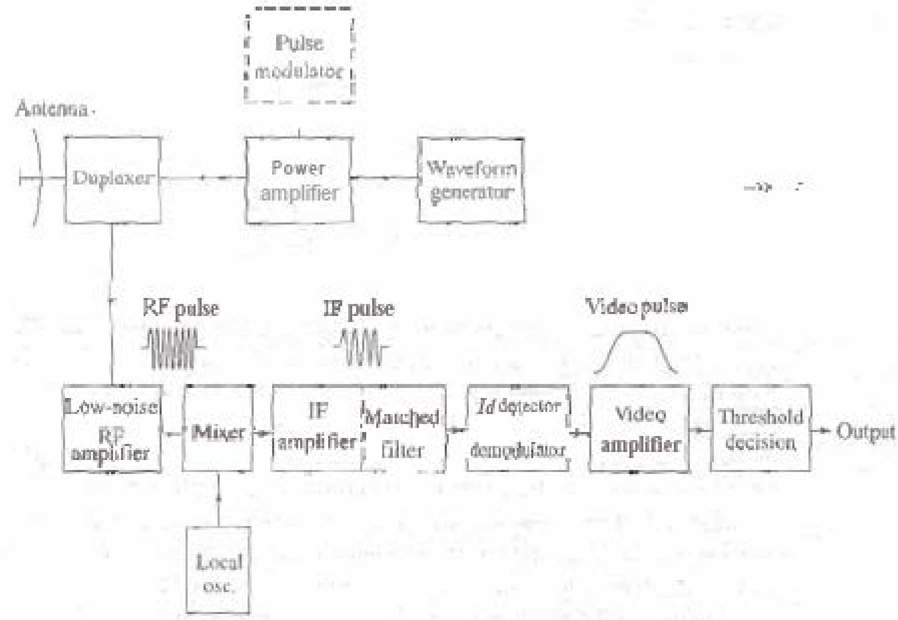
Q.2 a. Draw the block diagram of radar and explain the working of its each block. (10)

Answer:

The operation of a pulse radar may be described with the aid of the simple block diagram of Fig. 1.4. The transmitter may be a *power amplifier,* such as the klystron, traveling wave tube, or transistor amplifier. It might also be a power oscillator, such as the magnetron. The magnetron oscillator has been widely used for pulse radars of modest capability; but the amplifier is preferred when high average power is necessary, when other than simple pulse waveforms are required (as in pulse compression), or when good performance is needed in detecting moving targets in the midst of much larger clutter echoes based on the doppler frequency shift (the subject of Chap. 3). A power amplifier is indicated in Fig. 1.4. The radar signal is produced at low power by a *waveform generator,* which is then the input to the power amplifier. In most power amplifiers, except for solid-state power sources, a modulator (Sec. 10.7) turns the transmitter on and off in synchronism with the input pulses. When a power oscillator is used, it is also turned on and off by a *pulse modulator* to generate a pulse waveform.

The output of the transmitter is delivered to the *antenna* by a waveguide or other form of transmission line, where it is radiated into space. Antennas can be mechanically steered parabolic reflectors, mechanically steered planar arrays, or electronically steered phased arrays (Chap. 9). On transmit the parabolic reflector focuses the energy into a narrow

Figure 1.4 Block diagram of a conventional pulse radar with a superheterodyne receiver.



beam, just as does an automobile headlight or a searchlight. A phased array antenna is a collection of numerous small radiating elements whose signals combine in space to produce a radiating plane wave. Using phase shifters at each of the radiating elements, an electronically steered phased array can rapidly change the direction of the antenna beam in space without mechanically moving the antenna. When no other information is available about the antenna, the beamwidth (degrees) of a "typical" parabolic reflector is often approximated by the expression $65\ \lambda/D$, where D is the dimension of the antenna in the same plane as the beamwidth is measured, and λ is the radar wavelength. For example, an antenna with a horizontal dimension $D = 32.5$ wavelengths has an azimuth beamwidth of 2°. At a frequency of 3 GHz ($\lambda = 10$ cm), the antenna would be 3.25 m, or 10.7 ft, in extent. The rotation of a surveillance radar antenna through 360° in azimuth is called an antenna *scan*. A typical scan rate (or rotation rate) for a long-range civil air-traffic control air-surveillance radar might be 6 rpm. Military air-surveillance radars generally require a higher rotation rate.

The *duplexer* allows a single antenna to be used on a time-shared basis for both transmitting and receiving. The duplexer is generally a gaseous device that produces a short circuit (an arc discharge) at the input to the receiver when the transmitter is operating, so that high power flows to the antenna and not to the receiver. On reception, the duplexer directs the echo signal to the receiver and not to the transmitter. Solid-state ferrite circulators and receiver protector devices, usually solid-state diodes, can also be part of the duplexer.

The receiver is almost always a *superheterodyne*. The input, or RF,* *stage* can be a low-noise transistor amplifier. The mixer and local oscillator (LO) convert the RF signal

*In electrical engineering, *RF* is an abbreviation for *radio frequency*; but in radar practice it is understood to mean *radar frequency*. RF also is used to identify that portion of the radar that operates at RF frequencies, even though the inclusion of "frequencies" in this expression might seem redundant.

to an intermediate frequency (IF) where it is amplified by the IF amplifier. The signal bandwidth of a superheterodyne receiver is determined by *the* bandwidth of its IF stage. The IF frequency, for example. might be 30 or *60* MHz when the pulse width is of the order of 1 μs. (With a 1-μs pulse width. the IF bandwidth would be about 1 MHz.) The IF amplifier is designed as a *matched filter* (Sec. 5.2); that is, ane which maximizes the output peak-signal-to-mean-noise ratio. Thus the matched filter maximizes the detectability of weak echo signals and attenuates unwanted signals. With the approximately rectangular pulse shapes commonly used in many radars, conventional radar receiver filters are close to that of a matched filter when the receiver bandwidth B is the inverse of the pulse width τ, or $B\tau \approx 1$.

Sometimes the low-noise input stage is omitted and the mixer becomes the first stage of the receiver. A receiver with a mixer as the input stage will be less sensitive because of the mixer's higher noise figure; but it will have greater dynamic range, less susceptibility to overload, and less vulnerability to electronic interference than a receiver with a low-noise first stage (Sec. 11.3). These attributes of a mixer input stage might be of interest for military radars subject to the noisy environment of hostile electronic countermeasures (ECM).

l'hc IF amplifier is followed by a crystal diode, which is traditionally called the *second detector*, or *demodulator*. Its purpose is to assist in extracting the signal modulation from the carrier. The combination of IF amplifier, second detector, and video amplifier act as an *envelope detector* to pass the pulse modulation (envelope) and *reject* the carrier frequency. In radars that detect the doppler shift of the echo signal, the envelope detector is replaced by a *phase detector* (Sec. 3.1), which is different from the envelope detector shown here. The combination of IF amplifier and video amplifier is designed to provide sufficient amplification, or gain, to raise the level of the input signal to a magnitude where it can be seen on a display, such as a cathode-ray tube (CRT), or be the input to a digital computer for further processing.

At the output of the receiver a decision is made whether or not a target is present. The decision is based on the magnitude of the receiver output. If the output is large enough to exceed a predetermined threshold. the decision is that a target is present. If it does not cross the threshold, only noise is assumed to be present. The threshold level is set so that the rate at which false alarms occur due to noise crossing the threshold (in the absence of signal) is below some specified, tolerable value. This is fine if the noise remains constant, as when receiver noise dominates. If, on the other hand, the noise is external to the radar (as from unintentional interference or from deliberate noise jamming) or if clutter echoes (from the natural environment) are larger than the receiver noise, the threshold has to be varied adaptively in order to maintain the false alarm rate at a constant value. This is accomplished by a *constant false alarm* rare (CFAR) receiver (Sec. 5.7).

A radar usually receives many echo pulses from a target. The process of adding these pulses together to obtain a greater signal-to-noise ratio before the detection decision is made is called *integration*. The integrator is often found in the video portion of the receiver.

The *signal processor* is that part of the radar whose function is to pass the desired echo signal and reject unwanted signals, noise. or clutter. The signal processor is found in the receiver before the detection decision is made. The matched filter, mentioned previously, is an example of a signal processor. Another example is the doppler filter that separates desired moving targets (whose echoes are shifted in frequency due to the doppler effect) from undesired stationary clutter echoes.

Some radars process the detected target signal further, in the *data processor*, before displaying the information to an operator. An example is an *automatic tracker*, which uses the locations of the target measured over a period of time to establish the track (or path) of the target. Most modern air-surveillance radars and some surface-surveillance radars generate target tracks as their output rather than simply display detections. Following the data processor, or the decision function if there is no data processor, the radar output is displayed to an operator or used in a computer or other automatic device to provide some further action.

The signal processor and data processor are usually implemented with digital technology rather than with analog circuitry. The analog-to-digital (A/D) converter and digital memory are therefore important in modern radar systems. In some sophisticated radars in the past, the signal and data processors were larger and consumed more power than the transmitter and were a major factor in determining the overall radar system reliability; but this should not be taken as true in all cases.

b. What is meant by maximum unambiguous range & range to a target? (6)

Answer:

Range to a Target The most common radar signal, or waveform, is a series of short-duration, somewhat rectangular-shaped pulses modulating a sinewave carrier. (This is sometimes called a *pulse train.*) The range to a target is determined by the time T_R it takes the radar signal to travel to the target and back. Electromagnetic energy in free space travels with the speed of light, which is $c = 3 \times 10^8$ m/s. Thus the time for the signal to travel to a target located at a range R and return back to the radar is $2R/c$. The range to a target is then

$$R = \frac{cT_R}{2} \qquad [1.1]$$

*Webster's New Collegiate Dictionary defines *range* as "the horizontal distance to which a projectile can be pro- "the horizontal distance between a weapon and target." This is not how the term is used in radar. On the other had, the dictionary defines *range finder* as "an instrument . . . to determine the distance to a target," which is its meaning in radar.

With the range in kilometers or in nautical miles, and T in microseconds, Eq. (1.1) becomes:

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

Each microsecond of round-trip travel time corresponds to a distance of 150 meters, 164 yards, 492 feet, 0.081 nautical mile, or 0.093 statute mile. It takes 12.35 μs for a radar signal to travel a nautical mile and back.

Maximum Unambiguous Range Once a signal is radiated into space by a radar, sufficient time must elapse to allow all echo signals to return to the radar before the next pulse is transmitted. The rate at which pulses may be transmitted, therefore, is determined by the longest range at which targets are expected. If the time between pulses T_p is too short, an echo signal from a long-range target might arrive *after* the transmission of the next pulse and be mistakenly associated with that pulse rather than the actual pulse transmitted earlier. This can result in an incorrect or ambiguous measurement of the range. Echoes that arrive after the transmission of the next pulse are called *second-time-around echoes* (or *multiple-time-around echoes* if from even earlier pulses). Such an echo would appear to be at a closer range than actual and its range measurement 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 the *maximum unambiguous range*, R_{un}, and is given by

$$R_{un} = \frac{cT_p}{2} = \frac{c}{2f_p} \qquad [1.2]$$

where T_p = pulse repetition period = $1/f_p$, and f_p = pulse repetition frequency (prf), usually given in hertz or pulses per second (pps). A plot of the maximum unambiguous range as a function of the pulse repetition frequency is shown in Fig. 1.2. The term *pulse repetition rate* is sometimes used interchangeably with *pulse repetition frequency*.

Q.3 a. Derive an equation to show the relationship between maximum radar range and antenna gain. (8)

Answer:

Almost all radars use *directive antennas* with relatively narrow beamwidths that direct the energy in a particular direction. The antenna is an important part of a radar. As was found from the derivation of the radar equation in Sec. 1.2, it serves to place energy on target during transmission, collect the received echo energy reflected from the target, and determine the angular location of the target. There is always a trade between antenna size and transmitter size when long-range performance is required. If one is small the other must be large to make up for it. This is one reason why large antennas are generally desirable in most radar applications when practical considerations do not limit their physical size. Thus far, the antenna has been thought of as a mechanically steered reflector. Radar antennas can also be electronically steered or mechanically steered phased arrays, as described in Chap. 9.

Antenna Gain The antenna gain $G(\theta,\phi)$ is a measure of the power per unit solid angle radiated in a particular direction by a directive antenna compared to the power per unit solid angle which would have radiated by an omnidirectional antenna with 100 percent efficiency. The gain of an antenna is

$$G(\theta,\phi) = \frac{\text{power radiated per unit solid angle at an azimuth } \theta \text{ and an elevation } \phi}{(\text{power accepted by the antenna from the transmitter})/4\pi} \qquad [2.53]$$

This is the *power gain* and is a function of direction. If it is greater than unity in some directions, it must be less than unity in other directions. There is also the *directive gain*, which has a similar definition except that the denominator is the power *radiated* by the antenna per 4π steradians rather than the power *accepted* from the transmitter. The difference between the two is that the power gain accounts for losses within the antenna. The power gain is more appropriate for the radar equation than the directive gain, although there is usually little difference between the two in practical radar antennas, except for the phased array. The power gain and the directive gain of a radar antenna are usually considered to be the same in this text. When they are significantly different, then the distinction between the two must be made. In the radar equation, it is the maximum power gain that is meant by the parameter G.

b. List and explain some system losses. (8)

Answer:

At the beginning of this chapter it was said that one of the important factors omitted from the simple radar equation was the loss that occurs throughout the radar system. The loss due to the integration of pulses and the loss due to a target with a fluctuating cross section have been already encountered in this chapter. Propagation losses in the atmosphere are considered later, in Chap. 8. This section considers the various *system losses*, denoted L_s, not included elsewhere in the radar equation. Some system losses can be predicted beforehand (such as losses in the transmission line); but others cannot (such as degradation when operating in the field). The latter must be estimated based on experience and experimental observations. They are subject to considerable variation and uncertainty. Although the loss associated with any one factor may be small, there can be many small effects that add up and result in significant total loss. The radar designer, of course, should reduce known losses as much as possible in the design and development of the radar. Even with diligent efforts to reduce losses, it is not unusual for the system loss to vary from perhaps 10 dB to more than 20 dB. (A 12-dB loss reduces the range by one-half.)

All numerical values of loss mentioned in this section, including the above values of system loss, are meant to be illustrative. They can vary considerably depending on the radar design and how the radar is maintained.

System loss, L_s (a number greater than one), is inserted in the denominator of the radar equation. It is the reciprocal of efficiency (number less than one). The two terms (loss and efficiency) are sometimes used interchangeably.

Microwave Plumbing Losses There is always loss in the transmission line that connects the antenna to the transmitter and receiver. In addition, there can be loss in the various microwave components, such as the duplexer, receiver protector, rotary joint, directional couplers, transmission line connectors, bends in the transmission lines, and mismatch at the antenna.

Transmission Line Loss The theoretical one-way loss in decibels per 100 ft for standard waveguide transmission lines is shown in Table 2.2.[59] Since the same transmission line

Table 2.2 Attenuation of Rectangular Waveguides*

Frequency Band	EIA Waveguide Designation†	Frequency Range (GHz) for Dominant TE_{10} Mode	Outer Dimensions and Wall Thickness, inches	Theoretical Attenuation, Lowest to Highest Frequency, dB/100 ft (one-way)
UHF	WR-2100	0.35–0.53	21.25 × 10.75 × 0.125	0.054–0.034
L band	WR-770	0.96–1.45	7.95 × 4.1 × 0.125	0.201–0.136
S band	WR-284	2.6–3.95	3.0 × 1.5 × 0.08	1.102–0.752
C band	WR-187	3.95–5.85	2.0 × 1.0 × 0.064	2.08–1.44
X band	WR-90	8.2–12.40	1.0 × 0.5 × 0.05	6.45–4.48
K_u band	WR-62	12.4 18.0	0.702 × 0.391 × 0.04	9.51–8.31
Ka band	WR-28	26.5–40.0	0.36 × 0.22 × 0.04	21.9–15.0

*After "Reference Data for Engineers," 8th ed., M. E. Van Valkenburg, editor-in-chief, Chap. 30, *Waveguides and Resonators*, by T. Itoh, SAMS Prentice Hall Computer Publishing, Carmel, Indiana, 1993.

