2 + (X_L)^2}$$

Impedance with R, C, and L in Series

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

Parallel Circuit Impedance

$$Z = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

Sine-Wave Voltage Relationships

Average value

$$E_{ave} = \frac{2}{\pi} \times E_{max} = 0.637E_{max}$$

Effective or rms value

$$E_{eff} = \frac{E_{max}}{\sqrt{2}} = \frac{E_{max}}{1.414} = 0.707E_{max} = 1.11E_{ave}$$

Maximum value

$$E_{max} = \sqrt{2}(E_{eff}) = 1.414E_{eff} = 1.57E_{ave}$$

Voltage in an ac circuit

$$E = IZ = \frac{P}{I \times PF}$$

Current in an ac circuit

$$I = \frac{E}{Z} = \frac{P}{E \times PF}$$

Power in AC Circuit

Apparent power: $P = EI$

True power: $P = EI \cos \theta = EI \times PF$

Power Factor

$$PF = \frac{P}{EI} = \cos \theta$$

$$\cos \theta = \frac{\text{true power}}{\text{apparent power}}$$

Transformers

Voltage relationship

$$\frac{E_p}{E_s} = \frac{N_p}{N_s} \text{ or } E_s = E_p \times \frac{N_s}{N_p}$$

Current relationship

$$\frac{I_p}{I_s} = \frac{N_s}{N_p}$$

Induced voltage

$$E_{eff} = 4.44 \times BAfN \times 10^{-8}$$

Turns ratio

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}}$$

Secondary current

$$I_s = I_p \times \frac{N_p}{N_s}$$

Secondary voltage

$$E_s = E_p \times \frac{N_s}{N_p}$$

Three-Phase Voltage and Current Relationships

With wye connected windings

$$E_{line} = \sqrt{3}(E_{coil}) = 1.732E_{coil}$$

$$I_{line} = I_{coil}$$

With delta connected windings

$$E_{line} = E_{coil}$$

$$I_{line} = 1.732I_{coil}$$

With wye or delta connected winding

$$P_{coil} = E_{coil}I_{coil}$$

$$P_t = 3P_{coil}$$

$$P_t = 1.732E_{line}I_{line}$$

(To convert to true power multiply by $\cos \theta$)

Resonance

At resonance

$$X_L = X_C$$

Resonant frequency

$$F_o = \frac{1}{2\pi\sqrt{LC}}$$

Series resonance

$$Z \text{ (at any frequency)} = R + j(X_L - X_C)$$

$$Z \text{ (at resonance)} = R$$

Parallel resonance

$$Z_{max} \text{ (at resonance)} = \frac{X_L X_C}{R} = \frac{X_L{}^2}{R} = QX_L = \frac{L}{CR}$$

Bandwidth

$$\Delta = \frac{F_o}{Q} = \frac{R}{2\pi L}$$

Tube Characteristics

Amplification factor

$$\mu = \frac{\Delta e_p}{\Delta e_g} (i_p \text{ constant})$$

$$\mu = g_m r_p$$

AC plate resistance

$$r_p = \frac{\Delta e_p}{\Delta i_p} (e_g \text{ constant})$$

Grid-plate transconductance

$$g_m = \frac{\Delta i_p}{\Delta e_g} (e_p \text{ constant})$$

Decibels

NOTE: Wherever the expression "log" appears without a subscript specifying the base, the logarithmic base is understood to be 10.

Power ratio

$$dB = 10 \log \frac{P_2}{P_1}$$

Current and voltage ratio

$$dB = 20 \log \frac{I_2\sqrt{R_2}}{I_1\sqrt{R_1}}$$

$$dB = 20 \log \frac{E_2\sqrt{R_1}}{E_1\sqrt{R_2}}$$

NOTE: When R_1 and R_2 are equal they may be omitted from the formula. When reference level is one milliwatt

