Q.2 a. Explain about six applications of radars.

Answer:
Radar has been employed on the ground, in the air, on the sea, and in space. Ground-based radar has been applied chiefly to the detection, location, and tracking of aircraft or space targets.

Shipboard radar is used as a navigation aid and safety device to locate buoys, shore lines, and other ships as well as for observing aircraft. Airborne radar may be used to detect other aircraft, ships, or land vehicles, or it may be used for mapping of land, storm avoidance, terrain avoidance, and navigation. In space, radar has assisted in the guidance of spacecraft and for the remote sensing of the land and sea.

The major user of radar, and contributor of the cost of almost all of its development, has been the military: although there have been increasingly important civil applications, chiefly for marine and air navigation. The major areas of radar application, in no particular order of importance, are described below. Air Traffic Control (ATC): Radars are employed throughout the world for the purpose of safely controlling air traffic en route and in the vicinity of airports. Aircraft and ground vehicular traffic at large airports are monitored by means of high-resolution radar. Radar has been used with GCA (ground-control approach) systems to guide aircraft to a safe landing in bad weather. In addition, the microwave landing system and the widely used ATC radar-beacon system are based in large part on radar technology.

Aircraft Navigation: The weather-avoidance radar used on aircraft to outline regions of precipitation to the pilot is a classical form of radar. Radar is also used for terrain avoidance and terrain following. Although they may not always be thought of as radars, the radio altimeter (either FM/CW or pulse) and the doppler navigator are also radars. Sometimes ground-mapping radars of moderately high resolution are used for aircraft navigation purposes.

Ship Safety: Radar is used for enhancing the safety of ship travel by warning of potential collision with other ships, and for detecting navigation buoys, especially in poor visibility. In terms of numbers, this is one of the larger applications of radar, but in terms of physical size and cost it is one of the smallest. It has also proven to be one of the most reliable radar systems. Automatic detection and tracking equipments (also called plot extractors) are commercially available for use with such radars for the purpose of collision avoidance. Shore-based radar of moderately high resolution is also used for the surveillance of harbors as an aid to navigation.

Space: Space vehicles have used radar for rendezvous and docking, and for landing on the moon. Some of the largest ground-based radars are for the detection and tracking of satellites. Satellite-borne radars have also been used for remote sensing as mentioned below. Remote **Sensing**: All radars are remote sensors; however, as this term is used it implies the sensing of geophysical objects, or the "environment." For some time, radar has been used as a remote sensor of the weather. It was also used in the past to probe the moon and the planets (radar astronomy). The ionospheric sounder, an important adjunct for HF (short wave) communications, is a radar. Remote sensing with radar is also concerned with Earth resources, which includes the measurement and mapping of sea conditions, water resources, ice cover, agriculture, forestry conditions, geological formations, and environmental pollution. The platforms for such radars include satellites as well as aircraft.

Law Enforcement: In addition to the wide use of radar to measure the speed of automobile
traffic by highway police, radar has also been employed as a means for the detection of intruders.

Military: Many of the civilian applications of radar are also employed by the military. The traditional role of radar for military application has been for surveillance, navigation, and for the control and guidance of weapons. It represents, by far, the largest use of radar.

b. What are the desirable pulse characteristics and the factors that govern them in a Radar system?

Answer:
Radar is an electromagnetic system for the detection and location of objects. It operates by transmitting a particular type of waveform, a pulse-modulated sine wave for example, and detects the nature of the echo signal.

The most common radar waveform is a train of narrow, rectangular-shape pulses modulating a sinewave carrier. The distance, or range, to the target is determined by measuring the time TR taken by the pulse to travel to the target and return. Since electromagnetic energy propagates at the speed of light c = 3 x 108 m/s, the range R is

$$R = \frac{cT_R}{2}$$

The factor 2 appears in the denominator because of the two-way propagation of radar. With the range in kilometers or nautical miles, and TR in microseconds, Eq. above becomes

$$R(\text{km}) = 0.15T_R(\mu s) \qquad \text{or} \qquad R(\text{nmi}) = 0.081\,T_R(\mu s)$$

Each microsecond of round-trip travel time corresponds to a distance of 0.081 nautical mile, 0.093 statute mile, 150 meters, 164 yards, or 492 feet. Once the transmitted pulse is emitted by the radar, a sufficient length of time must elapse to allow any echo signals to return and be detected before the next pulse may be transmitted. Therefore the rate at which the pulses may be transmitted is determined by the longest range at which targets are expected. If the pulse repetition frequency is too high, echo signals from some targets might arrive after the transmission of the next pulse, and ambiguities in measuring range might result.

Echoes that arrive after the transmission of the next pulse are called second-time-around (or multiple-time-around) echoes. Such an echo would appear to be at a much shorter range than the actual and could be misleading if it were not known to be a second-time-around echo. The range beyond which targets appear as second-time-around echoes is called the maximum unambiguous range and is

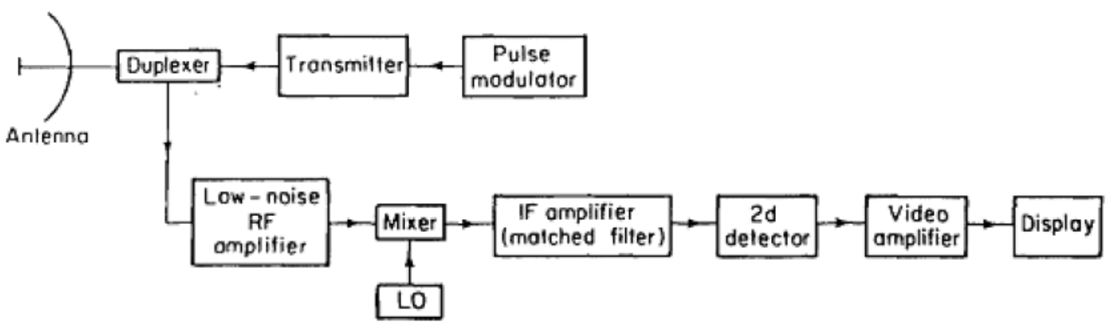
$$R_{unamb} = \frac{c}{2f_p}$$

Where fp = pulse repetition frequency, in Hz. A plot of the maximum unambiguous range

[Pulse Train, Timing equation, Mention of echo should be there in answer]

c. **Draw the functional block diagram of simple pulse radar and explain the purpose and functioning of each block in it.**

Answer:



The operation of a typical pulse radar may be described with the aid of the block diagram shown in Fig.. The transmitter may be an oscillator, such as a magnetron, that is "pulsed" (turned on and on) by the modulator to generate a repetitive train of pulses. The magnetron has probably been the most widely used of the various microwave generators for radar. A typical radar for the detection of aircraft at ranges of 100 or 200 nmi might employ a peak power of the order of a megawatt, an average power of several kilowatts, a pulse width of several microseconds, and a pulse repetition frequency of several hundred pulses per second.