†UHF and *L*-band guides are made of aluminum, K_u band is silver, the rest are copper-zinc alloy.

generally is used for both transmission and reception, the lass to be inserted in the radar equation is twice the one-way loss. Flexible waveguides and coaxial lines can have higher losses than conventional waveguides. At the lower radar frequencies, the transmission line introduces little loss unless its length is exceptionally long. At h e higher frequencies, attenuation may *not* always be small and may have to be taken into account. When practical, the transmitter and receiver should be placed close to the antenna to keep *the* transmission-line loss small. Additional loss can occur at each connection or bend in the line, Connector losses are normally negligible, but if the connection is poorly made, it can contribute measurable attenuation.*

Duplexer Loss The loss due to a gas duplexer that protects the receiver from the high power of the transmitter is generally different on transmission and reception. It also depends, of course, on the type of duplexer used, Manufacturers' catalogs give values for a duplexer's *insertion Oss* and (for a gas duplexer) the *arc loss* when in the fired condition. The radar might also have a waveguide shutter, with some insertion loss, that closes when the radar is shut down so as *to* protect the receiver from extraneous high-power signals whw its duplexer is not activated. A solid-stale receiver protector is often used as well as a solid-state attenuator in the receiver transmission line for applying sensitivity time control (STC). The duplexer and other related devices that might be used could, in some cases, contribute more then 2 dB of two-way loss.

*At a particular radar laboratory many years ago, an old *L*-band air-surveillance radar was used as an experimental test-bed system. Its range was poor, and the engineers attributed this to its "age"—whatever that meant, Its poor performance was tolerated for many years. One day, a technician working near the transmission line to the antenna happened to find, accidentally, that one of the transmission-line connectors had not been properly secured. He tightened a few bolts a d the radar "miraculously" achieved a significant increase in performance. Sometimes, it's the link things that count!

Example Although each radar can have different losses, an *S*-band (3-GHz) radar might have, by way of illustration, two-way microwave plumbing losses as follows:

100 ft of RG-113/U aluminum waveguide line	1.0 dB
Duplexer and related devices	2.0 dB
Rotary joint	0.8 dB
Connectors and bends (estimate)	0.3 dB
Other RF devices	0.4 dB
Total "example" microwave plumbing loss	4.5 dB

Antenna Losses The antenna efficiency, discussed in Chap. 9, is not included as a system loss. It should be accounted for in the antenna gain. Shaping of the antenna pattern, for example, to provide a $\csc^2$ pattern (Sec. 2.11), results in a loss that is included as an additional lowering of the antenna gain (Sec. 9.11) rather than as a system loss. The beam-shape loss of a surveillance radar, however, is usually included as part of the system losses.

Beam-Shape Loss The antenna gain that appears in the radar equation is assumed to be a constant equal to the maximum value. But in reality the train of pulses returned from a target by a scanning antenna is modulated in amplitude by the shape of the antenna beam, Fig. 2.26. Only one out of *n* pulses has the maximum antenna gain *G*, that which occurs when the peak of the antenna beam is in the direction of the target. Thus the computations of probability of detection (as given earlier in this chapter) have to take account of an amplitude-modulated train of pulses rather than constant-amplitude pulses. Some published probability of detection computations and computer programs for the radar equation account for the beam-shape loss. Others do not. When using published values of detection probability one needs to determine whether the beam-shape effect has been included or whether it must be accounted for separately. In this text, the approach is to

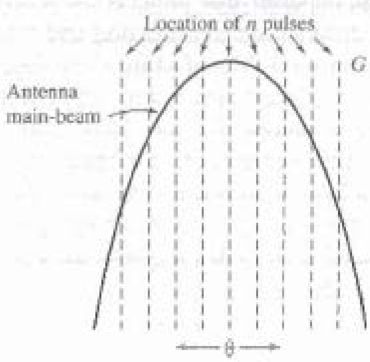


Figure 2.26 Nature of the beam-shape loss. The simple radar equation assumes *n* pulses are integrated, all with maximum antenna gain *G*. The dashed lines represent *n* pulses with maximum gain, and the solid curve is the antenna main-beam pattern $G(\theta)$. Except for the pulse at the center of the beam, the actual pulses illuminate the target with a gain less than the maximum.

assume a constant-amplitude pulse train as determined by the maximum antenna gain, and then add a beam-shape loss to the total system losses in the radar equation. This is a simpler, albeit less accurate, method. It is based on calculating the reduction in total signal energy received from a modulated train of pulses compared to what would have been received from a constant-amplitude pulse train. As defined, it does not depend on the probability of detection.

To obtain the beam-shape loss, the one-way-power antenna pattern is approximated by a gaussian shape given by exp $[-2.78\,\theta^2/\theta_B^2]$, where θ is the angle measured from the center of the beam and θ_B is the half-power beamwidth. If n_B is the number of pulses received within the one-way half-power beamwidth θ_B, and n the total number of pulses integrated (n does not necessarily equal n_B), the beam-shape loss is

$$\text{Beam-shape loss} = \frac{n}{1 + 2\sum_{k=1}^{(n-1)/2} \exp\left[-5.55k^2/(n_B - 1)^2\right]} \qquad 12.591$$

This expression applies for an odd number of pulses with the middle pulse appearing at the beam maximum. For example, if $n = 11$ pulses are integrated, all lying uniformly between the 3-dB beamwidth ($n = n_B$), the beam-shape loss is about 2 dB.

The above applies to a fan beam, It also applies to a pencil beam if the target passes directly through the center of the beam. If the target passes through any other part of the pencil beam, the maximum signal will be reduced. The beam-shape loss is increased, therefore, by the square of the maximum antenna gain seen (if the antenna were to pass through the beam center) to the squaw of the maximum gain actually seen (when the antenna passes through other than the beam maximum). The ratio is the square because of the two-way radar propagation.

When a large number of pulses are integrated, the scanning loss was found by Blake[60] to be 1.6 dB for a fan beam scanning in one coordinate and 3.2 dB for a pencil beam scanning in two coordinates. Blake's values are commonly used as the beam-shape loss in the radar equation, unless the number of pulses integrated is small.

A similar loss must be taken into account when searching a volume with a step-scanning pencil beam antenna (as with a phased array) since not all regions of space are illuminated with the same value of antenna gain. (In *step scanning,* the antenna beam is stationary and dwells in a fixed direction until all n pulses are collected, and then rapidly switches to dwell in a new direction.) Some tracking radars, such as conical scan, also have a loss due to the antenna beam not illuminating the target with maximum gain.

Scanning Loss When the antenna scans rapidly enough, relative to the round-trip time of the echo signal, the antenna gain in the direction of the target on transmit might not be the same as that on receive. This results in an additional loss called the *scanning loss.* It can be important for some long-range scanning radars such as those designed for space surveillance or ballistic missile defense, rather than for most air-surveillance radars.

Radome Loss introduced by a radome (Sec. 9.17) will depend on the type and the radar frequency. A "typical" ground-based metal space-frame radome might have a two-way transmission loss of *1.2* dB at frequencies ranging from *L* to *X* band.[61] Air-supported radomes can have lower loss; the lass with dielectric space-frame radomes can be higher.

Phased Array Losses Some phased array radars have additional transmission-line losses due to the distribution network that connects the receiver and transmitter to each of the many elements of the array. These losses are more correctly included as a reduction in the antenna power gain, but they seldom are. When not included as a loss in antenna gain, they should be included under the system losses.

Signal Processing Losses Sophisticated signal processing is prevalent in modern radars and is very important for detecting targets in clutter and in extracting information from radar echo signals. Unfortunately, signal processing can introduce loss that has to be tolerated. The factors described below can introduce significant loss that has to be accounted for; doppler processing radars might have even greater loss.

Nonmatched Filter There can be from *0.5* to 1.0 dB of loss due to a practical, rather than ideal, matched filter (Sec. 5.2) A similar loss can occur with a pulse-compression filter (which is an example of a matched filter).

Constant False-Alarm Rate (CFAR) Receiver As mentioned in Sec. 5.7, this loss can be more than 2.0 dB depending on the type of CFAR.

Automatic Integrators The binary moving-window detector, for example, can have a theoretical bss of 1.5 to 2.0 dB (Sec. *5.6).* Other automatic integrators might have more or less loss.

Threshold Level A threshold is established at the output of *the* radar receiver to achieve some specified probability of false alarm or average false-alarm time (Sec. 2.5). Because of the exponential relationship between false-alarm time and the threshold level, the threshold might be set at a slightly higher level as a safety factor to prevent excessive false alarms. Depending on how accurately the threshold can be set and maintained, the loss might be only a small fraction of a dB.

Limiting Loss Some radars might use a limiter in the radar receiver. Ah example is pulse-compression 'processing to remove amplitude fluctuations in the signal. The so-called Dicke-fix, an electronic counter-countermeasure to reduce the effects of impulsive noise, employs a hard limiter. Early MTI radars used bard limiters, but this is now considered poor practice and is almost never used since they reduce the clutter attenuation that can be obtained (Sec. 3.7). Analysis of an ideal bandpass hard limiter shows that, for small signal-to-noise ratio, the reduction in the signal-to-noise ratio 'of a sinewave imbedded in narrowband gaussian noise is theoretically $\pi/4$, which is about 1 dB.[62] Hard limiting with some forms of pulse compression *can* introduce greater loss (Table 6.6).

Straddling Loss A loss, called the *range straddling loss,* occurs when range gates are not centered on the pulse or when, for practical reasons, they are wider than optimum.

Likewise in a doppler filter bank (Sec. 3.4) there can be a *filter straddling loss* when the signal spectral line is not centered on the filter. These occur in both analog and digital processing.

Sampling Loss When digital processing is employed, a related loss to the straddling loss can occur when the video signal after the matched filter is sampled prior to digitizing by the analog-to-digital (A/D) converter. If there is only one sample per pulse width, sampling might not occur at the maximum amplitude of the pulse. The difference between the sampled value and the maximum pulse amplitude represents a *sampling loss*. The loss is about 2 dB when the sampling is at a rate of once per pulse width (applies for a probability of detection of 0.90 and probability of false alarm of 10^{-6}).[63] The larger values occur with higher probability of detection. Decreasing the sampling interval rapidly decreases the loss. When two samples per pulse are taken, the loss is approximately 0.5 dB; and with three samples per pulse, it is under 0.2 dB.

Losses in Doppler-Processing Radar* When range and/or doppler ambiguities exist (as discussed in Sec. 2.10 and Chap. 3 for MTI and pulse doppler radars), multiple redundant waveforms may be used to resolve the ambiguities or to prevent "blind speeds." Compared to a radar that has no ambiguities the redundant waveforms can represent a significant loss. The use of redundant waveforms is seldom considered as a system loss; but it can certainly affect the range of a radar. For example, in some medium prf pulse doppler radars (Sec. 3.9), eight different waveforms, each with a different prf, might need to be transmitted so as to obtain at least three dwells (with no missed targets due to blind speeds) in order to resolve the range ambiguities. The represents a loss of signal of 8/3, or 4.3 dB.

There can be an eclipsing loss in pulse doppler radars when echoes from (ambiguous) multiple-time-around targets arrive back at the radar at the same time that a pulse is being transmitted. MTI doppler processing also introduces loss due to the shape of the doppler (velocity) filters if the target velocity does not correspond to the maximum response of the doppler filter. Fill pulses in MTI and pulse doppler radar sometimes are necessary, but they represent wasted pulses from the point of view of detection of signals in noise.

Losses due to doppler processing are not always included as part of the system losses. This is justified in an MTI radar by noting that the clutter that the doppler processing is designed to remove generally does not occur at the maximum range of a radar, which is where the range performance of a radar usually is determined. Nevertheless, it should be recognized that MTI radars that employ the doppler frequency shift to detect moving targets in the presence of large clutter echoes (as discussed in Chap. 3) can seriously reduce the ability of a radar to detect a target when no clutter is present. This is why MTI processing is disconnected at ranges beyond where clutter is not expected.

Fill pulses are sometimes used in an MTI radar when the pulses are processed in batches, as in the MTD radar of Sec. 3.6. They are also sometimes used with high prf

*Doppler processing is described in Chap. 3. The reader not familiar with MTI and pulse doppler radar can skip this subsection.

doppler radar to avoid the effects of multiple-time-around clutter echoes. They are necessary for performing some types of doppler processing, but they are wasted pulses when signal-to-noise, and not signal-to-clutter, is important. They have not usually been considered as introducing a system loss, but they could if the fill pulses affect the detection of signals in noise.

Collapsing Loss If the radar were to integrate additional noise samples along with signal-plus-noise pulses, the added noise would result in a degradation called the *collapsing loss.* An example is in a 3-D radar that has a "stack" of multiple independent pencil beams in elevation. If the outputs from the N beams are superimposed on a single PPI display, at a given range resolution cell that contains the target echo the display will add $N - 1$ noise samples along with the single target echo. A collapsing loss can also occur when the output of a high-resolution radar is shown on a display whose resolution is coarser than that inherent in the radar. If the radar receiver output is automatically processed and thresholded rather than rely on an operator viewing a display to make the detection decision, there need not be a collapsing loss in the above two examples.

The mathematical derivation of the collapsing loss, assuming a square-law detector, may be carried out as suggested by Marcum.[64] He has shown that the integration of m noise pulses along with n signal-to-noise pulses with signal-to-noise ratio per pulse $(S/N)_n$, is equivalent to the integration of $m + n$ signal-to-noise pulses each with signal-to-noise ratio $n(S/N)_n/(m + n)$. Thus the collapsing loss, $L_c(m,n)$, is equal to the ratio of the integration loss L_i (Sec. 2.6) for $m + n$ pulses to the integration loss for n pulses, or

$$L_c(m,n) = \frac{L_i(m+n)}{L_i(n)} \qquad [2.60]$$

For example, assume there are 10 signal-plus-noise pulses integrated along with 30 noise-only pulses, and that $P_d = 0.90$ and $P_{fa} = 1/n_f = 10^{-8}$. From Fig. 2.8a, $L_i(40) = 3.5$ dB and $L_i(10) = 1.7$ dB, so that the collapsing loss according to Eq. (2.60) is 1.8 dB.

The above applies for a square-law detector. Trunk[65] has shown that the collapsing loss for a linear detector can be much greater than that for the square-law detector when the number of pulses integrated is small and the collapsing ratio is large, where collapsing ratio is defined as $(m + n)/n$. As the number of pulses becomes large, the difference between the two detectors becomes less, especially for low values of collapsing loss.

Operator Loss Most modern high-performance radars provide the detection decision automatically without intervention of a human operator. Processed information is presented directly to an operator or to a computer for some other action. In the early days of radar, operators were depended upon to find targets on a display. Sometimes, when the radar range performance was less than predicted, the degradation of performance was attributed to an operator loss. As engineers began to learn more about radar and the performance of the operator, it was found that an alert, motivated, well-trained operator can perform as well as indicated by theory for electronic detection. For this reason, an operator loss factor is seldom included even if the operator makes the detection decision. (When an operator is used to make detection decisions from the output of a radar display, he or she should be replaced with a rested, alert operator every *20* to 30 min., or else performance can seriously degrade.)

Equipment Degradation It is not uncommon for radars operated under field conditions to have lower performance than when they left the factory. This loss of performance can be recognized and corrected by regularly testing the radar, especially with built-in test equipment that automatically indicates when equipment deviates from specifications. It is not possible to be precise about the amount of loss *to* be assigned to field degradation. From one *to* three dB might be used when no other information is available.

Propagation Effects The effect of the environment on the propagation of radar waves can be significant and can make the actual range considerably different from that predicted as if the radar were operated in free space. Propagation effects can increase the free-space range as well as decrease it. The major effects of propagation on radar performance are: (1) reflections from the earth's *surface,* which cause the breakup of the antenna elevation pattern into lobes; (2) refraction, or bending, of *the* propagating wave by the variation of the atmosphere's index of refraction as a function of altitude, which usually increases the radar's range; (3) propagation *in* atmospheric ducts, which can significantly increase the range at low altitudes; and (4) attenuation in the clear atmosphere or in precipitation, which usually is negligible at most radar frequencies.* Propagation effects are not considered part of the system losses. They are accounted for separately by a *propagation factor,* usually denoted as F^4 and, when appropriate, by an *attenuation factor* exp $1 - 24$, where a is the attenuation coefficient (nepers per unit distance†), and R is the range. (This assumes the attenuation coefficient is independent of range.) The factor F^4 mainly includes the effects of lobing of the elevation antenna pattern due to reflection from the earth's surface (Sec. 8.2), but it can include all other propagation effects except attenuation. Both the propagation and the attenuation factors, as written, are in the numerator of the radar equation.