$$dBm = 10 \log \frac{P}{0.001} \text{ (when P is in watts)}$$

Synchronous Speed of Motor

$$rpm = \frac{120 \times \text{frequency}}{\text{number of poles}}$$

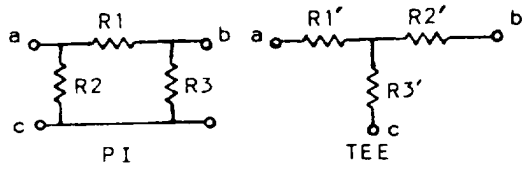
Wavelength

$$\text{wavelength (in meters)} = \frac{300}{\text{frequency (in megahertz)}}$$

$$\lambda = \frac{300}{f \text{ (MHz)}}$$

BRIDGE CIRCUIT CONVERSION FORMULAS

PI to Tee

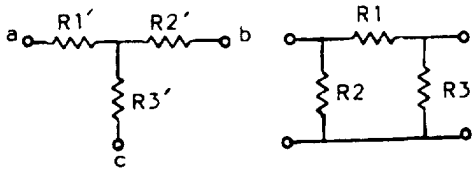


$$R_{1'} = \frac{R_1 R_2}{R_1 + R_2 + R_3}$$

$$R_{2'} = \frac{R_1 R_3}{R_1 + R_2 + R_3}$$

$$R_{3'} = \frac{R_2 R_3}{R_1 + R_2 + R_3}$$

Tee to PI

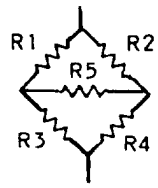


$$R_1 = \frac{R_{1'}R_{2'} + R_{2'}R_{3'} + R_{1'}R_{3'}}{R_{3'}}$$

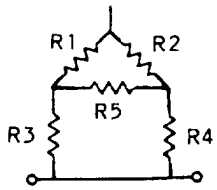
$$R_2 = \frac{R_{1'}R_{2'} + R_{2'}R_{3'} + R_{1'}R_{3'}}{R_{2'}}$$

$$R_3 = \frac{R_{1'}R_{2'} + R_{2'}R_{3'} + R_{1'}R_{3'}}{R_{1'}}$$

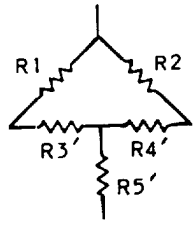
Calculating R_T for Bridge



1. Redraw



2. Convert PI network made up of resistors $R_3R_4R_5$ to Tee network made up of $R_{3'}R_{4'}R_{5'}$

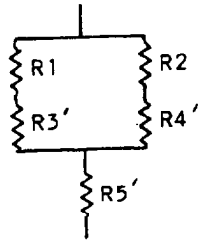


$$R_{3'} = \frac{R_3 R_5}{R_3 + R_4 + R_5}$$

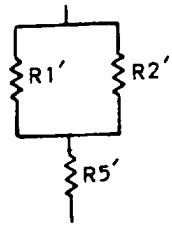
$$R_{4'} = \frac{R_4 R_5}{R_3 + R_4 + R_5}$$

$$R_{5'} = \frac{R_3 R_4}{R_3 + R_4 + R_5}$$

3. Redraw circuit

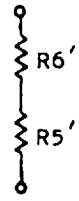


4. Simplify circuit by combining



$$R_{1'} = R_1 + R_{3'} \qquad R_{2'} = R_2 + R_{4'}$$

5. Simplify again



$$R_{6'} = \frac{R_{1'}R_{2'}}{R_{1'} + R_{2'}}$$

6. Solve for R_T

$$R_T = R_{6'} + R_{5'}$$

Comparison of Units in Electric and Magnetic Circuits

	Electric circuit	Magnetic circuit
Force	Volt, E, or emf	Gilberts, F, or mmf
Flow	Ampere, I	Flux , Φ, in maxwells
Opposition	Ohms, R	Reluctance, $\mathfrak{R}$
Law	Ohm's law, $I = \frac{E}{R}$	Rowland's law, $\Phi = \frac{F}{\mathfrak{R}}$
Intensity of force	Volts per cm of length.	$H = \frac{1.257IN}{L}$, gilberts per centimeter of length.
Density	Current density–for example, amperes per cm^2.	Flux density–for example, lines per cm^2 or gausses.

U.S. CUSTOMARY AND METRIC SYSTEM
UNITS OF MEASUREMENTS

Multiples and Submultiples		Prefixes	Symbols
1 000 000 000 000	$= 10^{12}$	tera (tĕr′ȧ)	T
1 000 000 000	$= 10^{9}$	giga (jĭ′gȧ)	G
1 000 000	$= 10^{6}$	mega (mĕg′ȧ)	M*
1 000	$= 10^{3}$	kilo (kĭl′ō)	k*
100	$= 10^{2}$	hecto (hĕk′tō)	h
10	$= 10$	deka (dĕk′ȧ)	da
0.1	$= 10^{-1}$	deci (dĕs′ĭ)	d
0.01	$= 10^{-2}$	centi (sĕn′tĭ)	c*
0.001	$= 10^{-3}$	milli (mĭl′ĭ)	m*
0.000 001	$= 10^{-6}$	micro (mī′krō)	μ*
0.000 000 001	$= 10^{-9}$	nano (năn′ō)	n
0.000 000 000 001	$= 10^{-12}$	pico (pē′kō)	p
0.000 000 000 000 001	$= 10^{-15}$	femto (fĕm′tō)	f
0.000 000 000 000 000 001	$= 10^{-18}$	atto (ăt′tō)	a