The waveform generated by the transmitter travels via a transmission line to the antenna, where it is radiated into space. A single antenna is generally used for both transmitting and receiving. The receiver must be protected from damage caused by the high power of the transmitter. This is the function of the duplexer. The duplexer also serves to channel the returned echo signals to the receiver and not to the transmitter. The duplexer might consist of two gas-discharge devices, one known as a TR (transmit-receive) and the other an ATR (anti-transmit-receive). The TR protects the receiver during transmission and the ATR directs the echo signal to the receiver during reception. Solid-state ferrite circulators and receiver protectors with gas-plasma TR devices and/or diode limiters are also employed as duplexers.

The receiver is usually of the superheterodyne type. The first stage might be a low-noise RF amplifier, such as a parametric amplifier or a low-noise transistor. However, it is not always desirable to employ a low-noise first stage in radar. The receiver input can simply be the mixer stage, especially in military radars that must operate in a noisy environment. Although a receiver with a low-noise front-end will be more sensitive, the mixer input can have greater dynamic range, less susceptibility to overload, and less vulnerability to electronic interference.

The mixer and local oscillator (LO) convert the RF signal to an intermediate frequency (IF). A " typical" IF amplifier for an air-surveillance radar might have a center frequency of 30 or 60 MHz and a bandwidth of the order of one megahertz. The IF amplifier should be designed as a notched filter; i.e., its frequency-response function H (f) should maximize the peak-signal-to-mean-noise-power ratio at the output. This occurs when the magnitude of the frequency-response function $|H(f)|$ is equal to the magnitude of the echo signal spectrum $|S(f)|$, and the phase spectrum of the matched filter is the negative of the phase spectrum of the echo signal . In a radar whose signal waveform approximates a rectangular pulse, the conventional IF filter bandpass characteristic approximates a matched filter when the product of the IF bandwidth B and the pulse width τ is of the order of unity, that is, $B\tau \approx 1$.

After maximizing the signal-to-noise ratio in the IF amplifier, the pulse modulation is extracted by the second detector and amplified by the video amplifier to a level where it can be properly displayed, usually on a cathode-ray tube (CRT).

Q.3 a. Derive the maximum range for a radar system from first principles.

Answer:

The radar equation relates the range of a radar to the characteristics of the transmitter, receiver, antenna, target, and environment. It is useful not just as a means for determining the maximum distance from the radar to the target, but it can serve both as a tool for understanding radar operation and as a basis for radar design.

If the power of the radar transmitter is denoted by Pt , and if an isotropic antenna is used (one which radiates uniformly in all directions), the power density (watts per unit area) at a distance R from the radar is equal to the transmitter power divided by the surface area 4Π R2 of an imaginary sphere of radius R, or

$$\text{Power density from isotropic antenna} = \frac{P_t}{4\pi R^2}$$

Radars employ directive antennas to channel, or direct, the radiated power Pt into some particular direction. The gain G of an antenna is a measure of the increased power radiated in the direction of the target as compared with the power that would have been radiated from an isotropic antenna. It may be defined as the ratio of the maximum radiation intensity from the subject antenna to the radiation intensity from a lossless, isotropic antenna with the same power input. (The radiation intensity is the power radiated per unit solid angle in a given direction.) The power density at the target from an antenna with a transmitting gain G is

$$\text{Power density from directive antenna} = \frac{P_t G}{4\pi R^2}$$

The target intercepts a portion of the incident power and reradiates it in various directions. The measure of the amount of incident power intercepted by the target and reradiated back in the direction of the radar is denoted as the radar cross section σ, and is defined by the relation

$$\text{Power density of echo signal at radar} = \frac{P_t G}{4\pi R^2}\frac{\sigma}{4\pi R^2}$$

The radar cross section σ has units of area. It is a characteristic of the particular target and is a measure of its size as seen by the radar. The radar antenna captures a portion of the echo power. If the effective area of the receiving antenna is denoted Ae, the power Pr, received by the radar is

$$P_r = \frac{P_t G}{4\pi R^2}\frac{\sigma}{4\pi R^2} A_e = \frac{P_t G A_e \sigma}{(4\pi)^2 R^4}$$

The maximum radar range Rmax is the distance beyond which the target cannot be detected. It occurs when the received echo signal power P, just equals the minimum detectable signal Smin, Therefore

$$R_{max} = \left[\frac{P_t G A_e \sigma}{(4\pi)^2 S_{min}}\right]^{1/4}$$

This is the fundamental form of the radar equation. Note that the important antenna parameters are the transmitting gain and the receiving effective area.

$$G = \frac{4\pi A_e}{\lambda^2}$$

Antenna theory gives the relationship between the transmitting gain and the receiving effective area of an antenna as Since radars generally use the same antenna for both transmission and reception, Eq. can be substituted into Eq. above, first for Ae, then for G, to give two other forms of the radar equation

$$R_{max} = \left[\frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 S_{min}}\right]^{1/4}$$

$$R_{max} = \left[\frac{P_t A_e^2 \sigma}{4\pi \lambda^2 S_{min}}\right]^{1/4}$$

b. Describe how threshold level for detection is decided in the presence of receiver noise for a specified probability of occurrence of false alarms by applying statistical noise theory.

Answer:

Consider an IF amplifier with bandwidth BIF followed by a second detector and a video amplifier with bandwidth Bv, (Fig. 1). The second detector and video amplifier are assumed to form an envelope detector, that is, one which rejects the carrier frequency but passes the modulation envelope. To extract the modulation envelope, the video bandwidth must be wide enough to pass the low-frequency components generated by the second detector, but not so wide as to pass the high-frequency components at or near the intermediate frequency. The video bandwidth Bv, must be greater than BIF /2 in order to pass all the video modulation. Most radar receivers used in conjunction with an operator viewing a CRT display meet this condition and may be considered envelope detectors. Either a square-law or a linear detector may be assumed since the effect on the detection probability by assuming one instead of the other is usually small.

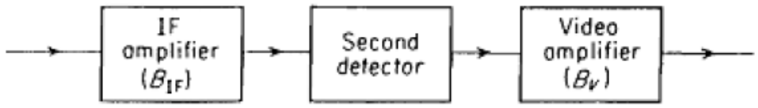


Fig.1 Envelope detector

The noise entering the IF filter (the terms filter and amplifier are used interchangeably) is assumed to be gaussian, with probability-density function given by

$$p(v) = \frac{1}{\sqrt{2\pi\psi_0}} \exp \frac{-v^2}{2\psi_0}$$

where p(v) dv is the probability of finding the noise voltage v between the values of v and v + dv, Ψo is the variance, or mean-square value of the noise voltage, and the mean value of v is taken to be zero. If gaussian noise were passed through a narrowband IF filter-one whose bandwidth is small compared with the mid frequency-the probability density of the envelope of the noise voltage output is shown by Rice to be

$$p(R) = \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right)$$

where R is the amplitude of the envelope of the filter output. Equation above is a form of the Rayleigh probability-density function. The probability that the envelope of the noise voltage will lie between the values of V1 and V2 is

$$\text{Probability } (V_1 < R < V_2) = \int_{V_1}^{V_2} \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) dR$$