The effect of the environment on radar propagation and performance is the subject of Chap. 8.

Radar ***System*** **Losses—the Seller and the Buyer** There is no universally agreed upon procedure for determining system losses *or* what losses should be considered when predicting radar performance. It is natural for a person selling a radar to be optimistic about the total system loss and claim a lower loss than might a potential buyer of the radar or an independent evaluator of a radar's performance. The advertised performance predicted by a radar manufacturer cannot be adequately verified or compared to the advertised predictions for similar radars by other manufacturers without complete knowledge of the losses that each radar designer has included.

Q.4 a. What are the differences between MTI radar and pulse Doppler radar? What are the limitations to MTI performance? (8)

Answer:

The radars discussed in the previous chapter were required to detect targets in the presence of noise. In the real world, radars have to deal with more than receiver noise when detecting targets since they can also receive echoes from the natural environment, such as land, sea, and weather. These echoes are called *clutter* since they can "clutter" the radar display. Clutter echoes can be many orders of magnitude larger than aircraft echoes. When an aircraft echo and a clutter echo appear in the same radar resolution cell, the aircraft might not be detectable. Chapter 7 describes the characteristics of clutter and discusses methods for reducing these unwanted echoes in order to detect the desired *target* echoes. However, *the* most powerful method for detecting moving targets in the midst of large clutter is by taking advantage of the doppler effect, which is the change of frequency of the radar echo signal due to the relative velocity between the radar and the moving target. The use of the doppler frequency shift with a pulse radar for the detection of moving targets in clutter is the subject of this chapter.

Radar deserves much credit for enabling the Allies (chiefly the United Kingdom and the Waited States) in the first half of World War II to prevail in *the* crucial air battles and night naval engagements against the Axis powers. Almost all of the radars used in World War II, however, were by today's standards relatively simple pulse systems that did not employ the doppler effect. Fortunately, these pulse radars were able to accomplish their mission without the use of doppler. This would not be possible today. All high-performance military air-defense radars and all civil air-traffic control radars for the detection and tracking of aircraft depend on the doppler frequency shift to separate the large clutter echoes from the much smaller echoes from moving targets. Clutter echoes can be greater than the desired target echoes by as much as 60 or 70 dB, or mare, depending on the type of radar and the environment.

MTI Radar and Pulse Doppler Radar A pulse radar that employs the doppler shift for detecting moving targets is either an MTI (moving target indication) radar' or a pulse doppler radar.[2] The MTI radar has a pulse repetition frequency (prf) low enough to not have any range ambiguities as defined by Eq. (1.2), $R_{un} = c/f_p$. It does, however, have many ambiguities in the doppler domain. The pulse doppler radar, on the other hand, is just the opposite. As we shall see later in this chapter, it has a prf large enough to avoid doppler ambiguities, but it can have numerous range ambiguities. There is a h a medium-prf pulse doppler that accepts both range and doppler ambiguities, as discussed in Sec. 3.9.

In addition to detecting moving targets in the midst of large clutter echoes, the doppler frequency shift has other important applications in radar; such as allowing CW (continuous wave) radar to detect moving targets and to measure radial velocity, synthetic aperture radar and inverse synthetic aperture radar for producing images of targets, and meteorological radars concerned with measuring wind shear. These other uses of the doppler frequency shift are not discussed in this chapter..

Doppler Frequency Shift The doppler effect used in radar is the same phenomenon that was introduced in high school physics courses to describe the changing pitch of an audible siren from an emergency vehicle as it travels toward or away from the listener. In this chapter we are interested is the doppler effect that changes the frequency of the electromagnetic signal that propagates from the radar to a moving target and back to the radar. If the range to the target is R, then the total number of wavelengths λ in the two-way path from radar to target and return is $2R/\lambda$. Each wavelength corresponds to a phase change of 2π radians. The total phase change in the two-way propagation path is then

$$\phi = 2\pi \times \frac{2R}{\lambda} = 4\pi R/\lambda \qquad [3.1]$$

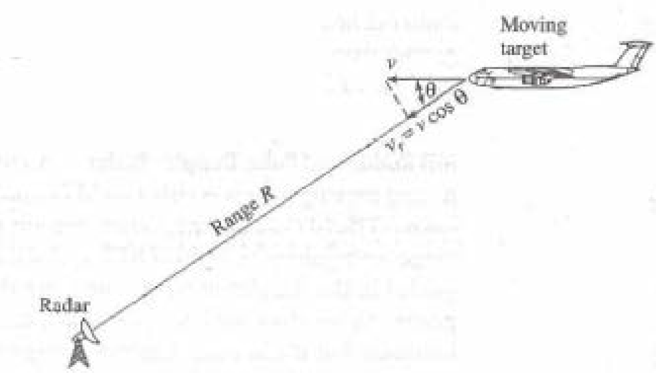
If the target is in motion relative to the radar, R is changing and so will the phase. Differentiating Eq. (3.1) with respect to time gives the rate of change of phase, which is the angular frequency

$$\omega_d = \frac{d\phi}{dt} = \frac{4\pi}{\lambda}\frac{dR}{dt} = \frac{4\pi v_r}{\lambda} = 2\pi f_d \qquad [3.2]$$

where $v_r = dR/dt$ is the radial velocity (meters/second), or rate of change of range with time. If, as in Fig. 3.1, the angle between the target's velocity vector and the radar line of sight to the target is θ_r then $v_r = v \cos \theta_r$ where v is the speed, or magnitude of the vector velocity. The rate of change of ϕ with time is the angular frequency $\omega_d = 2\pi f_d$, where f_d is the *doppler frequency shift*. Thus from Eq. (3.2),

$$f_d = \frac{2v_r}{\lambda} = \frac{2f_t v_r}{c} \qquad [3.3]$$

Figure 3.1 Geometry of radar and target in deriving the doppler frequency shift. Radar, target, and direction of target travel all lie in the same plane in this illustration.

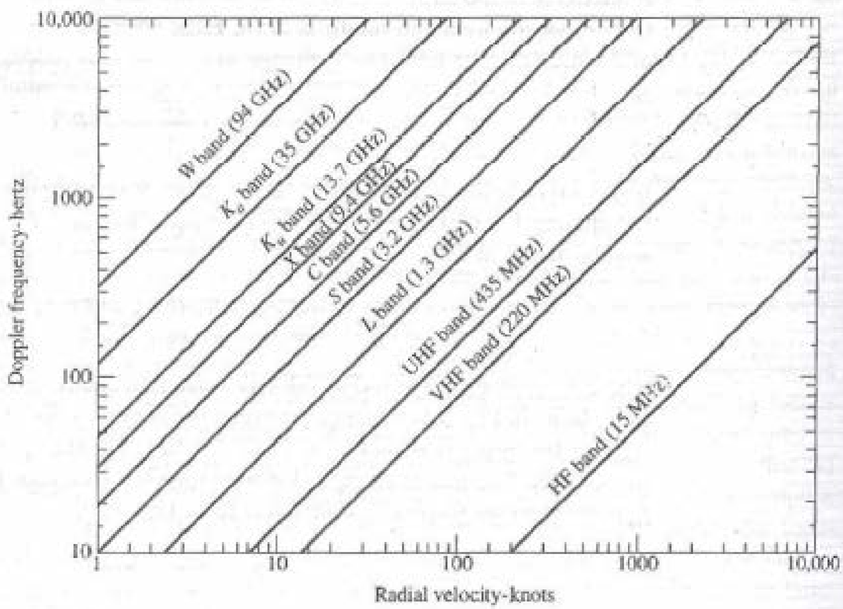


The radar frequency is $f_t = c/\lambda$, and the velocity of propagation $c = 3 \times 10^8$ m/s. If the doppler frequency is in hertz, the radial velocity in knots (abbreviated kt), and the radar wavelength in meters, we can write

$$f_d\ (\text{Hz}) = \frac{1.03 v_r\ (\text{kt})}{\lambda\ (\text{m})} \approx \frac{v_r\ (\text{kt})}{\lambda\ (\text{m})} \qquad [3.4]$$

The doppler frequency in hertz can also be approximately expressed as $3.43 v_r f_t$, where f_t is the radar frequency in GHz and v_r is in knots. A plot of the doppler frequency shift is shown in Fig. 3.2 as a function of the radial velocity and the various radar frequency bands.

Figure 3.3 Doppler frequency shift from a moving target as a function of the target's radial *velocity* and the radar frequency band.



Simple CW Doppler Radar Before discussing the use of doppler in pulse radar, it is instructive to begin by considering the doppler frequency shift experienced with a CW radar. The block diagram of a very simple CW radar that utilizes the doppler frequency shift to detect moving targets is shown in Fig. 3.3a. Unlike a pulse radar, a CW radar transmits while it receives. Without the doppler shift produced by the movement of the target, the weak CW echo signal would not be detected in the presence of the much stronger signal from the transmitter. Filtering in the frequency domain is used to separate the weak doppler-shifted echo signal from the strong transmitter signal in a CW radar.

The transmitter generates a continuous (unmodulated) sinusoidal oscillation at frequency f_t, which is then radiated by the antenna. On reflection by a moving target, the transmitted signal is shifted by the doppler effect by an amount $\pm f_d$, as was given by Eq. (3.3). The plus sign applies when the distance between radar and target is decreasing (a closing target); thus, the echo signal from a closing target has a larger frequency than that which was transmitted. The minus sign applies when the distance is increasing (a receding target). To utilize the doppler frequency shift a radar must be able to recognize that the received echo signal has a frequency different from that which was transmitted. This is the function of that portion of the transmitter signal that finds its way (or leaks) into the receiver, as indicated in Fig. 3.3a. The transmitter leakage signal acts as a reference to determine that a frequency change has taken place. The detector, or mixer, multiplies the echo signal at a frequency $f_t \pm f_d$ with the transmitter leakage signal f_t. The doppler filter allows the difference frequency from the detector to pass and rejects the higher frequencies. The filter characteristic is shown in Fig. 3.3a just below the doppler-filter block. It has a lower frequency cutoff to remove from the receiver output the transmitter leakage signal and clutter echoes. The upper frequency cutoff is determined by the maximum

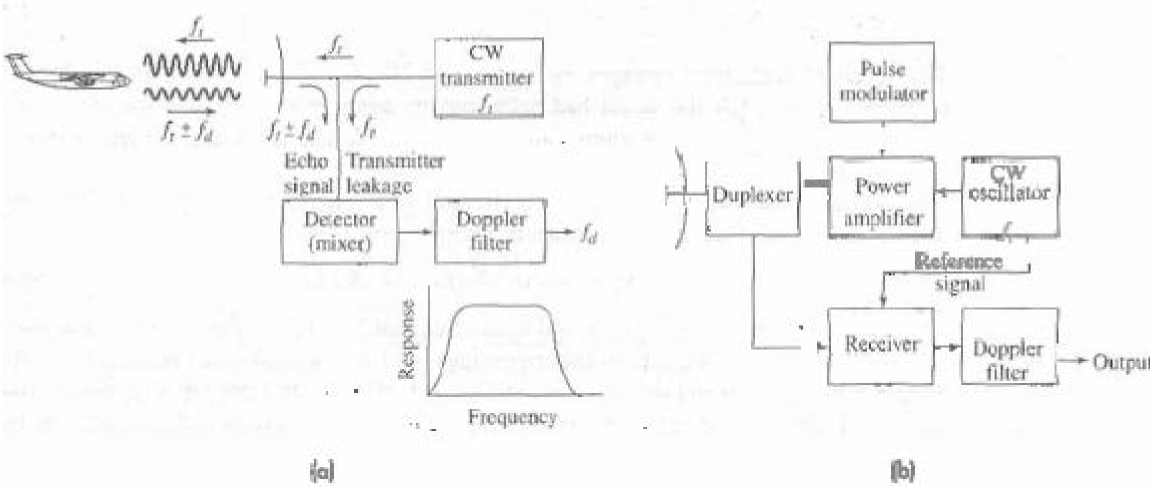


Figure 3.3 (a) Simple CW radar block diagram that extracts the doppler frequency shift from a moving target and rejects stationary clutter echoes. The frequency response of the doppler filter is shown at the lower right. (b) Block diagram of a simple pulse radar that extracts the doppler frequency shift of the echo signal from a moving target.

radial velocity expected of moving targets. The doppler filter passes signals with a doppler frequency f_d located within its pass band, but the sign af the doppler is lost along with the direction of the target motion. CW radars can be much more complicated than this simple example, but it is adequate as an introduction to a pulse radar that utilizes the doppler to detect moving targets in clutter.

Pulse Radar That Extracts the Doppler Frequency-Shifted Echo Signal One cannot simply convert the CW radar of Fig. 3.3a to a pulse radar by turning the CW oscillator on and off to generate pluses. Generating pulses in this manner also removes the reference signal at the receiver, which is needed to recognize that a doppler frequency shift has occurred. One way to introduce the reference signal is illustrated in Fig. 3.3b. The output of a stable CW oscillator is amplified by a high-power amplifier. The amplifier is turned on and off (modulated) to generate a series of high-power pulses. The received echa signal is mixed with the output of the CW oscillator which acts as a *coherent* reference to allow recognition of any change in the received echo-signal frequency. By *coherent* is meant that the phase of the transmitted pulse is preserved in the reference signal. The change in frequency is detected (recognized) by the doppler Alter.

The doppler frequency shift is derived next in a slightly different manner than was done earlier in this section. If the transmitted signal of frequency f_t is represented as $A_t \sin(2\pi f_t t)$, the received signal is $A_r \sin[2\pi f_t(t - T_R)]$, where A_t = amplitude of transmitted signal and A_r = amplitude of the received echo signal. The round-trip time T_R is equal to $2R/c$, where R = range and c = velocity of propagation. If rhe target is moving toward the radar, the range is changing and is represented as $R = R_0 - v_r t$, where v_r = radial velocity (nssumed constant). The geometry is the same as was shown in Fig. 3.1. With the above substitutions, the received signal is

$$V_{rec} = A_r \sin\left[2\pi f_t\left(1 + \frac{2v_r}{c}\right)t - \frac{4\pi f_t R_0}{c}\right] \qquad [3.5]$$

The received frequency changes by the factor $2f_t v_r/c = 2v_r/\lambda$, which is the doppler frequency shift f_d.* If the target had been moving away from the radar, the sign of the doppler frequency would be minus, and the received frequency would be less than that transmitted.

The received signal is heterodyned (mixed) with the reference signal $A_{ref} \sin 2\pi f_t t$ and the difference frequency is extracted, which is given as

$$V_d = A_d \cos(2\pi f_d t - 4\pi R_0/\lambda) \qquad [3.6]$$

where A_d = amplitude, $f_d = 2v_r/\lambda$ = doppler frequency, and the relation $f_t\lambda = c$ was used. (The cosine replaces the sine in the trigonometry of the heterodyning process.) For stationary targets $f_d = 0$ and the output signal is constant. Since the sine takes on values from +1 to −1, the sign of the clutter echo amplitude can be minus as well as plus. On the other hand, the echo signal from a moving target results in a time-varying output (due to the doppler shift) which is the basis for rejecting stationary clutter echoes (with zero doppler frequency) but allowing moving-target echoes to pass.

*The terms dopphr *frequency shift, doppler frequency,* and *doppler shift* are used interchangeably in his chapter.