*Most commonly used

GREEK ALPHABET

Name	Capital	Lower Case	Designates
Alpha	A	α	Angles, coefficient of thermal expansion.
Beta	B	β	Angles, flux density.
Gamma	Γ	γ	Conductivity.
Delta	Δ	δ	Variation of a quantity, increment.
Epsilon	E	ε	Base of natural logarithms (2.71828).
Zeta	Z	ζ	Impedance, coefficients, coordinates.
Eta	H	η	Hysteresis coefficient, efficiency, magnetizing force.
Theta	Θ	θ	Phase angle.
Iota	I	ι	
Kappa	K	κ	Dielectric constant, coupling coefficient, susceptibility.
Lambda	Λ	λ	Wavelength.
Mu	M	μ	Permeability, micro, amplification factor.
Nu	N	ν	Reluctivity.
Xi	Ξ	ξ	
Omicron	O	o	
Pi	Π	π	3.1416
Rho	P	ρ	Resistivity.
Sigma	Σ	ς	Summation symbol (cap).
Tau	T	τ	Time constant, time-phase displacement.
Upsilon	Y	υ	
Phi	Φ	ϕ	Angles, magnetic flux.
Chi	X	χ	
Psi	Ψ	ψ	Dielectric flux, phase difference.
Omega	Ω	ω	Ohms (capital), angular velocity ($2 \pi f$).

LAWS OF EXPONENTS

The International Symbols Committee has adopted prefixes for denoting decimal multiples of units. The National Bureau of Standards has followed the recommendations of this committee, and has adopted the following list of prefixes:

Numbers	Powers of ten	Prefixes	Symbols
1,000,000,000,000	10^{12}	tera	T
1,000,000,000	10^{9}	giga	G
1,000,000	10^{6}	mega	M
1,000	10^{3}	kilo	k
100	10^{2}	hecto	h
10	10	deka	da
.1	10^{-1}	deci	d
.01	10^{-2}	centi	c
.001	10^{-3}	milli	m
.000001	10^{-6}	micro	u
.000000001	10^{-9}	nano	n
.000000000001	10^{-12}	pico	p
.000000000000001	10^{-15}	femto	f
.000000000000000001	10^{-18}	atto	a

To multiply like (with same base) exponential quantities, add the exponents. In the language of alegebra the rule is $a^m \times a^n = a^{m+n}$

$$10^4 \times 10^2 = 10^{4+2} = 10^6$$

$$0.003 \times 825.2 = 3 \times 10^{-3} \times 8.252 \times 10^2$$

$$= 24.756 \times 10^{-1} = 2.4756$$

To divide exponential quantities, subtract the exponents. In the language of algebra the rule is

$$\frac{a^m}{n} = a^{m-n}$$

or

$$10^8 \div 10^2 = 10^6$$

$$3{,}000 + 0.015 = 0.015 = (3 \times 10^3) + (1.5 \times 10^{-2})$$

$$= 2 \times 10^5 = 200{,}000$$

To raise an exponential quantity to a power, multiply the exponents. In the language of algebra $(x^m)^n = x^{mn}$.

$$(10^3)^4 = 10^{3 \times 4} = 10^{12}$$

$$2{,}500^2 = (2.5 \times 10^3)^2 = 6.25 \times 10^6 = 6{,}250{,}000$$

Any number (except zero) raised to the zero power is one. In the language of algebra $x^0 = 1$

$$x^3 \div x^3 = 1$$

$$10^4 \div 10^4 = 1$$

Any base with a negative exponent is equal to 1 divided by the base with an equal positive exponent. In the language of algebra $x^{-a} = \frac{1}{x^a}$

$$10^{-2} = \frac{1}{10^2} = \frac{1}{100}$$

$$5a^{-3} = \frac{5}{a^3}$$

$$(6a)^{-1} = \frac{1}{6a}$$

To raise a product to a power, raise each factor of the product to that power.

$$(2 \times 10)^2 = 2^2 \times 10^2$$

$$3{,}000^3 = (3 \times 10^3)^3 = 27 \times 10^9$$

To find the nth root of an exponential quantity, divide the exponent by the index of the root. Thus, the nth root of $a^m = a^{m/n}$.

$$\sqrt{x^6} = x^{6/2} = x^3$$

$$\sqrt[3]{64 \times 10^3} = 4 \times 10 = 40$$

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