The probability that the noise voltage envelope will exceed the voltage threshold VT is

$$\text{Probability } (V_T < R < \infty) = \int_{V_T}^{\infty} \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) dR$$

$$= \exp\left(-\frac{V_T^2}{2\psi_0}\right) = P_{fa}$$

Whenever the voltage envelope exceeds the threshold, target detection is considered to have occurred, by definition. Since the probability of a false alarm is the probability that noise will cross the threshold, Eq. above gives the probability of a false alarm, denoted Pfa The average time interval between crossings of the threshold by noise alone is defined as the false-alarm time T_{fa},

$$T_{fa} = \lim_{N\to\infty} \frac{1}{N} \sum_{k=1}^{N} T_k$$

where T_k is the time between crossings of the threshold V_T by the noise envelope, when the slope of the crossing is positive. The false-alarm probability may also be defined as the ratio of the duration of time the envelope is actually above the threshold to the total time it could have been above the threshold, or

$$P_{fa} = \frac{\sum_{k=1}^{N} t_k}{\sum_{k=1}^{N} T_k} = \frac{\langle t_k \rangle_{av}}{\langle T_k \rangle_{av}} = \frac{1}{T_{fa} B}$$

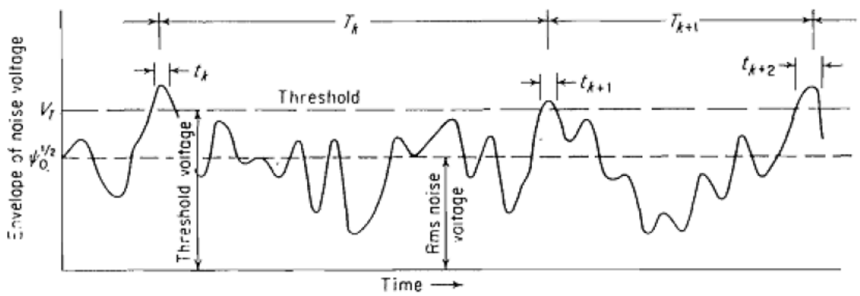


Fig.2 Envelope of receiver output illustrating false alarms due to noise.
where t_k and T_k are defined in Fig.1. The average duration of a noise pulse is approximately the reciprocal of the bandwidth B, which in the case of the envelope detector is BIF. Equating equations discussed above we get

$$T_{fa} = \frac{1}{B_{IF}} \exp \frac{V_T^2}{2\psi_0}$$

A plot of Eq. above is shown in Fig.2, with V2T / 2Ψo as the abscissa. If, for example, the bandwidth of the IF amplifier were 1MHz and the average false-alarm time that could be tolerated were 15 min, the probability of a false alarm is 1.11 x 10-9 . From Eq. above the threshold voltage necessary to achieve this false-alarm time is 6.45 times the rms value of the noise voltage.
The false-alarm probabilities of practical radars are quite small. The reason for this is that the false-alarm probability is the probability that a noise pulse will cross the threshold during an interval of time approximately equal to the reciprocal of the bandwidth. For a 1- MHz bandwidth, there are of the order of 106 noise pulses per second. Hence the false alarm probability of any one pulse must be small (< 10-6) if false-alarm times greater than 1s are to be obtained.

c. An X band monopulse tracking radar operates at 9.7 GHz and is required to have S/N=30 db in the receiver sum channel when a 1 m2 constant RCS target is engaged at arrange of 40 Km. The radar has an antenna with a diameter of 1.8m and an efficiency of 65%. The transmitter has pulse duration of 0.5 μs and the system noise temperature for all channels is 1540 K, the radar integrates 40 pulses. RF loses are 5 dB on transmit and 4 dB on receive. Do not add any miscellaneous losses. Use the optimum receiver bandwidth and taking account of the receive filter mismatch loss; determine the transmitter pulse power required to meet this specification.

Answer:

Optimum receiver bandwidth $B = 1.4 / \tau = 2.8$ MHz, with mismatch loss of 0.9 dB.
Antenna gain is $G = 0.6 \times (\pi D / \lambda)^2 = 0.65 \times (\pi\, 1.8 / 0.03093)^2 = 21{,}727$ or 43.4 dB.
With $T_s = 1540$ K the noise power in the receiver is

$$k\, T_s\, B = -228.6 + 31.9 + 64.5 = -132.3 \text{ dBW}$$

For a S/N ratio of 30 dB we need a receiver power of $P_r = -102.3$ dBW.
Total losses in the radar are $5 + 4 + 0.9 = 9.9$ dB
(There is no scan loss, since we are tracking the target).
Integration gain for 40 pulses is $16 - 2.7 = 13.3$ dB
Applying the radar equation at a range of 40 km

$$P_r = P_t + 2\,G + 2\lambda + \sigma - 33 - 4R + G_{int} - \text{losses}$$

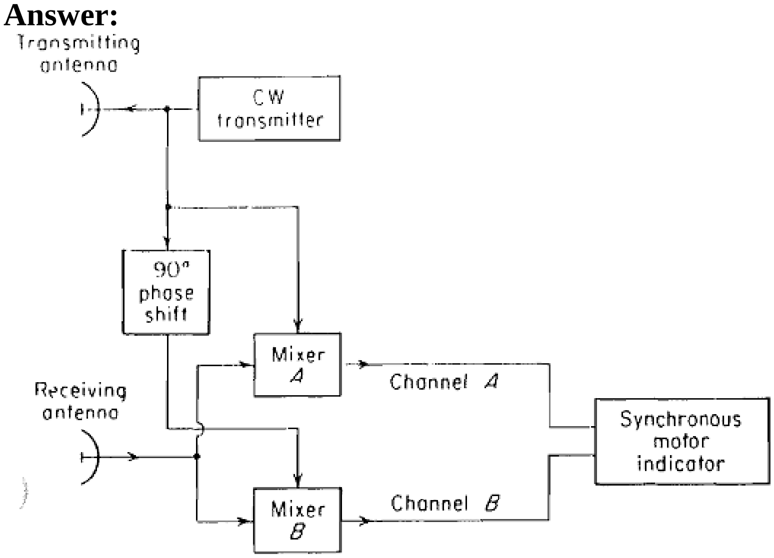
$$-102.3 = P_t \text{ dBW} + 86.8 - 30.2 + 0 - 33 - 184.0 + 13.3 - 9.9$$

Hence

$$P_t = 54.7 \text{ dBW or } 295 \text{ kW}$$

Q.4 a. Explain with necessary block diagram, how doppler direction is identified with CW radar.