If the radar pulse width is long enough and if the target's doppler frequency is large enough, it may be possible to detect the doppler frequency shift on the basis of the frequency change within a single pulse. If Fig. 3.4a represents the RF (or IF) echo pulse train, Fig. 3.4b is the pulse train when there is a recognizable doppler frequency shift. To detect a doppler shift on the basis of a single pulse of width τ generally requires that there be at least one cycle of the doppler frequency f_d within the pulse; or that $f_d\tau > 1$. This condition, however, is not usually met when detecting aircraft since the doppler frequency f_d is generally much smaller than $1/\tau$. Thus the doppler effect cannot be utilized with a *single* short pulse in this case. Figure 3.4c is more representative of the doppler frequency for aircraft-detection radars. The doppler is shown sampled at the pulse repetition frequency (prf). More than one pulse is needed to recognize a change in the echo frequency due to the doppler effect. (Figure 3.4c is exaggerated in that the pulse width is usually small compared to the pulse repetition period. For example, τ might be of the order of 1 μs, and the pulse repetition period might be of the order of 1 ms.)

Sweep*-Sweep Subtraction and the Delay-Line Canceler Figures 3.5a and b represent (in a very approximate manner) the bipolar video (both positive and negative amplitudes) from two successive sweeps* of an MTI (moving target indication) radar defined at the beginning of this chapter. The fixed clutter echoes in this figure remain the same from sweep to sweep. The output of the MTI radar is called *bipolar video*, since the signal has negative as well as positive values. *(Unipolar video* is rectified bipolar video with

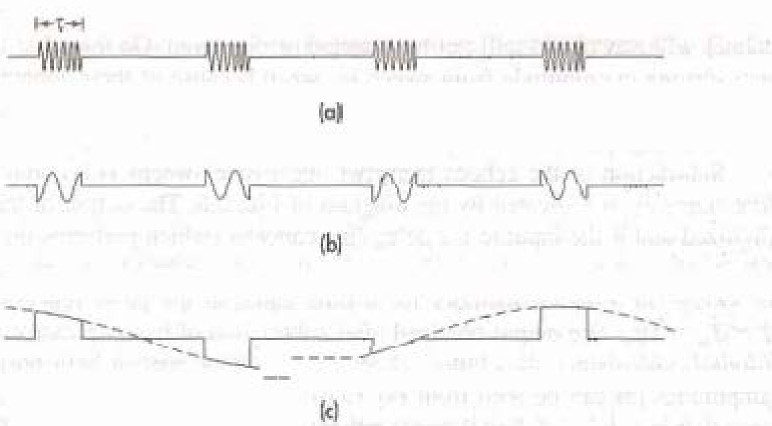


Figure 3.4 (a) Representation of the echo pulse train at either the RF or IF portion of the receiver; (b) video pulse train after the phase detector when the doppler frequency $f_d > 1/\tau$; (c) video pulse train for the doppler frequency $f_d < 1/\tau$, which is usually the case for aircraft-surveillance radar. The doppler frequency signal is shown dashed in (c), as if it were CW. Note that the pulses in (c) have an exaggerated width compared to the period of the doppler frequency.

*Sweep as used here is what occurs in the time between two transmitted pulses, or the pulse repetition interval. It is a more convenient term to use than is pulse repetition period, but the latter is more descriptive. The term sweep originally signified the action of moving the electron beam of a cathode ray tube display across the face of the tube during the time of a pulse repetition.

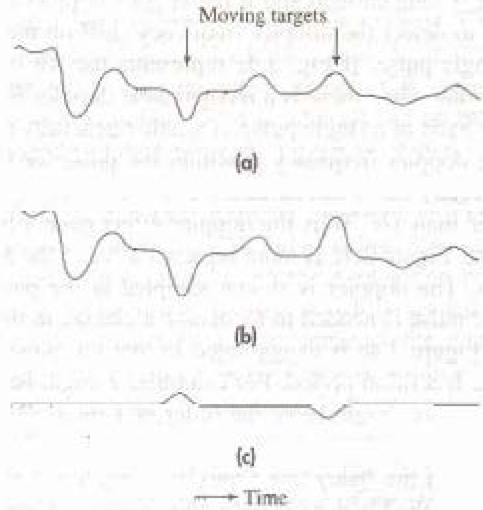
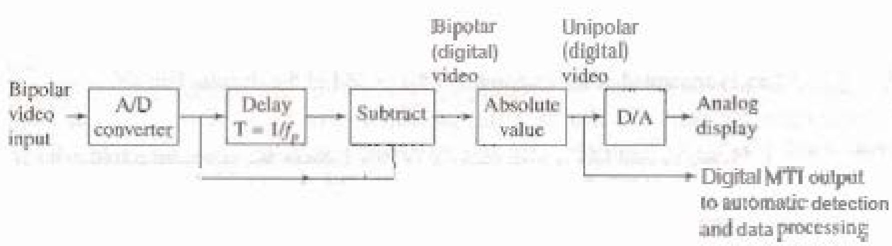


Figure 3.5 Two successive sweeps, (a) and (b), of an MTI radar A-scope display (amplitude as a function of time, or range). Arrows indicate the positions of moving targets. When (b) is subtracted from (a), the result is (c) and echoes from stationary targets are canceled, leaving only moving targets.

only positive values.) If one sweep is subtracted from the previous sweep, fixed clutter echoes will cancel and will not be detected or displayed. On the other hand, moving targets change in amplitude from sweep to sweep because of their doppler frequency shift. If one sweep is subtracted from the other, the result will be an uncancelled residue, as shown in Fig. 3.5c.

Subtraction of the echoes from two successive sweeps is accomplished in a *delay-line canceler*, as indicated by the diagram of Fig. 3.6. The output of the MTI receiver is digitized and is the input to the delay-line canceler (which performs the role of a doppler filter). The delay T is achieved by storing the radar output from one pulse transmission, or sweep, in a digital memory for a time equal to the pulse repetition period so that $T = T_p = 1/f_p$. The output obtained after subtraction of two successive sweeps is *bipolar (digital) video* since the clutter echoes in the output contain both positive and negative amplitudes [as can be seen from Eq. (3.6) when $f_d = 0$]. It is usually called *video*, even though it is a series of digital words rather than an analog video signal. The absolute value

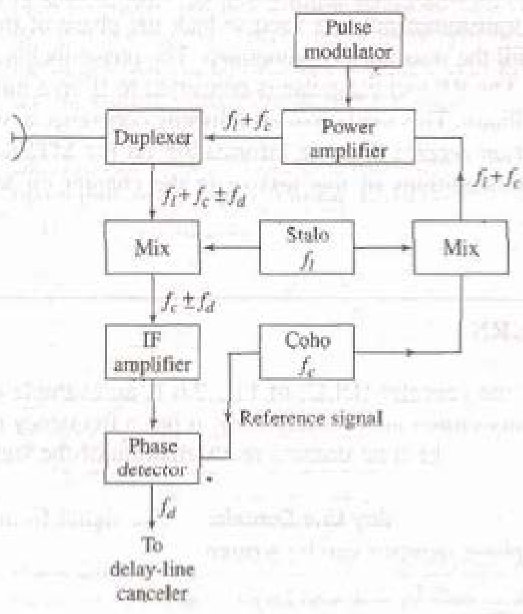
Figure 3.6 Block diagram of a single delay-line canceler.



of the bipolar video is taken, which is then *unipolar video.* Unipolar video is needed if an analog display is used that requires positive signals only. The unipolar digital video is then converted to an analog signal by the digital-to-analog (D/A) converter if the processed signal is to be displayed on a PPI (plan position indicator). Alternatively, the digital signals may be used for automatically making the detection decision and for further data processing, such as automatic tracking and/or target recognition. The name *delay-line can*-e s & was originally applied when analog delay lines (usually acoustic) were used in the early MTI radars. Even though analog delay lines have been replaced by digital memories, the name delay-line canceler is still used to describe the operation of Fig. 3.6.

MTI Radar Block Diagram The block diagram of Fig. 3.3 illustrated the reference signal necessary for an MTI radar, but it is oversimplified. A slightly more elaborate block diagram of an MTI radar employing a power amplifier as the transmitter is shown in Fig. 3.7. The local oscillator of an MTI radar's superheterodyne receiver must be more stable than the local oscillator for a radar that does not employ doppler. If the phase of the local oscillator were to change significantly between pulses, an uncancelled clutter residue can result at the output of the delay-line canceler which might be mistaken for a moving target even though only clutter were present. To recognize the need for high stability, the local oscillator of an MTI receiver is called the *stalo,* which stands for *stable local oscillator.* The IF stage is designed as a matched filter, as is usually the case in radar. Instead of an amplitude detector, there is a *phase detector* following the IF stage. This is a mixer-like device that combines the received signal (at IF) and the reference signal from the *coho* so as to produce the difference between the received signal and the reference signal frequencies.[3] This difference is the doppler frequency. The name *coho* stands for

Figure 3.7 Block diagram of an MTI radar that uses a power amplifier as the transmitter.



coherent oscillator to signify that it is the reference signal that has the phase of the transmitter signal. Coherency with the transmitted signal is obtained by using the sum of the coho and the stalo signals as the input signal to the power amplifier. Thus the transmitter frequency is the sum of the stalo frequency f_l and the coho frequency f_c. This is accomplished in the mixer shown on the upper right side of Fig. 3.7. The combination of the stalo and coho sometimes is called the *receiver-exciter* portion of the MTI radar. Using the receiver stalo and coho to also generate the transmitter signal insures better stability than if the functions were performed with two different sets of oscillators. The output of the phase detector is the input to the delay-line canceler, as in Fig. 3.6. The delay-line canceler acts as a high-pass filter to separate the doppler-shifted echo signals of moving targets from the unwanted echoes of stationary clutter. The doppler filter might be a single delay-line canceler as in Fig. 3.6; but it is more likely to be one of several other more elaborate filters with greater capability, as described later in this chapter.

The power amplifier is a good transmitter for MTI radar since it can have high stability and is capable of high power. The pulse modulator turns the amplifier on and off to generate the radar pulses. The klystron and the traveling wave tube have usually been the preferred type of vacuum-tube amplifier for MTI radar. The crossed-field amplifier has also been used, but it is generally noisier (less stable) than other devices; hence, it might not be capable of canceling large clutter echoes. Triode and tetrode vacuum tubes have also been used with success for radars that operate at UHF and lower frequencies, but they have been largely replaced at these lower radar frequencies by the solid-state transistor. The transistor has the advantage of good stability and it does not need a pulse modulator.

Before the development of the high-power klystron amplifier for radar application in the 1950s, the only suitable RF power generator at microwave frequencies was the magnetron oscillator. In an oscillator, the phase at the start of each pulse is random so that the receiver-exciter concept of Fig. 3.7 cannot be used. To obtain a coherent reference in this case, a sample of each transmitter pulse is used to lock the phase of the coho to that of the transmitted pulse until the next pulse is generated. The phase-locking procedure is repeated with each pulse. The RF locking-pulse is converted to IF in a mixer that also uses the stalo as the local oscillator. This method of establishing coherence at the receiver sometimes is called *coherent on receive*. Further information on the MTI using an oscillator may be found in previous editions of this text or in the chapter on MTI in the *Radar Handbook*.[1]

b. Draw the block diagram of delay line canceller and explain how it works. (8)

Answer:

The simple MTI delay-line canceler (DLC) of Fig. 3.6 is an example of a *time-domain filter* that rejects stationary clutter at zero frequency. It has a frequency response function $H(f)$ that can be derived from the time-domain representation of the signals.

Frequency Response of the Single Delay-Line Canceler The signal from a target at range R_0 at the output of the phase detector can be written

$$V_1 = k \sin(2\pi f_d t - \phi_0) \qquad [3.7]$$

where f_d = doppler frequency shift, ϕ_0 = a constant phase equal to $4\pi R_0/\lambda$, R_0 = range at time equal to zero, λ = wavelength, and k = amplitude of the signal. [For convenience, the cosine of Eq. (3.6) has been replaced by the sine.] The signal from the previous radar transmission is similar, except it is delayed by a time T_p = pulse repetition interval, and is

$$V_2 = k \sin [2\pi f_d(t - T_p) - \phi_0] \qquad [3.8]$$

The amplitude k is assumed to be the same for both pulses. The delay-line canceler subtracts these two signals. Using the trigonometric identity $\sin A - \sin B = 2 \sin[(A - B)/2] \cos [(A + B)/2]$, we get

$$V = V_1 - V_2 = 2k \sin (\pi f_d T_p) \cos \left[2\pi f_d\left(t - \frac{T_p}{2}\right) - \phi_0\right] \qquad [3.9]$$

The output from the delay-line canceler is seen to consist of a cosine wave with the same frequency f_d as the input, but with an amplitude $2k \sin (\pi f_d T_p)$. Thus the amplitude of the canceled video output depends on the doppler frequency shift and the pulse repetition period. The frequency response function of the single delay-line canceler (output amplitude divided by the input amplitude k) is then

$$H(f) = 2 \sin (\pi f_d T_p) \qquad [3.10]$$

Its magnitude $|H(f)|$ is sketched in Fig. 3.8.

The single delay-line canceler is a filter that does the job asked of it: it eliminates fixed clutter that is of zero doppler frequency. Unfortunately, it has two other properties that can seriously limit the utility of this simple doppler filter: (1) the frequency response function also has zero response when moving targets have doppler frequencies at the prf and its harmonics, and (2) the clutter spectrum at zero frequency is not a delta function of zero width, but has a finite width so that clutter will appear in the pass band of the delay-line canceler. The result is there will be target speeds, called *blind speeds*, where the target will not be detected and there will be an uncanceled clutter residue that can interfere with the detection of moving targets. These limitations will be discussed next.

Blind Speeds The response of the single delay-line canceler will be zero whenever the magnitude of $\sin (\pi f_d T_p)$ in Eq. (3.10) is zero, which occurs when $\pi f_d T_p = 0, \pm\pi, \pm 2\pi, \pm 3\pi, \ldots$ Therefore,

$$f_d = \frac{2v_r}{\lambda} = \frac{n}{T_p} = nf_p \qquad n = 0,1,2,\ldots \qquad [3.11]$$

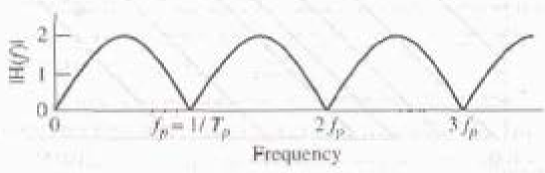


Figure 3.8 Magnitude of the frequency response $|H(f)|$ of a single delay-line canceler as given by Eq. (3.10), where f_p = pulse repetition frequency and $T_p = 1/f_p$.

This states that in addition to the zero response at zero frequency, there will also be zero response of the delay-line canceler whenever the doppler frequency $f_d = 2v_r/\lambda$ is a multiple of the pulse repetition frequency f_p. (The doppler shift can be negative or positive depending on whether the target is receding or approaching. When considering the blind speed and its effects, the sign of the doppler can be ignored—which is what is done here.) The radial velocities that produce blind speeds are found by equating Eqs. (3.11) and (3.3), and solving for the radial velocity, which gives

$$v_n = \frac{n\lambda}{2T_p} = \frac{n\lambda f_p}{2} \qquad n = 1,2,3\ldots \tag{3.12}$$

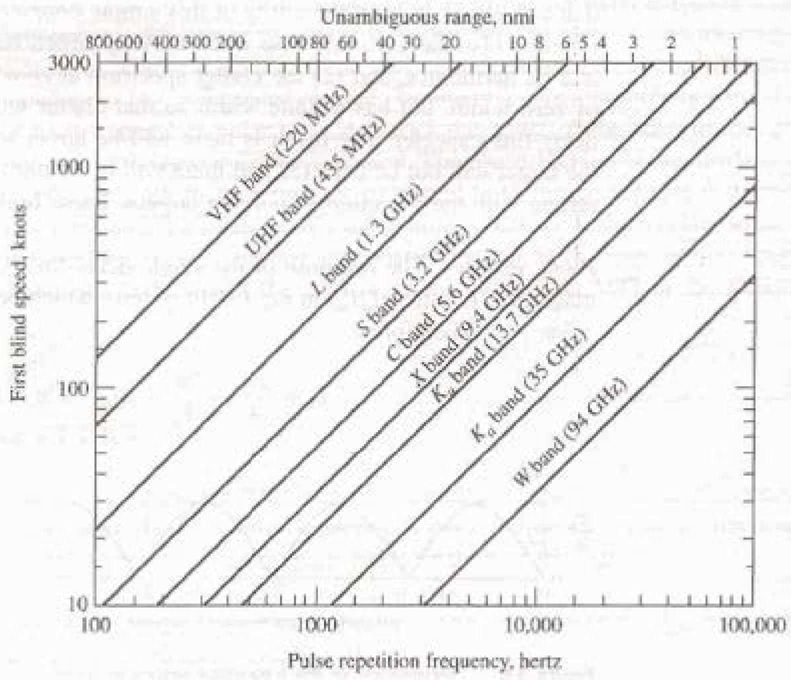
where v_r has been replaced by v_n, the nth blind speed. Usually only the first blind speed v_1 is considered, since the others are integer multiples of v_1. If λ is measured in meters, f_p in hertz, and the radial velocity in knots, the first blind speed can be written

$$v_1\,(\text{kt}) = 0.97\,\lambda\,(\text{m})\,f_p\,(\text{Hz}) \approx \lambda\,(\text{m})\,f_p\,(\text{Hz}) \tag{3.13}$$

A plot of the first blind speed as a function of the pulse repetition frequency and the various radar frequency bands is shown in Fig. 3.9.