Answer:



The sign of the doppler frequency, and therefore the direction of target motion, may be found by splitting the received signal into two channels as shown in Fig. . In channel A the signal is processed as in the simple CW radar. The received signal and a portion of the transmitter heterodyne in the detector (mixer) to yield a difference signal
EA = K2 E0 cos (±wd t + φ)
EA = amplitude of transmitter signal
K 2 = a constant determined from the radar equation
w d = dopper angular frequency shift
φ = a constant phase shift, which depends upon range of initial detection
The other channel is similar, except for a 90° phase delay introduced in the reference signal. The output of the channel B mixer is
EB = K2 E0 cos (±wd t + φ + π / 2)
If the target is approaching (positive Doppler), the outputs from the two channels are
EA (+) = K2 E0 cos (wd t + φ)

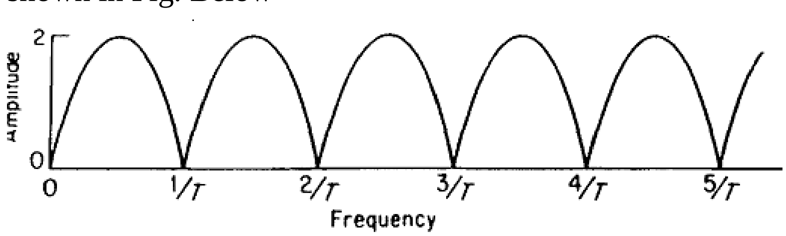
EB (+)= K2 E0 cos (wd t + φ + π / 2)
If the targets are receding (negative doppler), the outputs from the two channels are
EA (-) = K2 E0 cos (wd t - φ)
EB (-)= K2 E0 cos (wd t - φ - π / 2)

The sign of wd and the direction of the target's motion may be determined according to whether the output of channel B leads or lags the output of channel A. One method of determining the relative phase relationship between the two channels is to apply the outputs to a synchronous two-phase motor. The direction of motor rotation is an indication of the direction of the target motion.

Electronic methods may be used instead of a synchronous motor to sense the relative phase of the two channels. One application of this technique has been described for a rate-of climb meter for vertical take-off aircraft to determine the velocity of the aircraft with respect to the ground during take-off and landing. It has also been applied to the detection of moving targets in the presence of heavy foliage.

b. Explain in detail the Filter characteristics of the delay-line canceller.

Answer:
Filter characteristics of the delay-line canceller. The delay-line canceller acts as a filter which rejects the d-c component of clutter. Because of its periodic nature, the filter also rejects energy in the vicinity of the pulse repetition frequency and its harmonics. The video signal received from a particular target at a range R0 is
V1 = k sin (2πfdt – φ0)
Where, φ0 = phase shift
k = amplitude of video signal.
The signal from the previous transmission, which is delayed by a time T = pulse repetition interval, is
V2 = k sin (2πfd(t – T) – φ0)
Everything else is assumed to remain essentially constant over the interval T so that k is the same for both pulses. The output from the subtractor is
V = V1 – V2 = 2*k sin πfdT cos [2πfd(t – T / 2) - φ0]
It is assumed that the gain through the delay-line canceller is unity. The output from the canceller V consists of a cosine wave at the doppler frequency fd with an amplitude 2*k sin πfdT . Thus the amplitude of the canceled video output is a function of the Doppler frequency shift and the pulse-repetition interval, or prf. The magnitude of the relative frequency-response of the delay-line canceller [ratio of the amplitude of the output from the delay-line canceller, 2*k sin πfdT, to the amplitude of the normal radar video kj is shown in Fig. Below



c. Explain the Butterfly effect that is produced by MTI.

Answer:

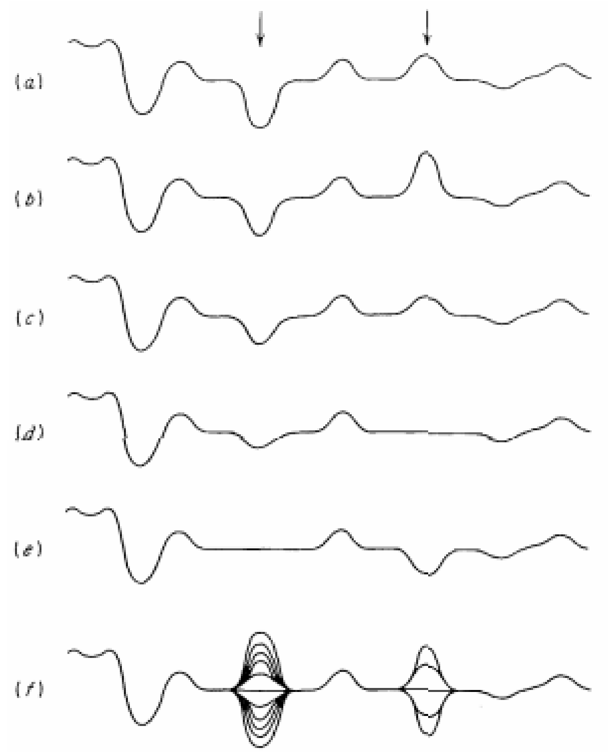


Fig (a-e) Successive sweeps of an MTI radar A-scope display (echo amplitude as a function of time); (f) superposition of many sweeps; arrows indicate position of moving targets.

Moving targets may be distinguished from stationary targets by observing the video output on an A-scope (amplitude vs. range). A single sweep on an A-scope might appear as in Fig. (a). This sweep shows several fixed targets and two moving targets indicated by the two arrows. On the basis of a single sweep, moving targets cannot be distinguished from fixed targets. (It may be possible to distinguish extended ground targets from point targets by the stretching of the echo pulse. However, this is not a reliable means of discriminating moving from fixed targets since some fixed targets can look like point targets, e.g., a water tower. Also, some moving targets such as aircraft flying in formation can look like extended targets.) Successive A scope sweeps (pulse-repetition intervals) are shown in Fig. (b) to (e). Echoes from fixed targets remain constant throughout but echoes from moving targets vary in amplitude from sweep to sweep at a rate corresponding to the doppler frequency. The superposition of the successive A-scope sweeps is shown in Fig. (J). The moving targets produce, with time, a butterfly effect on the A-scope.

Q.5 a. Derive the expression for frequency response of the matched filter with non-white noise.

Answer:
Matched filter with nonwhite noise: In the derivation of the matched-filter characteristic, the spectrum of the noise accompanying the signal was assumed to be white; that is, it was independent of frequency. If this assumption were not true, the filter which maximizes the output signal-to-noise ratio would not be the same as the matched filter. It has been shown that if the input power spectrum of the interfering noise is given by [Ni (f)]2, the frequency-response function of the filter which maximizes the output signal-to-noise ratio is

$$H(f) = \frac{G_a S^*(f) \exp(-j2\pi f t_1)}{[N_i(f)]^2}$$

When the noise is nonwhite, the filter which maximizes the output signal-to-noise ratio is called the NWN (nonwhite noise) matched filter. For white noise [Ni(f)]2 = constant and the NWN matched-filler frequency-response function of Eq. above reduces to that of Eq. discussed earlier in white noise. Equation above can be written as

$$H(f) = \frac{1}{N_i(f)} \times G_a \left(\frac{S(f)}{N_i(f)} \right)^* \exp(-j2\pi f t_1)$$

This indicates that the NWN matched filter can be considered as the cascade of two filters. The first filter, with frequency-response function l/Ni (f), acts to make the noise spectrum uniform, or white. It is sometimes called the whitening filter. The second is the matched filter when the input is white noise and a signal whose spectrum is S(f)/Ni(f).

b. Derive the surface clutter radar equation.