Blind speeds can be a serious limitation in MTI radar since they cause some desired moving targets to be canceled along with the undesired clutter at zero frequency. Based on Eq. (3.13) there are four methods for reducing the detrimental effects of blind speeds:

Figure 3.9 Plot of the first blind speed, Eq. (3.13), as a function of the pulse repetition frequency for the various radar frequency bands.



1. Operate the radar at long wavelengths (low frequencies).
2. Operate with a high pulse repetition frequency.
3. Operate with more than one pulse repetition frequency.
4. Operate with more than one RF frequency (wavelength).

Combinations of two or more of the above are also possible to further alleviate the effect of blind speeds. Each of these four methods has particular advantages as well as limitations, so there is not always a clear choice as to which to use in any particular application.

Consider the case where a low RF frequency is chosen to avoid blind speeds. If, for example, the first blind speed is to be no lower than 640 kt (approximately Mach 1) and the prf is selected as 330 Hz (which gives an unambiguous range of 245 nmi), then the radar wavelength from Eq. (3.13) is 2 m. This corresponds to a frequency of 150 MHz (the VHF region of the spectrum). Although there were many radars at VHF built in the 1930s and early 1940s and there are still advantages to operating a radar at these frequencies, VHF is not usually considered a desirable frequency choice for a long-range air-surveillance radar for a number of reasons: (1) resolution in range and angle are poor due to narrow bandwidths and large beamwidths, (2) this portion of the electromagnetic spectrum is crowded with other than radar services (such as broadcast FM and TV), and (3) low-altitude coverage generally is poor. Thus attempting to use low frequencies to avoid the blind speed problem is not usually a desirable option for many radar applications.

On the other hand, if we choose to operate at a high RF frequency and increase the prf to avoid blind speeds, we would then have to tolerate the many range ambiguities that result. For example, if the first blind speed was again chosen to be 640 kt and the wavelength were 0.1 m (an *S*-band frequency of 3000 MHz), the prf would have to be 6600 Hz. This results in a maximum unambiguous range of 12.3 nmi, which is small for many radar applications. (Such an approach, however, is used successfully in pulse doppler radars, as discussed later in this chapter.)

When two or more prfs are used in a radar, the blind speeds at one prf generally are different from the blind speeds at the other prfs. Thus targets that are highly attenuated with one prf might be readily seen with another prf. This technique is widely used with air-surveillance radars, especially those for civil air-traffic control. A disadvantage of a multiple-prf waveform is that multiple-time-around clutter echoes (from regions beyond the maximum unambiguous range) are not canceled.

A radar that can operate at two or more RF frequencies can also unmask blind speeds, but the required frequency change is often larger than might be possible within the usual frequency bands allocated for radar use. A limitation of multiple frequencies is the need for greater system bandwidth.

In some circumstances, it might be desirable to tolerate the blind speeds rather than accept the limitations of the above methods. As in many aspects of engineering, there is no one single solution best for all cases. The engineer has to decide which of the above limitations can be accepted in any particular application.

Blind speeds occur because of the sampled nature of the pulse radar waveform. Thus it is sampling that is the cause of ambiguities, or aliasing, in the measurement of the doppler frequency; just as sampling in a pulse radar (at the prf) can give rise to ambiguities in the range measurement.

Clutter Attenuation The other limitation of the single delay-line canceler is insufficient attenuation of clutter that results from the finite width of the clutter spectrum. The single delay-line canceler whose frequency response was shown in Fig. 3.8 does what it is supposed to do, which is to cancel stationary clutter with zero doppler shift. In the "real world," however, the clutter spectrum has a finite width due to such things as the internal motions of the clutter, instabilities of the stalo and coho oscillators, other imperfections of the radar and its signal processor, and the finite signal duration. (The factors that widen the clutter spectrum will be discussed later in Sec. 3.7.) For present purposes we will assume the clutter power spectral density is represented by a gaussian function, and is written as

$$W(f) = W_0 \exp\left(-\frac{f^2}{2\sigma_c^2}\right) = W_0 \exp\left(-\frac{f^2\lambda^2}{8\sigma_v^2}\right) \qquad f \geq 0 \qquad [3.14]$$

where W_0 = peak value of the clutter power spectral density, at $f = 0$; σ_c = standard deviation of the clutter spectrum in hertz, and σ_v = standard deviation of the clutter spectrum in meters/second. The relation between the two forms of the clutter-spectrum standard deviation is based on applying the doppler frequency expression of Eq. (3.3) such that $\sigma_c = 2\sigma_v/\lambda$. The advantage of using the standard deviation σ_v is that it is often independent of the frequency; whereas σ_c is in hertz and depends on the radar frequency. Nevertheless, we will generally use σ_c in this chapter for the clutter spectrum.

The consequences of a finite-width clutter spectrum can be seen from Fig. 3.10. The frequency response of the single delay-line canceler shown by the solid curve encompasses a portion of the clutter spectrum; therefore, clutter will appear in the output. The greater the standard deviation σ_c, the greater the amount of clutter that will be passed by the filter to interfere with moving target detection. The clutter attenuation (CA) produced by a single delay-line canceler is

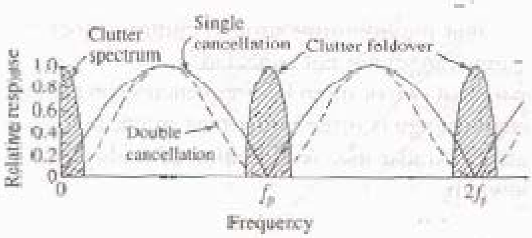


Figure 3.10 Relative frequency response of a single delay-line canceler (solid curve) and the double delay-line canceler (dashed curve), along with the frequency spectrum of the clutter (shaded area). Note the clutter spectrum is folded over at the prf and its harmonics because of the sampled nature of a pulse radar waveform

$$\mathrm{CA} = \frac{\int_0^\infty W(f)\,df}{\int_0^\infty W(f)|H(f)^2|\,df} \qquad [3.15]$$

where $H(f)$ is the frequency response of the delay-line canceler. Substituting for $H(f)$ [Eq. (3.10)], the clutter attenuation becomes

$$\mathrm{CA} = \frac{\int_0^\infty W_0 \exp\,[-f^2/2\sigma_c^2]\,df}{\int_0^\infty W_0 \exp[-f^2/2\sigma_c^2]\,4\sin^2(\pi f T_p)\,df} = \frac{0.5}{1-\exp\,(-2\pi^2 T_p^2 \sigma_c^2)} \qquad [3.16]$$

If the exponent in the denominator at the right-hand part of this equation is small, the exponential term $\exp\,[-x]$ can be replaced by $1 - x$, so that

$$\mathrm{CA} \approx \frac{f_p^2}{4\pi^2\sigma_c^2} = \frac{f_p^2\lambda^2}{16\pi^2\sigma_v^2} \qquad [3.17]$$

In this equation the pulse repetition period T_p has been replaced by $1/f_p$. The clutter attenuation provided by a single delay-line canceler is not sufficient for most MTI radar applications.

If a second delay-line canceler is placed in cascade, the frequency response of the two filters is the square of the single delay-line canceler, or

$$H(f) = 4\sin^2(\pi f T_p) \qquad [3.18]$$

This is indicated in Fig. 3. 10 by the dashed curve (except that we have plotted the relative response rather than the absolute response). Less of the clutter spectrum is included within the frequency response of the double delay-line canceler; hence, it attenuates more of the clutter. The clutter attenuation for the double delay-line canceler is

$$\mathrm{CA} \approx \frac{f_p^4}{48\pi^4\sigma_c^4} = \frac{f_p^4\lambda^4}{768\pi^4\sigma_v^4} \qquad [3.19]$$

Additional delay-line cancelers can be cascaded to obtain a frequency response $H(f)$ which is the nth power of the single delay-line canceler given by Eq. (3.10), where n is the number of delay-line cancelers.

MTI Improvement Factor The clutter attenuation is a useful measure of the performance of an MTI radar in canceling clutter, but it has an inherent weakness if one is not careful. The clutter attenuation can be made infinite by turning off the radar receiver! This, of course, would not be done knowingly since it also eliminates the desired moving-target echo signals. To avoid the problem of increasing clutter attenuation at the expense of desired signals, the IEEE defined[4] a measure of performance known as the ***MTI improvement factor*** which includes the signal gain as well as the clutter attenuation. It is defined as "The signal-to-clutter ratio at the output of the clutter filter divided by the signal-to-clutter ratio at the input of the clutter filter, averaged uniformly over all target radial velocities of interest." It is expressed as

$$\text{improvement factor} = I_f = \left.\frac{(\text{single/clutter})_{\text{out}}}{(\text{single/clutter})_{\text{in}}}\right|_{f_d} = \left.\frac{C_{\text{in}}}{C_{\text{out}}} \times \frac{S_{\text{out}}}{S_{\text{in}}}\right|_{f_d} =$$

$$= \text{CA} \times \text{average gain} \quad [3.20]$$

The vertical line on the right in the above equation indicates that the average is taken with respect to doppler frequency f_d. The improvement factor can be expressed as the clutter attenuation $\text{CA} = (C_{in}/C_{out})$ times the average filter gain. The average gain is determined from the filter response $H(f)$ and is usually small compared to the clutter attenuation. The average gain for a single delay-line canceler is 2 and for a double delay-line canceler is 6. The improvement factors for single and double delay-line cancelers are

$$I_f(\text{single DLC}) \approx \frac{1}{2\pi^2(\sigma_c/f_p)^2} = \frac{\lambda^2}{8\pi^2(\sigma_v/f_p)^2} \quad [3.21]$$

$$I_f(\text{double DLC}) \approx \frac{1}{8\pi^4(\sigma_c/f_p)^4} = \frac{\lambda^4}{128\pi^4(\sigma_v/f_p)^4} \quad [3.22]$$

The general expression for the improvement factor for a canceler with n delay-line cancelers in cascade is[5]

$$I_f(n \text{ cascaded DLCs}) \approx \frac{2^n}{n!}\left(\frac{1}{2\pi(\sigma_c/f_p)}\right)^{2n} \quad [3.23]$$

As with the previous expressions, this applies when σ_c/f_p is small. A plot of the improvement factor as a function of σ_c/f_p is provided later in Fig. 3.13. The ratio σ_c/f_p is a measure of the amount of "doppler space" occupied by clutter. Equation (3.23) also applies for the so-called *N-pulse canceler* with $N = n + 1$, to be discussed next.

N-Pulse Delay-Line Canceler A double delay-line canceler is shown in Fig. 3.11a. A canceler with two delay lines that has the same frequency response as the double delay-line canceler, but which is arranged differently, is shown in Fig. 3.11b. This is known as a *three-pulse canceler* since three pulses are added, with appropriate weights as shown. To obtain a $\sin^2(@,)$ response, the weights of the three pulses are +1, −2, +1. When the input is $s(t)$, the output of the three-pulse canceler is then

$$s(t) - 2s(t + T_p) + s(t + 2T_p)$$

which is the same as the output from the double delay-line canceler

$$s(t) - s(t + T_p) - [s(t + T_p) - s(t + 2T_p)].$$

Thus the double delay-line canceler and the three-pulse canceler have the same frequency response function.

A four-pulse canceler with weights +1, −3, +3, −1 has a frequency response proportional to $\sin^3(\pi f T_p)$. A five-pulse canceler has weights +1, −4, +6, −4, +1. If n is

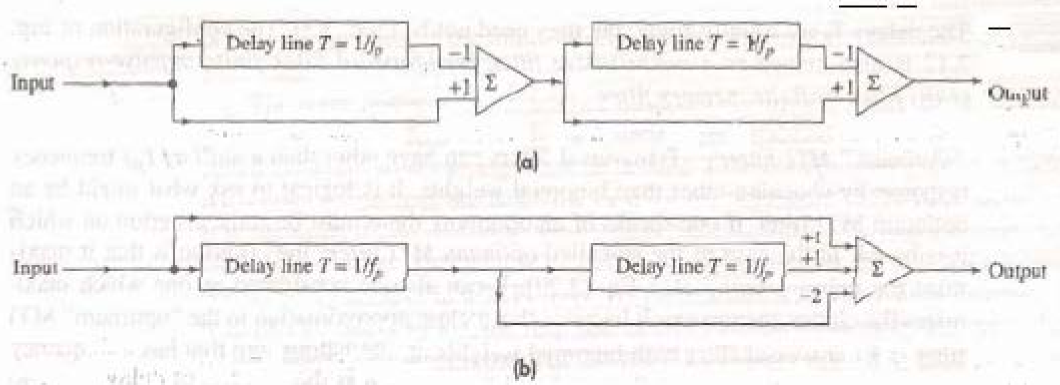


Figure 3.11 (a) Double delay-line canceler; (b) three-pulse canceler. The two configurations have the same frequency response. The three-pulse canceler of (b) is an example of a transversal filter.

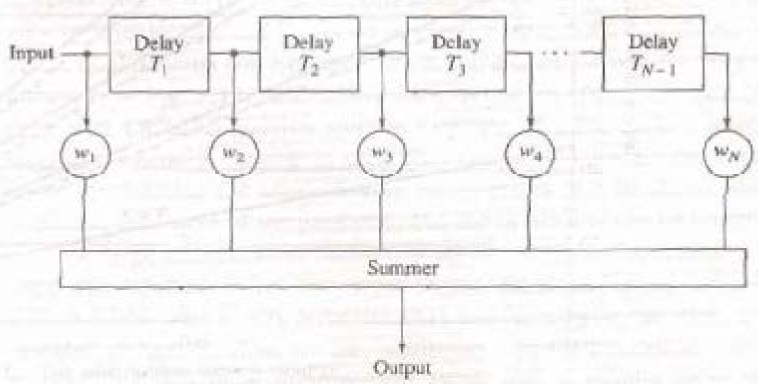
the number of delay lines, there are $n + 1 = N$ pulses available to produce a frequency response function proportional to $\sin^n(\pi/T_p)$ when the weights are the coefficients of the expansion of the binomial series $(1 - x)^n$ with alternating signs. The binomial weights with alternating sign are given by

$$w_i = (-1)^{i-1} \frac{n!}{(n - i + 1)!(i - 1)!} \qquad i = 1,2,\cdots, n + 1 \qquad [3.24]$$

The N-pulse canceler has the same frequency response as n single delay-line cancelers in cascade, where $n = N - 1$. The greater the value of N, the greater will be the clutter attenuation.

Transversal (Nonrecursive) Filter The three-pulse canceler of Fig. 3.11b is an example of a *transversal filter*. Its general form with n delay lines is shown in Fig. 3.11.The weights w_i are applied to the $N = n + 1$ pulses and then combined in the summer, or adder. The transversal filter is a time-domain filter with feed forward lines and taps with weights w_i.