Answer:

Since the clutter signal is not a steady signal, it fluctuates with time and space; therefore it is better to consider the clutter signal as being a random sequence and to study its statistical properties are beyond the scope of this research.

In designing surface clutter models, Wetzel L.B (1990) has developed different models based on the grazing angle which can be effective to model sea clutter.

Figure 3.3, depicts radar illuminating the surface at a grazing angle ψ, it is assumed the width of area A_c is determined by the azimuth beam width θ_B

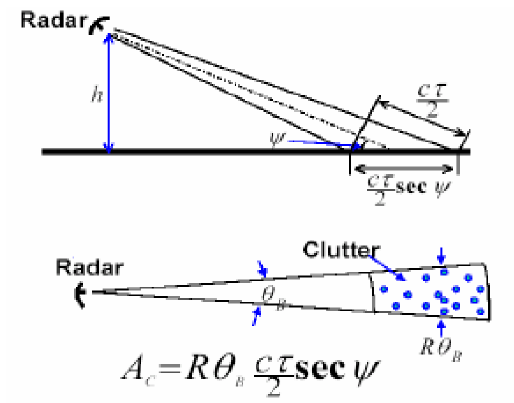


Figure 3.3. Geometry of radar clutter. Top: Elevation view showing the extent of the surface illuminated by the radar pulse, Bottom: Plan view showing clutter resolution cell consisting of individual independent scatterers.

Low Grazing Angle ψ, depicts a radar illuminating the surface at a small grazing angle **ψ.** For low grazing angles the range is determined by the radar pulse width τ. The cell width is determined by the azimuth beam width θ_B the range R. Utilizing the simple radar equation, the received echo power P_r is

$$P_r = \frac{P_t G A_e \sigma}{(4\pi)^2 R^4} \qquad (8)$$

where P_t = transmitter power, W
G = antenna gain
A_e = antenna effective aperture m²
R = range, m
σ = radar cross section of the scatterer, m²

To model target echoes (rather than clutter), we let $P_r = S$ (received target signal power) and $\sigma = \sigma_v$ (target cross section). The signal power returned from target power is then

$$S = \frac{P_t G A_e \sigma_t}{(4\pi)^2 R^4} \qquad (9)$$

When the echo is from clutter, the cross section σ becomes $\sigma_c = \sigma_0 A_c$, where the area A_c of the radar resolution cell is

$$A_c = R\,\theta_B\,(c\tau/2)\sec\psi \qquad (10)$$

The smallest area A_c, of the sea surface within which individual targets can no longer be individually observed is termed resolution cell.

With θ_B = two way azimuth beamwidth, c = velocity of propagation, τ = pulse width and ψ = grazing angle (defined with respect to surface tangent). The Area A_c in range resolution, is (c τ / 2), where the factor of 2 in the denominator represents for the two way propagation of radar. With these premises and definitions, the radar equation for the surface clutter each signal power C is

$$C = \frac{P_t G A_e [\theta_B (c\tau/2)\sec\psi}{(4\pi)^2 R^3} \qquad (11)$$

When the echo from the surface clutter is large compared to receiver noise, the signal-to-clutter ratio is

$$\frac{S}{C} = \frac{\sigma_t}{\sigma_0 R\,\theta_B\,(c\tau/2)\sec\psi} \qquad (12)$$

If the maximum range, R_{max}, corresponds to the minimum discernible signal-to-clutter ratio (S/C_{min}), then the radar equation for the detection of target in clutter at low grazing angle is

$$R_{max} = \frac{\sigma_t}{(S/C)_{min}\, \sigma^0\, \theta_B\, (c\,\tau/2)\, \sec\psi} \qquad (13)$$

Notice the range appears as the first power rather than the fourth power as in the usual noise-dominated radar equation, which results in greater variation of the maximum range of a clutter dominated radar. When there is uncertainty in the range, statistical representations are useful.

The characteristics of clutter echoes are normally given in statistical terms. It is often described either by the mean (average) value of σ_0 or the median (the value exceeded 50 percent of the time). For a Rayleigh pdf, the difference between the mean and median is small (a few percent), but for non-Rayleigh distributions, the mean and the median can be quite different.

Q.6a. A weather radar operates at a frequency of 2.85 GHz. At this frequency, the radar cross section of an individual spherical drop is given by

$$\sigma_{drop} = \pi^5 D^6 |k^2| / \lambda^4 m^2$$

(i) What radar band is being used by this radar? Calculate the RCS of a spherical drop with diameter of 1, 3 and 5 mm.

(ii) A weather has a beam with a circular cross-section and beamwidth of 1.5 degrees. The pulse length is 1μs. calculate target volume of the radar at a range of 50 Km. On average there are 8000 drops per cubic meter in a rainstorm. Determine the number of drops in the target volume and hence the rain radar cross sectional area in meters squared and dB meters^2 assuming all drops are 1 mm in diameter. Ignore all the beam shape factors.

Answer:

(i)

Answers: This is an S-band radar. Check the calculation for 1 mm: $\sigma_{drop} = \pi^5 (.001)^6 (0.93) / (0.105)^4$ m^2 = 2.34×10^{-12} m^2. For 3 mm and 5 mm drops we have σ_{drop} = $1.71\ 10^{-9}$ m^2 and σ_{drop} = $3.7\ 10^{-8}$ m^2, respectively.

(ii)

Answers: Beam volume = R Θ_B x R Θ_B x $c\tau/2$ = $(50\ 10^3\ m)^2 (1.5\pi/180)^2 (150\ m)$ = $0.258\ 10^9$ m^3 for a rectangular beam area (or $0.16\ 10^9$ m^3 for a circular beam area). The number of drops in the target volume is N = 8000 x $0.258\ 10^9$ = $2.06\ 10^{12}$ drops. The rain RCS of the target volume assuming 1 mm drops is RCS = n x σ_{drop} = 4.8 m^2 = 6.8 dB m^2.

b. Explain various reflector antennas.

Answer: Types of Reflector – Paraboloid, Offset-Fed, Cassegrain, refer Pages 553, 556, 558 of Text Book-I

Q.7 a. Explain Balanced type duplexer.