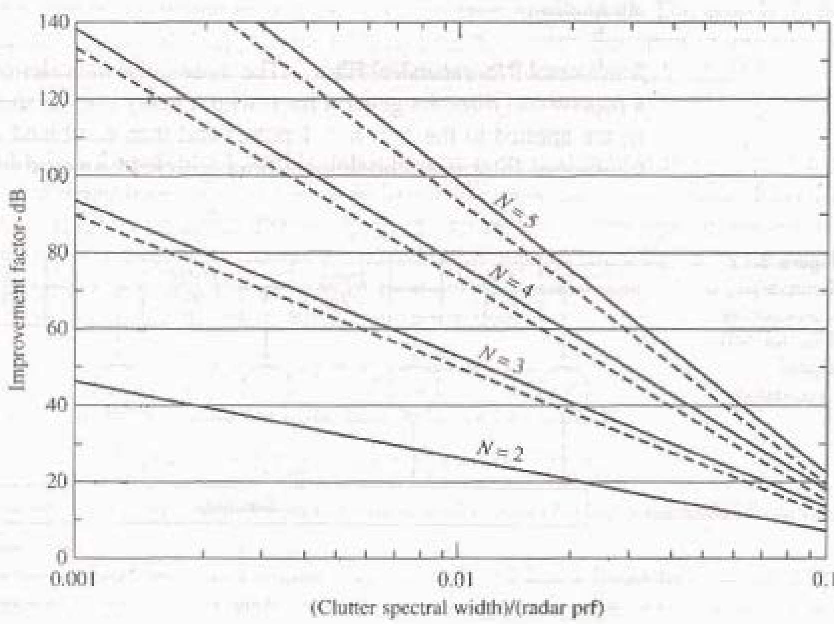
Figure 3.12 Transversal, a nonrecursive, filter for MTI signal processing.



The delays T_i are usually equal, but they need not be (Sec. 3.3). The configuration of Fig. 3.12 is also known as a *nonrecursive filter, feed-forward filter, finite impulse-response (FIR) filter,* or *finite memory filter.*

"Optimum" MTI Filter Transversal filters can have other than a $\sin^n(\pi f T_p)$ frequency response by choosing other than binomial weights. It is logical to ask what might be an optimum MTI filter. If one speaks of an optimum, there must be some criterion on which it is based. In the case of the so-called *optimum MTI filter*[6] the criterion is that it maximize the *improvement factor,* Eq. (3.20). It can also be considered as one which maximizes the clutter attenuation. It happens that a close approximation to the "optimum" MTI filter is a transversal filter with binomial weights of alternating sign that has a frequency response function proportional to $\sin^n(\pi f T_p)$, where n is the number of delay lines, as was discussed above. The optimum weights $(1, -1)$ of a transversal filter with a single delay line are the same as that of the single delay-line canceler, if the clutter spectrum can be modeled as a gaussian probability density function.[7] The difference between the three-pulse canceler and the "optimum" three-pulse MTI filter is less than 2 dB.[8] The difference is also small for higher values of n. Figure 3.13 is a plot of the improvement factor as a function of σ_c/f_p, where σ_c = the standard deviation of the clutter spectrum assuming it is of gaussian shape, and f_p = the pulse repetition frequency.[9] The solid curves apply for optimum weights and the dashed curves apply for binomial weights. There are two things to be noted from this plot. First, the differences between the two sets of curves are small so that we will take the optimum MTI filter to be adequately approximated by the filter with binomial weights whose response is proportional to $\sin^n(\pi f T_p)$. Second, sufficient improvement factor for many applications might be obtained with no more than two

Figure 3.13 MTI improvement factor for an N-pulse delay-line canceler with binomial weights (dashed curves) and optimum weights (solid curves) as a function of the clutter spectral width σ_c. (After Andrews.[9])



or three delay lines, if the improvement factor is the major criterion for the design of the MTI filter.

The term *optimum* is sometimes mistaken for the *best* that can be achieved. Optimum, however, is defined as the best under some implied or specified conditions. A so-called optimum solution might not be desired if the conditions for which it is defined are not suitable. This happens to be the case for the "optimum" MTI filter mentioned above. It is optimum if one wishes to have a filter that maximizes the improvement factor or the clutter attenuation. This may seem to be a suitable criterion, but it is not necessarily what one wants to achieve in a MTI filter. As the number n of delay lines increases in a filter with $H(f) \sim \sin^n(\pi f T_p)$, the response of $H(f)$ narrows and more and more clutter is rejected. The narrower bandwidth of the filter also means that fewer moving targets will be detected. If, for example, the -10 dB width of $H(f)$ is taken as the threshold for detection, and if all targets are uniformly distributed across the doppler frequency band, the following reductions in performance occur:

- 20 percent of all targets will be rejected by a two-pulse canceler
- 38 percent of all targets will be rejected by a three-pulse canceler
- 48 percent of all targets will be rejected by a four-pulse canceler

Thus if the "optimum" clutter filter is used, the loss of desired target detections is another reason it should not employ a large value of n.

"*Rectangula*" ... If one examines the clutter spectrum, such as the shaded region in Fig. 3.10, it can be seen that a desirable filter should approximate a rectangular passband that attenuates the clutter but has uniform response over as much of the doppler space as practical. It would not have as much clutter attenuation as an "optimum" filter of the same number of delay lines; but as we have seen from Fig. 3.13, the clutter attenuation of the "optimum" generally is far greater than can be used in practice when the number of delay lines n is large. A transversal filter, as in Fig. 3.12 approximate a rectangular passband if it contains a sufficient number of delay lines and if the weights w_i are chosen appropriately.

Some examples of transversal, or nonrecursive, filters for MTI applications that have appeared in the literature have been summarized by Y. H. Mao.[10] Procedures for nonrecursive filter design can be found in classical text books on digital *filters*. An early example due to Houts and Burlage,[11] based & a Chebyshev filter response that employs 15 pulses, is in Fig. 3.14. Also shown for comparison is the response of a three-pulse can-&& with binomial weights and the response of a five-pulse canceler with "optimum" weights. Generally, the goal in such filter design is to achieve the *necessary* improvement factor by choosing the attenuation in the stopband of a bandpass filter, the extent of the stopband, the extent and the ripple that can be tolerated in the passband.

The large improvement factor that results with the "optimum" MTI filter when n is large can be traded for increased doppler-frequency passband. For example, when $\sigma_c/f_p = 0.001$, Eq. (3.22) indicates that the three-pulse-canceler, or double delay-line canceler, (which is close to the "optimum") has a theoretical improvement factor of 91 dB. This is a large improvement factor and is usually more than is required for

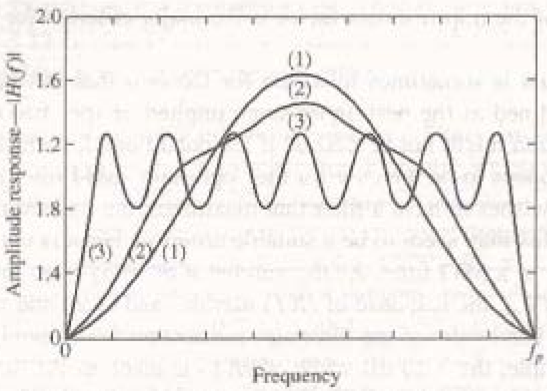


Figure 3.14 Amplitude response for three MTI delay-line cancelers. (1) Classical three-pulse canceler, (2) five-pulse delay-line canceler with optimum weights, and (3) 15-pulse Chebyshev design.
(After Houts and Burlage.[11])

routine MTI radar applications. Furthermore, it might be difficult to achieve such high values in practice considering the problems of equipment instability and other factors that can limit the improvement factor. The five-pulse "optimum" of Fig. 3.14 is indicated by Houts and Barlage[11] as having an improvement factor of 85 dB for this clutter spread and the 15-pulse Chebyshev design has an improvement factor of 52 dB. If there are many pulses available for MTI processing, an approximately rectangular filter response may be preferred over the "optimum" since increased doppler-frequency passband is more important than extremely large theoretical values of improvement factor which are not needed or cannot be achieved in practice. It has been said[11] that even with only five pulses available, a five-pulse Chebyshev design provides significantly wider doppler space than the five-pulse "optimum" design.

When only a few pulses are available for processing, there is probably little that can be done to control the shape of the nonrecursive filter characteristic, and there might not be much gained by using other than a filter with binomial weights that has a $\sin^n(\pi f T_p)$ response.

Recursive Filters The N-pulse nonrecursive canceler discussed above allows the designer N zeros for synthesizing the frequency response using the classical z-plane procedure for filter design. Each feedforward line and its weight w_i correspond to a zero in classical filter design on the z-plane. Filter design using only zeros does not have the flexibility of filter design based on poles as well as zeros. Poles can be obtained with delay-line cancelers by employing feedback. With both feedback and feedforward lines providing both poles and zeros, arbitrary filter frequency-response functions can be synthesized from cascaded delay lines, within the limits of realizability.[12] These are known as *recursive filters* or *infinite impulse response (IIR) filters.* Significantly fewer delay lines (and fewer pulses) are needed to achieve desirable frequency-response functions than with nonrecursive filters that only have zeros available for design in the z-plane.

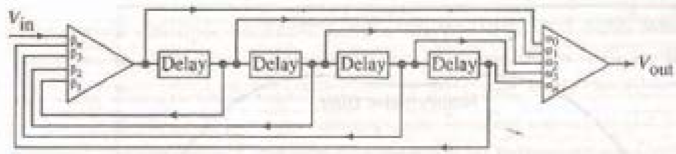


Figure 3.15 Canonical configuration of a recursive delay-line filter with both feedforward and feedback.
(After White and Ruvin, *IRE Natl. Conv. Rec.*, vol. 5, pt. 2, 1957.)

The canonical configuration of a time-domain filter with both feedback and feedforward is shown in Fig. 3.15. More usually, the canonical configuration is broken into sections with feedback and feedforward around individual delay lines. An example is the three-pole Chebyshev filter of Fig. 3.16a. The frequency response of this recursive filter is shown in Fig. 3.16b, with 0.5 dB ripple in the passband.[13] The width of the passband can be changed with different sets of weights. Figure 3.17, due to J. S. Shreve,[14] compares the response of a nonrecursive and recursive filter. It is seen that the recursive filter provides a frequency response that better resembles the rectangular shape than the nonrecursive, and does so with only two delay lines rather than the four of the nonrecursive filter.

Q.5 a. Summarize the characteristics of the matched filter for an input signal s(t). (8)

Answer:

Summary of the Matched Filter The characteristics of the matched filter for an input signal $s(t)$ are summarized below in short notation, omitting realizability factors and constants. The symbols have been defined previously in this section.

1. Frequency response function: $S^*(f)$
2. Maximum output signal-to-noise ratio: $2E/N_0$
3. Magnitude of the frequency response: $|H(f)| = |S(f)|$
4. Phase of the frequency response: $\phi_m(f) = -\phi_s(f)$
5. Impulse response: $s(-t)$
6. Output signal waveform for large signal-to-noise ratio: autocorrelation function of $s(t)$
7. Relation between bandwidth and pulse width for a rectangular-like pulse and conventional filter: $B\tau \approx 1$
8. Frequency response function for nonwhite noise: $S^*(f)/[N_i(f)]^2$

The matched filter makes radar-signal detection quite different from detection in conventional communication systems. The detectability of signals with a matched-filter receiver is a function only of the received signal energy E and the input noise spectral density N_0. Detection capability and the range of a radar do not depend on the shape of the signal or the receiver bandwidth. The shape of the transmitted signal and its bandwidth therefore can be selected to optimize the extraction of information without, in theory, affecting detection. Also different from communications is that the signal out of the matched filter is not the same shape as the input signal. It should be no surprise that the output signal's shape is different from the input since the criterion for the matched filter states only that detectability is to be maximized, not that the shape of the signal is to be preserved.

b. How the automatic detection of radar signals achieved and by what means it is different from conventional detection method? (8)

Answer:

An operator viewing a PPI display or an A-scope "integrates" in his/her eye-brain combination the echo pulses available from the target. Although an operator in many cases can be as effective as an automatic integrator, performance is limited by operator fatigue, boredom, overload, and the integrating characteristics of the phosphor of the CRT display. With automatic detection by electronic means, the operator is not depended on to make the detection decision. *Automatic detection* is the name applied to the part of the radar that performs the operations required for the detection decision without operator intervention. The detection decision made by an automatic detector might be presented to an operator for action or to a computer for further processing.

In many respects, automatic detection requires much better receiver design than when an operator makes the detection decision. Operators can recognize and ignore clutter echoes and interference that would limit the recognition abilities of some automatic devices. An operator might have better discrimination capabilities than automatic methods for sorting clutter and interference; but the automatic, computer-based decision devices can operate with far greater number of targets than an operator can handle.

Automatic detection of radar signals involves the following:

- *Quantization* of the radar coverage into range, and maybe angle, resolution cells.
- *Sampling* of the output of the range-resolution cells with at least one sample per cell, more than one sample when practical.
- *Analog-to-digital conversion* of the analog samples.
- *Signal processing* in the receiver to remove as much noise, clutter echoes, and interference as practicable before the detection decision is attempted.
- *Integration* of the available samples at each resolution cell.
- *Constant false-alarm rate* (CFAR) circuitry to maintain the false-alarm rate when the receiver cannot remove all the clutter and interference.
- *Clutter map* to provide the location of clutter so as to ignore known clutter echoes.
- *Threshold detection* to select target echoes for further processing by an automatic track or other dam processor.
- *Measurement of range and angle* after the detection decision is made.

The automatic detection and tracking (ADT) system, which includes the above. was discussed in Sec. 4.9. We next consider the automatic integration of signals and the application of CFAR in the automatic detection process.

The detector is that portion of the radar receiver that extracts the modulation from the carrier in order to decide whether or not a signal is present. It extends from the IF amplifier to the output of the video amplifier; thus, it is much more than a rectifying element. The conventional pulse radar as described in Chaps. 1 and 2 employs an *envelope detector* which extracts the amplitude modulation and rejects the carrier. By eliminating the carrier and passing only the envelope, the envelope detector destroys the phase information. There are other "detectors" in radar that are different from the above description. The MTI radar uses a *phase detector* to extract the phase of the radar echo relative to the phase of a coherent reference, as described in Chap. 3. In Chap. 4, the *phase-sensitive detector* employed in tracking radars for extracting angle information was mentioned.

Optimum Envelope Detector Law The envelope detector consists of the IF amplifier with bandpass filter characteristic, a rectifying element (such as a diode), and a video amplifier with a low-pass filter characteristic. The detector is called a *linear detector* if the relation between the input and output signal is linear for positive voltage signals, and zero for negative voltage. (The detector, of course, is a nonlinear device even though it bears the name *linear.*) When the output is the square of the input for positive voltage, the detector is called *square law.* The detector law is usually considered the combined law of the rectifying element and the video integrator that follows it, if an integrator is used. For example, if the rectifying element has a linear characteristic and the video integrator has a square-law characteristic, the combination would be considered a square-law detector. There can be, of course, many other detector laws beside the linear and the square law.

The *optimum detector* law can be found based on the use of the likelihood-ratio receiver. It can be expressed as[17–19]

$$y = \ln I_0(av) \qquad [5.22]$$

where y = output voltage of the detector

a = amplitude of the sinewave signal divided by the rms noise voltage

v = amplitude of the IF voltage envelope divided by the rms noise voltage

$I_0(x)$ = modified Bessel function of zero order

This equation specifies the form of the detector law that maximizes the likelihood ratio for a fixed probability of false alarm. A suitable approximation is[20]

$$y = \ln I_0(av) \approx \sqrt{(av)^2 + 4} - 2 \qquad [5.23]$$

For large signal-to-noise ratios ($a \gg 1$), this is approximately

$$y \simeq av$$

which is a linear law. For small signal-to-noise ratios, the approximation of Eq. (5.23) becomes

$$y \approx (av)^2/4$$

which has the characteristic of a square-law detector. Hence, for large signal-to-noise ratio, the optimum $\ln I_0$ detector may be approximated by a linear detector, and for small signal-to-noise ratios it is approximated by a square-law detector.