Answer:
Balanced duplexer: The balanced duplexer, Fig.2, is based on the short-slot hybrid junction which consists of two sections of waveguides joined along one of their narrow walls with a slot cut in the common narrow wall to provide coupling between the two. The short-slot hybrid may be considered as a broadband directional coupler with a coupling ratio of 3 dB. In the transmit condition (Fig. 2 a) power is divided equally into each waveguide by the first short slot hybrid junction. Both TR tubes break down and reflect the incident power out the antenna arm as shown

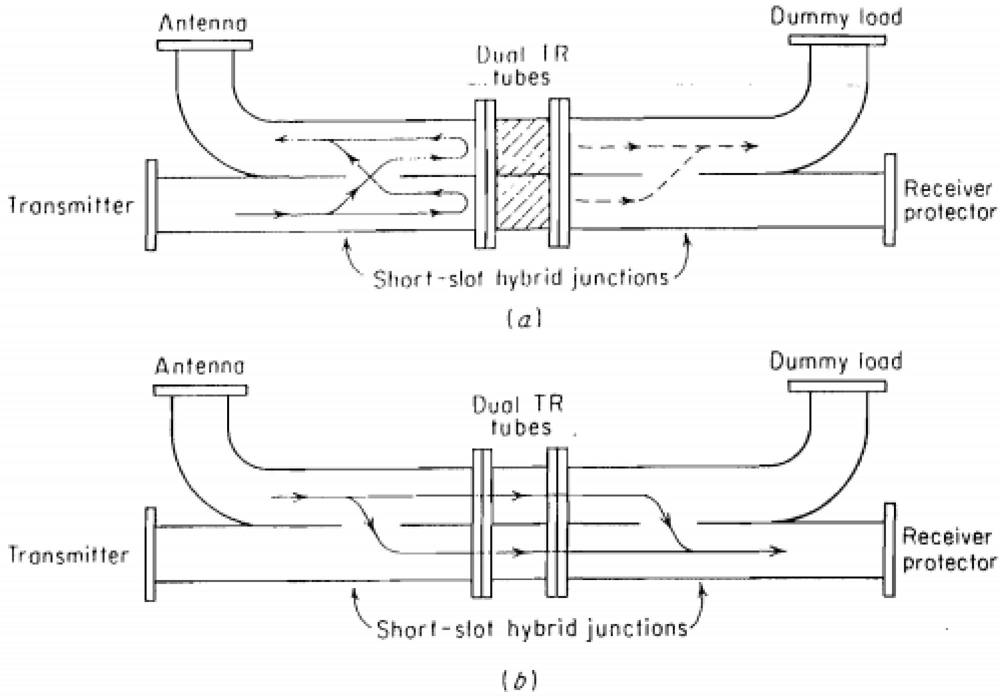


Fig.2. Balanced duplexer using dual TR tubes and two short-slot hybrid junctions (a) Transmit condition; (b) receive condition.

The short-slot hybrid has the property that each time the energy passes through the slot in either direction, its phase is advanced 90° . Therefore, the energy must travel as indicated by the solid lines. Any energy which leaks through the TR tubes (shown by the dashed lines) is directed to the arm with the matched dummy load and not to the receiver, In addition to the attenuation provided by the TR tubes, the hybrid junctions provide an additional 20 to 30 dB of isolation.

On reception the TR tubes are unfired and the echo signals pass through the duplexer and into the receiver as shown in Fig. 2 b. The power splits equally at the first junction and because of the 90° phase advance on passing through the slot, the energy recombines in the receiving arm and not in the dummy-load arm. The power-handling capability of the balanced duplexer is inherently greater than that of the branch-type duplexer and it has wide bandwidth, over ten percent with proper design. A receiver protector is usually inserted between the duplexer and the receiver for added protection.

b. What are Radar displays? Explain their principle of Operation with examples and sketches.

Answer:

The A-scope display, shown in the figure, presents only the range to the target and the relative strength of the echo. Such a display is normally used in weapons control radar systems. The bearing and elevation angles are presented as dial or digital readouts that correspond to the actual physical position of the antenna. The A-scope normally uses an electrostatic-deflection crt. The sweep is produced by applying a sawtooth voltage to the horizontal deflection plates. The electrical length (time duration) of the sawtooth voltage determines the total amount of range displayed on the crt face.

B-scope displays were common in airborne and fire-control radars in the 1950s and 60s, which were mechanically or electronically scanned from side to side, and sometimes up and down as well. The center of the bearing usually is movable through hand wheels in fire-control radars. The antenna turntable then is turned into the new direction. The screens middle is defined as the main reception direction of the antenna normally. The bearing area is covered through an electro-mechanical or electronic beam steering.

The used designation “B-scope” is ambiguous sometimes. The term refers to two completely different types of scopes. In radar devices without measurement of the azimuth angle, the term “B-scope” is also used (e.g.: Ground Penetration Radars). The abscissa is a time coded scale then, and shows a history of the pulse periods.

The PPI-scope shown in this figure, is by far the most used radar display. It is a polar coordinate display of the area surrounding the radar platform. Own position is represented as the origin of the sweep, which is normally located in the center of the scope, but may be offset from the center on some sets. The ppi uses a radial sweep pivoting about the center of the presentation. The sweep rotates on the display just as fast as the radar antenna. This results in a map-like picture of the area covered by the radar beam. A long-persistence screen is used so that the targets remain visible until the sweep passes again.

Q.8 a. Describe the principle behind the operation of a phased array antenna in a radar system. Explain its radiation pattern.

Answer:

Phased Array Antenna:

The phased array is a directive antenna made up of individual radiating antennas, or elements, which generate a radiation pattern whose shape and direction is determined by

the relative phases and amplitudes of the currents at the individual elements. By properly varying the relative phases it is possible to steer the direction of the radiation. The radiating elements might be dipoles open-ended waveguides, slots cut in waveguide, or any other type of antenna. The inherent flexibility offered by the phased-array antenna in steering the beam by means of electronic control is what has made it of interest for radar. It has been considered in those radar applications where it is necessary to shift the beam rapidly from one position in space to another or where it is required to obtain information about many targets at a flexible, rapid data rate. The full potential of a phased-array antenna requires the use of a computer that can determine in real time, on the basis of the actual operational situation, how best to use the capabilities offered by the array.
Radiation pattern for Phased array Antenna: Consider a linear array made up of N elements equally spaced a distance d apart (Fig.). The elements are assumed to be isotropic point sources radiating uniformly in all directions with equal amplitude and phase. Although isotropic elements are not realizable in practice, they are a useful concept in array theory, especially for the computation of radiation patterns. The array is shown as a receiving antenna for convenience, but because of the reciprocity principle, the results obtained apply equally well to a transmitting antenna. The outputs of all the elements are summed via lines of equal length to give a sum output voltage Ea. Element 1 will be taken as the reference signal with zero phase.