The linear detector usually is preferred in practice since it results in a higher dynamic range than the square law and is less likely to introduce distortion. On the other hand, the square-law detector usually is easier to analyze than the linear, so many analyses assume a square-law characteristic. Fortunately, the theoretical difference in detection performance between the square-law and linear detectors when performing noncoherent integration often is relatively insignificant.[21,22] Marcum[23] also showed that for a single pulse (no integration) the probability of detecting a given signal is independent of the detector law.

Logarithmic Detector If the output of the receiver is proportional to the logarithm of the input envelope, it is called a *logarithmic detector*, or *logarithmic receiver*. It finds application where large variations of input signals are expected. Its purpose is to prevent receiver saturation and/or to reduce the effects of unwanted clutter echoes in certain types of non-MTI receivers (as in the discussion of the log-FTC receiver in Sec. 7.8). A logarithmic characteristic is not used with MTI receivers since a nonlinear characteristic can limit the MTI improvement factor that can be achieved.

There is a loss in detectability with a logarithmic receiver. For 10 pulses integrated the loss in signal-to-noise ratio is about 0.5 dB, and for 100 pulses integrated, the loss is about 1.0 dB.[24] As the number of pulses increase, the loss approaches a maximum value of 1.1 dB.[25]

I,Q Detector The *I* and *Q*, or *in-phase* and *quadrature*, channels were mentioned in Sec. 3.5 in the discussion of the MTI radar. There it was noted that a single phase-detector fed by a coherent reference could produce a significant loss in signal depending on the relative timing (or "phase") of the pulse train and the doppler-shifted echo signal. In an MTI radar, the term *blind phase* (not a truly descriptive term) was used to describe this loss.

The loss due to blind phases was avoided if a second parallel detector channel, called the *quadrature*, or Q channel, were used with a reference signal 90° out of phase from the reference signal of the first channel, called the *in-phase*, or I channel. Most signal processing analyses now use I and Q channels as the receiver model especially when the doppler frequency is extracted.

The I,Q detector is more general than just for avoiding loss due in blind phases in an MTI radar. Figure 5.3 illustrates the I,Q detector. It is sometimes called a *synchronous detector*.[26] If the input is a narrowband signal having a carrier frequency f_0 (which could be the IF frequency) with a time-varying amplitude $a(t)$ and time-varying phase $\phi(t)$, then

$$\text{input signal: } s(t) = a(t) \sin [2\pi f_0 t + \phi(t)]$$

The output of the in-phase channel is $I(t) = a(t) \cos [\phi(t)]$ and the output of the quadrature channel is $Q(t) = a(t) \sin [\phi(t)]$. The input signal then can be represented as $s(t) = I(t) \sin 2\pi f_0 t + Q(t) \cos 2\pi f_0 t$. Thus the I and Q channels together provide the amplitude and phase modulations of the input signal.

If the outputs of the I and Q channels of Fig. 5.3 are squared and combined (summed), then the square root of the sum of the squares is the envelope $a(t)$ of the input signal. This describes an envelope detector. The phase $\phi(t)$ of the input signal is arctan (Q/I).

The I,Q representation is commonly used in digital signal processing.[27] The digitized signals are represented by complex numbers derived from the I and Q components. In each channel, the signal is digitized by an analog-to-digital (A/D) converter to produce a series of complex digital samples from the signal $I + jQ$. According to the sampling theorem, if the input signal has a bandwidth B there must be at least $2B$ samples per second (the Nyquist rate) to faithfully reproduce the signal. Because there are two channels in the I,Q detector, the A/D converter in each of the I and Q channels needs only to sample at the rate of B samples per second, thus reducing the complexity required of the A/D converters.

With a rate of B samples per second, there is a loss of about 0.6 dB compared to continuous sampling, since the sampling is not guaranteed to occur at the peak of the

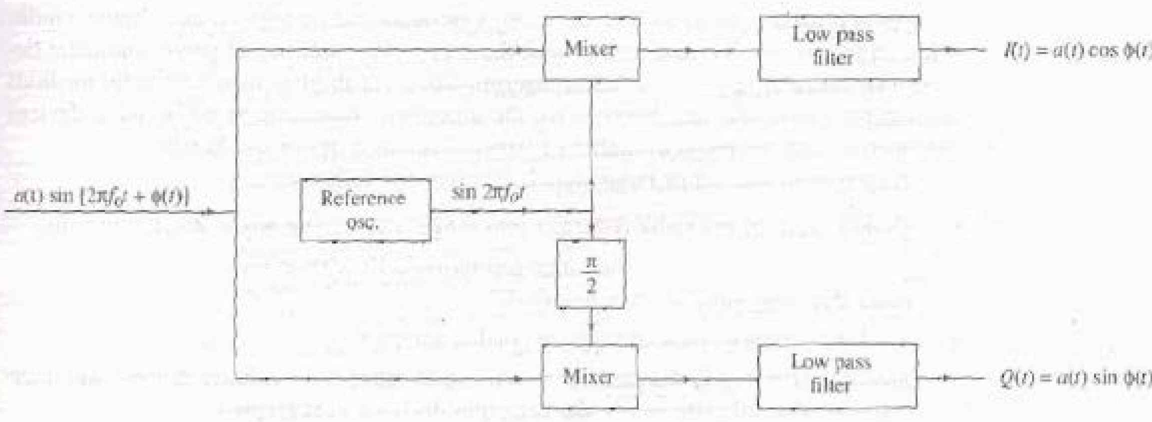


Figure 5.3 I,Q detector

output.[27] Much of this loss can be recovered by sampling at a rate of $2B$ samples per second. In some applications, further loss might occur due to the two channels not being precisely 90° out of phase, not being of equal gain, or if they are not perfectly linear.[28]

When I,Q channels are used for MTI processing, a doppler filter such as a delay-line canceler is included in each channel to separate moving targets from stationary clutter, as was discussed in Sec. 3.5.

Coherent Detector The so-called "coherent detector" sometimes has been described in the past literature as a single-channel detector similar to the in-phase channel of the I,Q detector, but with the reference signal at the same exact frequency and same exact phase as that of the input signal. Compared to the normal envelope detector of Chap. 2, the signal-to-noise ratio from a coherent detector might be from 1 to 3 dB greater. Unfortunately, the phase of the received radar signal is seldom known, so the single-channel coherent detector as described generally is not applicable to radar. The I,Q detector of Fig. 5.3 can also be considered as a coherent detector, but without the limitation of the coherent detector described above.

Q.6 a. What do you understand by term clutter? Enlist the different types of clutter (names only) and explain detection of target in sea clutter? (8)

Answer:

Clutter is the term used by radar engineers to denote *unwanted* echoes from the natural environment. It implies that these unwanted echoes "clutter" the radar and make difficult the detection of wanted targets. Clutter includes echoes from land, sea, weather (particularly rain), birds, and insects. At the lower radar frequencies, echoes from ionized meteor trails and aurora also can produce clutter. The electronic warfare technique known as *chaff*,* although not an example of the natural environment, is usually considered as clutter since it is unwanted and resembles clutter from rain. Clutter is generally distributed in spatial extent in that it is much larger in physical size than the radar resolution cell. There are also "point," or discrete, clutter echoes, such as TV and water towers, buildings, and other similar structures that produce large backscatter. Large clutter echoes can mask echoes from desired targets and limit radar capability. When clutter is much larger than receiver noise, the optimum radar waveform and signal processing can be quite different from that employed when only receiver noise is the dominant limitation on sensitivity.

Radar echoes from the environment are not always undesired. Reflections from storm clouds, for example, can be a nuisance to a radar that must detect aircraft; but storm clouds

*Chaff is an electronic countermeasure that consists of a large number of thin passive reflectors, often metallic foil strips. When released from an aircraft they are quickly dispersed by the wind to form a highly reflecting cloud. A relatively small bundle of chaff can form a cloud with a radar cross section comparable to that of a large aircraft.

containing rain are what the radar meteorologist wants to detect in order to measure rainfall rate over a large area. The backscatter echoes from land can interfere with many applications of radar, but they are the target of interest for ground-mapping radar, synthetic aperture radars, and radars that observe earth resources. Thus the same environmental echo might be the desired signal in one application and the undesired clutter echo in another. The observation of land, sea, weather and other natural phenomena by radar and other sensors for the purpose of determining something about the environment is known as *remote sensing of the environment,* or simply *remote sensing*. All radars, strictly speaking, are remote sensors; but the term is usually applied only to those radars whose major function is to observe the natural environment for the purpose of extracting information about the environment. A prime example of a radar used for remote sensing is the doppler weather radar.

Echoes from land or sea are examples of *surface clutter.* Echoes from rain and chaff are examples of *volume clutter.* The magnitude of the echo from distributed surface clutter is proportional to the area illuminated. In order to have a measure of the clutter echo that is independent of the illuminated area, the *clutter cross section per unit area,* denoted by the symbol σ^0, is commonly used to describe surface clutter. It is given as

$$\sigma^0 = \frac{\sigma_c}{A_c} \qquad [7.1]$$

where σ_c is the radar cross section of the clutter occupying an area A_c. The symbol σ^0 is spoken, and sometimes written, as *sigma zero.* It has also been called the *scattering coefficient, differential scattering cross section, normalized radar reflectivity, backscattering coefficient,* and *normalized radar cross section* (*NRCS*). The zero is a superscript since the subscript is reserved for the polarization employed. Sigma zero is a dimensionless quantity, and is often expressed in decibels with a reference value of one m^2/m^2.

Similarly, a cross section per unit volume is used to characterize volume clutter. It is defined as

$$\eta = \frac{\sigma_c}{V_c} \qquad [7.2]$$

where σ_c in this case is the radar cross section of the clutter that occupies a volume V_c. Clutter cross section per unit volume, η, is sometimes called the *reflectivity.*

In the next section, the radar range equation for targets in surface clutter is derived along with a brief review of the general character of scattering from surface clutter. This is followed by descriptions of radar echoes from land, sea, weather, and the atmosphere. The chapter concludes by describing methods that might be used to enhance the detection of targets in clutter.

The radar echo from the sea when viewed at low grazing angles is generally smaller than the echo from land. It usually does not extend as far in range as land clutter and is more uniform over the oceans of the worlds than typical land clutter. It has been difficult, however, to establish reliable quantitative relationships between sea echo measurements and the environmental factors that determine the sea conditions. Another difficulty in dealing with sea echo is that the surface of the sea continually changes with time. Nevertheless, there does exist a large body of information regarding the radar echo from the sea that can be used for radar design and provide a general understanding of its effect on radar performance.

The nature of the radar echo (clutter) from the sea depends upon the shape of the sea surface. Echoes are obtained from those parts of the sea whose scale sizes (roughness) are comparable in dimension to the radar wavelength. The shape, or roughness, of the sea depends on the wind. Sea clutter also depends on the pointing direction of the radar antenna beam relative to the direction of the wind. Sea clutter can be affected by contaminants that change the water surface-tension. The temperature of the water relative to that of the air is also thought to have an effect on sea clutter.

The *sea* generally consists of waves that result from the action of the wind blowing on the water surface. Such waves, called *wind waves*, cause a random-appearing ocean-height profile. *Swell waves* occur when wind waves move out of the region where they were originally excited by the wind or when the wind ceases to blow. Swell waves are less random and sometimes appear to be somewhat sinusoidal. They can travel great distances (sometimes thousands of miles) from the place where they originated. The echoes from an X-band radar viewing swell at low grazing angles will be small if there is no wind blowing, even if the swell waves are large. If a wind occurs, the surface will roughen and radar echoes will appear.

Sea state is a term used by mariners as a measure of wave height, as shown in Table 7.2. The sea state description shown in this table is that of the World Meteorological Organization. Sea state conditions can also be described by the Douglas scale, the Hydrographic Office scale, and the Beaufort scale. The Beaufort is actually a wind-speed scale.[21] Each gives slightly different values, so when a sea state is mentioned one should check which scale is being used.

Although sea state is commonly used to describe the roughness of the sea, it is not a complete indicator of the strength of sea clutter. Wind speed is often considered a better measure of sea clutter, but it is also limited since the effect of the wind on the sea depends on how long a time it has been blowing (called the *duration*) and over how great a distance (called the *fetch*). Once the wind starts to blow, the sea takes a finite time to grow and reach equilibrium conditions. When equilibrium is reached it is known as a *fully developed sea*. For example, a wind speed of 10 kt with a duration of 2.4 h and a fetch of at least 10 nmi produces a fully developed sea with a significant wave height (average height of the one-third highest waves) of 1.4 ft.[22] It corresponds to sea state 2. A 20 kt

Table 7.2 World Meteorological Organization sea state

	Wave Height		
Sea State	**Feet**	**Meters**	**Descriptive Term**
0	0	0	Calm, glassy
1	$0-\frac{1}{3}$	0–0.1	Calm, rippled
2	$\frac{1}{3}-1\frac{2}{3}$	0.1–0.5	Smooth, wavelets
3	2–4	0.6–1.2	Slight
4	4–8	1.2–2.4	Moderate
5	8–13	2.4–4.0	Rough
6	13–20	4.0–6.0	Very rough
7	20–30	6.0–9.0	High
8	30–45	9.0–14	Very high
9	over 45	over 14	Phenomenal

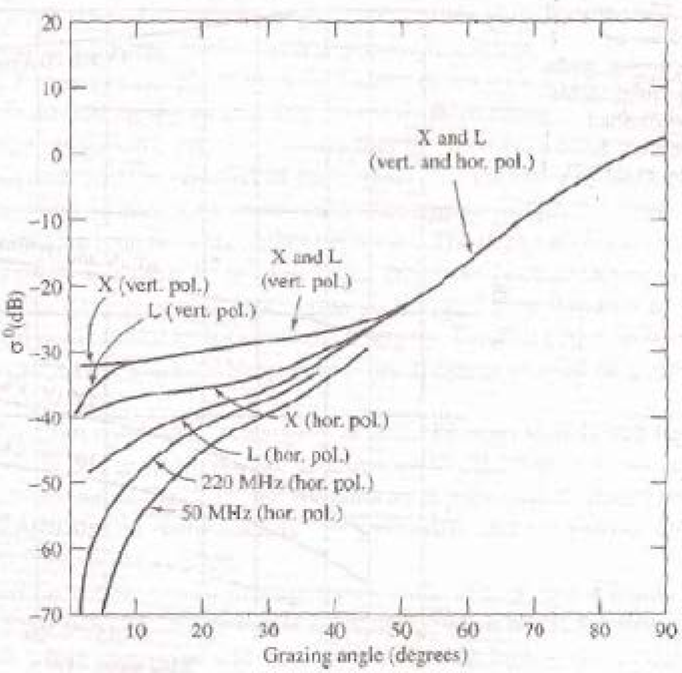
wind blowing for 10 h over a fetch of 75 nmi results in a significant wave height of about 8 ft and corresponds to sea state 4.

Average Value of σ^0 as a Function of Grazing Angle A composite of sea clutter data from many sources is shown in Fig. 7.13. This figure was derived from data for winds varying from approximately 10 to 20 kt, and can be considered representative of sea state 3. (Sea state 3 is roughly the medium value of sea state; i.e., about half the time over the oceans of the world, the sea state is 3 or less.) It is believed that Fig. 7.13 is representative of the average behavior of sea clutter, but there is more uncertainty in the data than is indicated by the thin lines with which the curves were drawn.

The curves of Fig. 7.13 provide the following conclusions for sea clutter with winds from 10 to 20 kt:

- At high grazing angles, above about 45°, sea clutter is independent of polarization and frequency.
- *Sea* clutter with vertical polarization is larger than with horizontal polarization. (At higher wind speeds the differences between the two polarizations might be less.)
- *Sea* clutter with vertical polarization is approximately independent of frequency. (This seems to hold at low grazing angles even down to frequencies in the HF band.)
- At low grazing angles, sea clutter with horizontal polarization decreases with decreasing frequency. This is apparently due to the interference effect at low angles between the direct radar signal and the multipath signal reflected from the sea surface.

Figure 7.13 Composite of averaged sea clutter σ^0 data from various sources for wind speeds ranging from 10 to 20 kt.