The difference in the phase of the signals in adjacent elements is $\Psi = 2\Pi (d/\lambda) \sin\theta$, where θ is the direction of the incoming radiation. It is further assumed that the amplitudes and phases of the signals at each element are weighted uniformly. Therefore the amplitudes of the voltages in each element are the same and, for convenience, will be taken to be unity. The sum of all the voltages from the individual elements, when the phase difference between adjacent elements is Ψ, can be written

$$E_a = \sin \omega t + \sin(\omega t + \psi) + \sin(\omega t + 2\psi) + \cdots + \sin[\omega t + (N-1)\psi]$$

where ω is the angular frequency of the signal. The sum can be written

$$E_a = \sin\left[\omega t + (N-1)\frac{\psi}{2}\right] \frac{\sin(N\psi/2)}{\sin(\psi/2)}$$

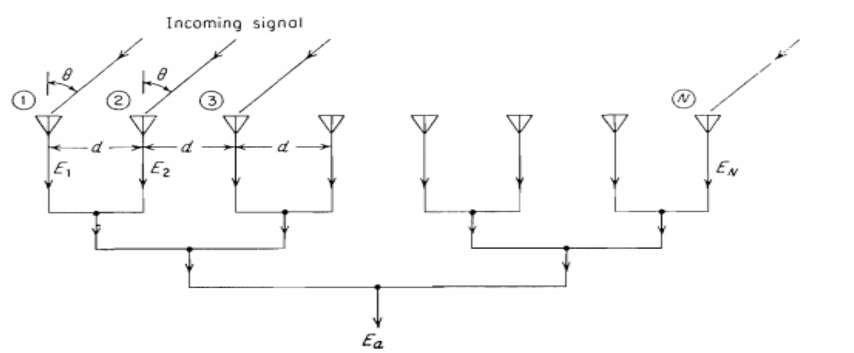


Fig. N-element linear array

The first factor is a sine wave of frequency ω with a phase shift (N - l)Ψ/2 (if the phase reference were taken at the center of the array, the phase shift would be zero), while the second term represents an amplitude factor of the form sin (NΨ/2)/sin(Ψ/2). The field intensity pattern is the magnitude of the Eq. above or

$$|E_a(\theta)| = \left|\frac{\sin\,[N\pi(d/\lambda)\sin\theta]}{\sin\,[\pi(d/\lambda)\sin\theta]}\right|$$

The pattern has nulls (zeros) when the numerator is zero. The latter occurs when NΠ (d/λ) sinθ = 0, ± Π, ± 2Π..., ± nΠ, where n = integer. The denominator, however, is zero when Π (d/λ) sinθ, = 0, ± Π, ± 2Π..., ± nΠ. Note that when the denominator is zero, the numerator is also zero. The value of the field intensity pattern is indeterminate when both the denominator and numerator are zero. However, by applying L'Hopital's rule (differentiating numerator and denominator separately) it is found that |Ea| is a maximum whenever sinθ = ± nλ/d. These maxima all have the same value and are equal to N. The maximum at sinθ = 0 defines the main beam. The other maxima are called grating lobes. They are generally undesirable and are to be avoided. If the spacing between elements is a half-wavelength (d/λ = 0.5), the first grating lobe (n = ± 1) does not appear in real space since sinθ > 1, which cannot be. Grating lobes appear at ± 90° when d = λ. For a non scanning array (which is what is considered here) this condition (d = λ) is usually satisfactory for the prevention of grating lobes. Equation discussed above applies to isotropic radiating elements, but practical antenna elements that are designed to maximize the radiation at θ = 0°, generally have negligible radiation in the direction θ = ± 90°. Thus the effect of a realistic element pattern is to suppress the grating lobes at ±90°. It is for this reason that an element spacing equal to one wavelength can be tolerated for a non scanning array.

From Eq. above, Ea (θ) = Ea (Π - θ). Therefore an antenna of isotropic elements has a similar pattern in the rear of the antenna as in the front. The same would be true for an array of dipoles. To avoid ambiguities, the backward radiation is usually eliminated by placing a reflecting screen behind the array. Thus only the radiation over the forward half of the antenna (-90° ≤ θ ≤ 90°) need be considered.

The radiation pattern is equal to the normalized square of the amplitude, or

$$G_a(\theta) = \frac{|E_a|^2}{N^2} = \frac{\sin^2\,[N\pi(d/\lambda)\sin\theta]}{N^2\sin^2\,[\pi(d/\lambda)\sin\theta]}$$

If the spacing between antenna elements is λ/2 and if the sine in the denominator of Eq. above is replaced by its argument, the half-power beamwidth is approximately equal to

$$\theta_B = \frac{102}{N}$$

The first sidelobe, for N sufficiently large, is 13.2 dB be1ow the main beam. The pattern of a uniformly illuminated array with elements spaced λ/2 apart is similar to the pattern produced by a continuously illuminated uniform aperture.

When directive elements are used, the resultant array antenna radiation pattern is

$$G(\theta) = G_e(\theta)\,\frac{\sin^2[N\pi(d/\lambda)\sin\theta]}{N^2\sin^2[\pi(d/\lambda)\sin\theta]} = G_e(\theta)G_a(\theta)$$

where Ge (θ) is the radiation pattern of an individual element. The resultant radiation pattern is the product of the element factor Ge(θ) and the array factor Ga(θ), the latter being the pattern of an array composed of isotropic elements. The array factor has also been called the space factor. Grating lobes caused by a widely spaced array may therefore be eliminated with directive elements which radiate little or no energy in the directions of the undesired lobes. For example, when the element spacing d = 2λ, grating lobes occur at θ = ±30° and ±90° in addition to the main beam at θ = 0°. If the individual elements have a beamwidth somewhat less than 60°, the grating lobes of the array factor will be suppressed.

b. List out the applications of phased array antennas.

Answer:

The phased-array antenna has been of considerable interest to the radar systems engineer because its properties are different from those of other microwave antennas. The array antenna takes several forms:

Mechanically scanned array:

The array antenna in this configuration is used to form a fixed beam that is scanned by mechanical motion of the entire antenna. No electronic beam steering is employed. This is an economical approach to air-surveillance radars at the lower radar frequencies, such as VHF. It is also employed at higher frequencies when a precise aperture illumination is required, as to obtain extremely low sidelobes. At the lower frequencies, the array might be a collection of dipoles or Yagis, and at the higher frequencies the array might consist of slotted waveguides.

Linear array with frequency scan:

The frequency-scanned, linear array feeding a parabolic cylinder or a planar array of slotted waveguides has seen wide application as a 3D air-surveillance radar. In this application, a pencil beam is scanned in elevation by use of frequency and scanned in azimuth by mechanical rotation of the entire antenna.

Linear array with phase scan:

Electronic phase steering, instead of frequency scanning, in the 3D air-surveillance radar is generally more expensive, but allows the use of the frequency domain for purposes other than beam steering. The linear array configuration is also used to generate multiple, contiguous fixed beams (stacked beams) for 3D radar. Another application is to use either phase- or frequency steering in a stationary linear array to steer the beam in one angular coordinate, as for the GCA radar.

Phase-frequency planar array:

A two-dimensional (planar) phased array can utilize frequency scanning to steer the beam in one angular coordinate and phase shifters to steer in the orthogonal coordinate. This approach is generally easier than using phase shifters to scan in both coordinates, but as with any frequency-scanned array the use of the frequency domain for other purposes is limited when frequency is employed for beam-steering.

Phase-phase planar array:
The planar array which utilizes phase shifting to steer the beam in two orthogonal coordinates is the type of array that is of major interest for radar application because of its inherent versatility. Its application, however, has been limited by its relatively high cost. The phase-phase array is what is generally implied when the term electronically steered phased array is used.

Q.9 a. Explain the block diagram of the AGC portion of tracking radar receiver.

Answer:
The echo-signal amplitude at the tracking-radar receiver will not be constant but will vary with time. The three major causes of variation in amplitude are (1) the inverse-fourth-power relationship between the echo signal and range, (2) the conical- scan modulation (angle-error signal), and (3) amplitude fluctuations in the target cross section. The function of the automatic gain control (AGC) is to maintain the d-c level of the receiver output constant and to smooth or eliminate as much of the noise like amplitude fluctuations as possible without disturbing the extraction of the desired error signal at the conical-scan frequency.

One of the purposes of AGC in any receiver is to prevent saturation by large signals. The scanning modulation and the error signal would he lost if the receiver were to saturate. In the conical-scan tracking radar an AGC that maintains the d-c level constant results in an error signal that is a true indication of the angular pointing error. The d-c level of the receiver must be maintained constant if the angular error is to be linearly related to the angle-error signal voltage.

An example of the AGC portion of a tracking-radar receiver is shown in Fig. A portion of the video-amplifier output is passed through a low-pass or smoothing filter and fed back to control the gain of the IF amplifier. The larger the video output, the larger will be the feedback signal and the greater will be the gain reduction. The filter in the AGC loop should pass all frequencies from direct current to just below the conical-scan-modulation frequency. The loop gain of the AGC filter measured at the conical-scan frequency should be low so that the error signal will not be affected by AGC action.

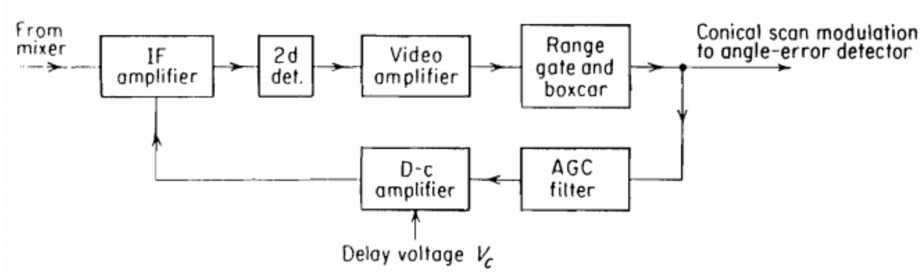


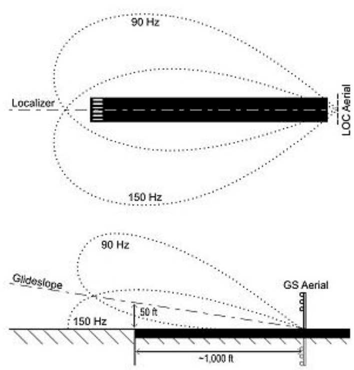
Fig Block diagram of the AGC portion of a tracking-radar receiver

The output of the feedback loop will be zero unless the feedback voltage exceeds a prespecified minimum value Vc. In the block diagram the feedback voltage and the voltage Vc are compared in the d-c amplifier. If the feedback voltage exceeds Vc, the AGC is operative, while if it is less, there is no AGC action. The voltage Vc is called the delay voltage. The purpose of the delay voltage is to provide a reference for the constant output signal and permit receiver gain for weak signals. If the delay voltage were zero, any output which might appear from the receiver would be due to the failure of the AGC circuit to regulate completely. In many applications of AGC the delay voltage is actually zero. This is called undelayed AGC. In such cases the AGC can still perform satisfactorily since the loop gain is usually low for small signals. Thus the AGC will not regulate weak signals. The effect is similar to having a delay voltage, but the performance will not be as good.

b. What is an instrument landing system? Explain how elevation guidance is provided in this system. Give the configuration of localizer antenna also.

Answer:

An ILS consists of two independent sub-systems, one providing lateral guidance (localizer), the other vertical guidance (glide slope or glide path) to aircraft approaching a runway. Aircraft guidance is provided by the ILS receivers in the aircraft by performing a modulation depth comparison.



The emission patterns of the localizer and glideslope signals. Note that the glide slope beams are partly formed by the reflection of the glideslope aerial in the ground plane.

A localizer (LOC, or LLZ until ICAO designated LOC as the official acronym) antenna array is normally located beyond the departure end of the runway and generally consists of several pairs of directional antennas. Two signals are transmitted on one out of 40 ILS channels between the carrier frequency range 108.10 MHz and 111.95 MHz (with the 100 kHz first decimal digit always odd, so 108.10, 108.15, 108.30, and so on are LOC frequencies but 108.20, 108.25, 108.40, and so on are not). One is modulated at 90 Hz, the other at 150 Hz and these are transmitted from separate but co-located antennas. Each antenna transmits a narrow beam, one slightly to the left of the runway centerline, the other to the right.

The localizer receiver on the aircraft measures the difference in the depth of modulation (DDM) of the 90 Hz and 150 Hz signals. For the localizer, the depth of modulation for each of the modulating frequencies is 20 percent. The difference between the two signals varies depending on the position of the approaching aircraft from the centerline.

If there is a predominance of either 90 Hz or 150 Hz modulation, the aircraft is off the centerline. In the cockpit, the needle on the horizontal situation indicator (HSI, the instrument part of the ILS), or course deviation indicator (CDI), will show that the aircraft needs to fly left or right to correct the error to fly down the center of the runway. If the DDM is zero, the aircraft is on the centerline of the localizer coinciding with the physical runway centerline.

A glide slope (GS) or glide path (GP) antenna array is sited to one side of the runway touchdown zone. The GP signal is transmitted on a carrier frequency between 328.6 and 335.4 MHz using a technique similar to that of the localizer. The centerline of the glide slope signal is arranged to define a glide slope of approximately 3° above horizontal (ground level). The beam is 1.4° deep; 0.7° below the glideslope centerline and 0.7° above the glideslope centerline.

These signals are displayed on an indicator in the instrument panel. This instrument is generally called the omni-bearing indicator or nav indicator. The pilot controls the aircraft so that the indications on the instrument (i.e., the course deviation indicator) remain centered on the display. This ensures the aircraft is following the ILS centreline (i.e., it provides lateral guidance). Vertical guidance, shown on the instrument by the glideslope indicator, aids the pilot in reaching the runway at the proper touchdown point. Many aircraft possess the ability to route signals into the autopilot, allowing the approach to be flown automatically by the autopilot.

TEXT BOOKS

I Introduction to Radar Systems, Merrill I. Skolnik, 3e, TMH, 2001

II Electronic and Radio Engineering, F.E. Terman, McGraw Hill Publications.