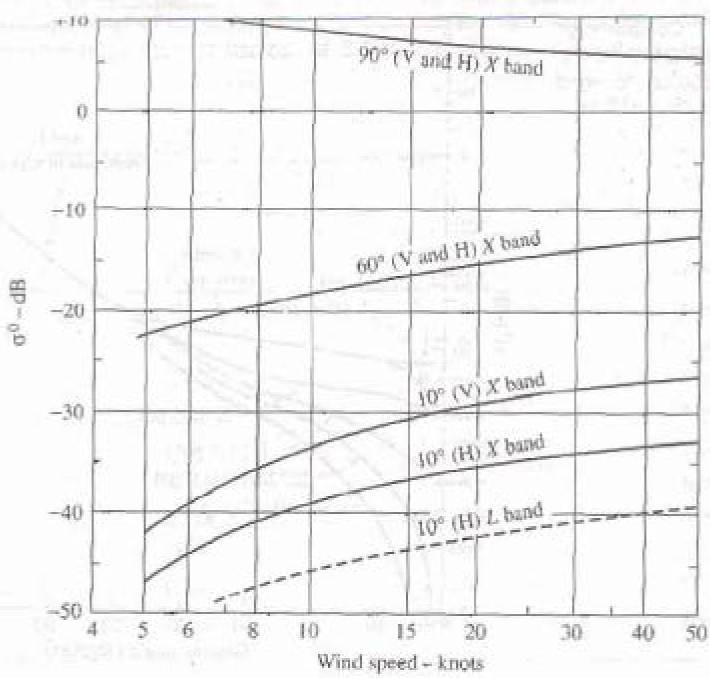
There is no simple law that describes the frequency dependence of sea clutter with horizontal polarization.

Effect of Wind The wind is the most important environmental factor that determines the magnitude of the sea clutter. At low grazing angles and at microwave frequencies, backscatter from the sea is quite low when the wind speed is less than about 5 kt. It increases rapidly with increasing wind from about 5 to 20 kt, and increases more slowly at higher wind speeds. At very high winds, the increase is small with increasing wind.

Figure 7.14 was derived from experimental data of John Daley et al. using the Naval Research Laboratory Four-Frequency Airborne Radar.[23-25] As mentioned, sea clutter at the higher microwave frequencies and low grazing angles increases with increasing wind speed, but begins to level off at winds of about 15 to 25 kt. When viewed at vertical incidence (grazing angle of 90°) and with zero or low wind speed, the sea surface is flat and a large echo is directed back to the radar. As the wind speed increases and the sea surface roughens, some of the incident radar energy is scattered in directions other than back to the radar, so that σ^0 will decrease. According to Daley et al., the value of σ^0 at vertical incidence in Fig. 7.14 can be represented by $25\, w^{-0.6}$, where w is the wind speed in knots.

At low grazing angles, less than about one degree, it is difficult to provide a quantitative measure of the effect of the wind on sea clutter. This is due to the many factors that influence σ^0 at low angles, such as shadowing of parts of the sea by waves, multipath interference, diffraction, surface traveling (electromagnetic) waves, and ducted propagation.

Figure 7.14 Effect of wind speed on sea clutter at several grazing angles. Radar looking upwind. Solid curves apply to X band, dashed curve applies to L band. (After J. C. Daley, et al.[23-25])



Another factor affecting the ability to obtain reproducible measurements of sea clutter is that a finite time and a finite fetch are required for the sea to become fully developed. Unfortunately, few measurements of sea clutter mention the fetch and duration of the wind.

Sea clutter is largest when the radar looks into the wind (upwind), smallest when looking with the wind (downwind), and intermediate when looking perpendicular to the wind (crosswind). There might be as much as 5 to 10 dB variation in σ^0 as the antenna rotates 360° in azimuth.[26] Backscatter is more sensitive to wind direction at the higher frequencies than at lower frequencies; horizontal polarization is more sensitive to wind direction than vertical polarization; the ratio of σ^0 measured upwind to that measured downwind decreases with increasing grazing angle and sea state; and at UHF the backscatter is practically insensitive to wind direction at grazing angles greater than 10°.

The orthogonal component of polarization from sea clutter (cross polarization response) at grazing angles from 5 to 60° appears to be about 5 to 15 dB less than the echo from the same polarization (co-pol) as transmitted.[25]

Sea Clutter with High-Resolution Radar (Sea Spikes) The use of a clutter density, σ^0 (clutter cross section per unit area), to describe sea clutter implies that the clutter echo is independent of the illuminated area. When sea clutter is viewed by a high-resolution radar, especially at the higher microwave frequencies (such as X band), sea clutter is not uniform and cannot be characterized by σ^0 alone. High-resolution sea clutter is spiky. The individual echoes seen with high-resolution radar are called *sea spikes*. They are sporadic and have durations of the order of seconds. They are nonstationary in time, spatially nonhomogeneous, and have a probability density function that is non-Rayleigh. Sea spikes are important since they are the major cause of sea clutter at the higher microwave frequencies at low grazing angles, with any radar resolution.

Fig. 7.15 is an example of the time history of sea spikes in sea state 3 for pulse widths varying from 400 to 40 ns and with vertical polarization.[27,28] With 40-ns pulse width (6-m range resolution), Fig. 7.15 shows that the time between sea spikes can be several tens of seconds and the duration of each spike is of the order of one or a few seconds. As the pulse width is increased, more sea spikes appear within the larger resolution cell of the radar and the time between spikes decreases. The peak radar cross section of sea spikes in this example is almost 10 m^2. (At times they have been observed to be of even higher cross section.) The relatively large cross section and time duration of sea spikes can result in their being mistaken for small radar targets. This is a major problem with sea spikes; they can cause false alarms when a conventional detector based on gaussian receiver noise is used.

Under calm conditions, sea state 1 or less, the echo signals still have the same spiky appearance as in Fig. 7.15, but are as much as 40 dB lower in cross section. A time history similar to that of Fig. 7.15, but for horizontal polarization, would show that sea spikes with this polarization occur slightly less frequently and are sharper (of briefer duration) than with vertical polarization.

Based on airborne radar measurements made at L, S, and X bands with pulse widths ranging from 0.5 to 5 μs, clutter is more spiky when the radar looks upwind or downwind rather than crosswind and with low rather than high grazing angles.[29]

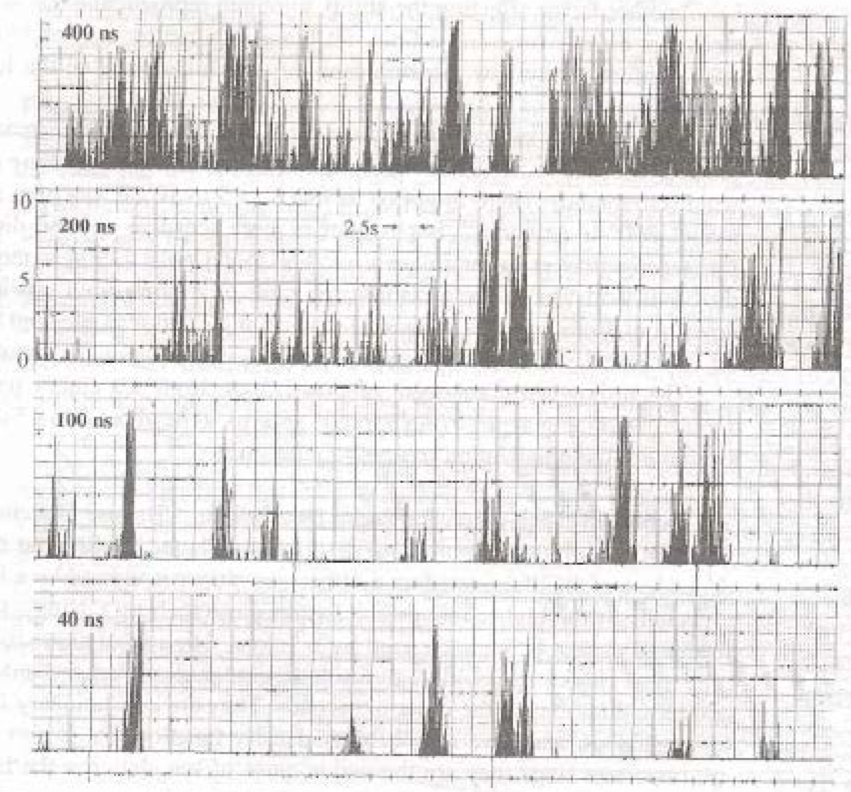


Figure 7.15 Amplitude as a function of time at a fixed range-resolution cell for low grazing angle *X*-band (9.2 GHz) sea clutter obtained off Boca Raton, Florida, with pulse widths ranging from 400 ns to 40 ns in a windblown sea with many white caps (sea state 3). Cross-range resolution is approximately 9 m, grazing angle of 1.4°, and vertical polarization.
I. (From J. P. Hansen and V. F. Cavaleri.[27])

Sea spikes are evident when the radar resolution is less than the water wavelength. The physical size of sea spikes usually is less than the resolution of the radar which observes them, and it is said they appear to move at approximately the surface wave velocity.[30] Sea spikes obviously are present with low as well as high resolution. At low resolution, the sum of many individual sea spikes within the resolution cell produces an almost continuous noiselike echo. At the lower frequencies where the radar wavelength is large compared to the sea-surface features that give rise to sea spikes, it would be expected that the backscatter is no longer characterized by sea spikes.

Another characteristic of sea spikes is a relatively rapid and high-percentage pulse-to-pulse amplitude modulation. According to Hansen and Cavaleri,[27] measured modulation frequencies at *X* band vary from 20 to 500 Hz. The modulating frequency seems to be affected by the type of physical surface (breaking water, sharp wave crests, ripples, etc.), the relative wind speed and direction, and polarization of the radar. The characteristic amplitude modulations can be used for recognizing and rejecting sea spike echoes from true target echoes.[31]

In addition to changes in clutter characteristics with narrow pulse widths, as illustrated in Fig. 7.15, similar effects have been noted with narrow antenna beamwidths.[32] In some cases, target detection in clutter can actually be enhanced when a smaller antenna with a lower resolution is used rather than one with a high resolution that is bothered by sea spikes.[32] On the other hand, with very high resolution, targets can be seen in the clear regions between the sea spikes, so that subclutter visibility in the traditional sense might not be required.

Origin of Sea Spikes Sea spikes are associated with breaking waves or waves about to break. Visible whitecaps are also associated with breaking waves, but whitecaps themselves do not appear to be the cause of sea spikes since whitecaps are mainly foam with entrapped air (which does not result in significant backscatter). It has been reported[33] that about 50 percent of the time the whitecap is seen visually either simultaneously or a fraction of a second later than the appearance of a sea spike on the radar display. About 35 to 40 percent of the time a spike is observed when the waves have a very peaked crest, but with no whitecap developed. A sea spike echo can appear without the presence of a whitecap, but no whitecap is seen in the absence of a radar observation of a spike. Thus it can be concluded that the whitecap is not the cause of the sea spike echo.

Wetzel[34] has offered an explanation for the origin of sea spikes based on the entraining plume model of a spilling breaker.[35] In this model "a turbulent plume emerges from the unstable wave crest and accelerates down the forward face of the breaking wave, entraining air as it goes." Wetzel further assumes "that the breaking 'event' involves a cascade of discrete plumes emitted along the wave crest at closely spaced times." Based on this model and some assumptions about the characteristics of the plume, Wetzel was able to account for the peak radar cross section, frequency dependence, the spikier nature of horizontal polarization, similarity of the appearance of sea spikes with both horizontal and vertical polarizations at very low grazing angles, appearance of the characteristic modulation, and other properties. He has pointed out that this model is based on a simplistic hypothesis that requires further elaboration. It does not account, however, for the scattering when the radar views the sea downwind and it does not adequately explain the internal amplitude modulations of the sea spikes.

Detection of Signals in High-Resolution Sea Clutter The nature of sea spikes as seen by high-resolution radar results in a probability density function (pdf) that is not Rayleigh. Therefore, conventional methods for detection of signals in gaussian noise do not apply. (A Rayleigh pdf for clutter power is equivalent to a gaussian pdf for receiver noise voltage.) Non-Rayleigh pdfs have "high tails"; that is, there is a higher probability of obtaining a large value of clutter than when it is Rayleigh. A receiver detector designed as in Chap. 2 on the basis of gaussian noise, or Rayleigh clutter, will result in a high false-alarm rate when confronted with sea spikes. The statistics of non-Rayleigh sea clutter are not easily quantified and can vary with resolution and sea state. Thus conventional receiver detector design based on the assumption of gaussian noise cannot be applied. To avoid excessive false alarms with non-Rayleigh sea clutter, the detection-decision threshold might have to be increased (perhaps 20 to 30 dB). The high thresholds necessary to avoid false alarms reduce the probability of detecting desired targets. When sea spikes are a concern, detection criteria other than those based on gaussian or Rayleigh statistics must be used if a severe penalty in detection capability is to be avoided.

One method for dealing with a sea spike is to recognize its characteristic amplitude modulations and delete the sea spike from the receiver.[31] Another method is to employ a receiver with a log-log output-input characteristic whose function is to provide greater suppression of the higher values of clutter than a conventional logarithmic receiver. In a log-log receiver, the logarithmic characteristic progressively declines faster than the usual logarithmic response by a factor of 2 to 1 over the range from noise level to +80 dB above noise.[36].

Conventional pulse-to-pulse integration does not improve the detection of targets in spiky clutter because of the long correlation time of sea spikes. However, a high antenna scan rate (several hundred rpm) allows independent observations of the clutter to be made, so that scan-to-scan integration can be performed.[37] Similarly, if the target is viewed over a long period of time before a detection decision is made, the target can be recognized by its being continuously present on the display while sea spikes come and go. One method for achieving this is *time compression*, as mentioned later in Sec. 7.8.

The effect of sea spikes is less important for radars that are to detect ships since ship cross sections are large compared to the cross sections of sea spikes. Sea spikes might interfere, however, with the detection of small targets such as buoys, swimmers, submarine periscopes, debris, and small boats. With ultrahigh resolution (ultrawideband radar) where the range resolution might be of the order of centimeters, sea spikes are relatively sparse in both time and space so that it should be possible to see physically small targets when they are located in between the spiky clutter.

Sea Clutter at Very Low Grazing Angles[38] Clutter at very low grazing angles differs from that at higher angles because of shadowing, ducted propagation (Sec. 8.5), and the changing angle at which the radar ray strikes the fluctuating sea surface.

The surface of the sea is seldom perfectly flat. It usually has a time-varying angle with respect to the radar. A grazing angle might be defined with respect to the horizontal, but it is difficult to determine the angle made with respect to the dynamic sea surface. Refraction by the atmosphere can also change the angle the radar ray makes with the surface.

Shadowing of wave troughs by wave crests can occur. Scatterers as seen by the radar are thought to be mainly those from the crests, especially for horizontal polarization. Attempts to compute the effect of shadowing by simple geometrical considerations (to determine how much of the sea surface is masked) have not proven successful. One reason for failure of geometrical shadowing is that scattering features (such as sloshes, plumes, and other effects of breaking waves) are not uniformly distributed and are more likely to be near the crests of the waves. Diffraction effects can occur which complicate shadowing calculations. Thus the effect of shadowing is more complicated than just simple masking.[38,39]

Another factor to consider when searching for a theoretical understanding of sea spikes (and microwave sea clutter in general) is the effect of a surface traveling radar wave that is launched when the incident wave has a component of electric field in the plane of incidence.[40] The effect was mentioned in Sec. 2.7 and shown in Fig. 2.10 for scattering from a long thin road. The surface traveling wave and its reflection from a discontinuity can be a reason why microwave sea clutter seen with vertical polarization usually is larger than sea clutter seen with horizontal polarization.

At very low grazing angles (less than about one degree), sea clutter should decrease rapidly with decreasing grazing angle, especially for horizontal polarization. Figure 7.3, which is a generic plot of clutter echo strength as a function of grazing angle, attempts to show this behavior. The decrease of clutter at low angles results from the cancellation of the direct and surface scattered waves as illustrated in the ideal representation given in Sec. 8.2. Measurements[41] have demonstrated this rapid decrease in sea clutter below some *critical angle* (the angle at which clutter changes from a R^{-3} dependence at short range to an R^{-7} at longer range and low grazing angle, where R = range). Not all low-angle sea clutter measurements, however, show this effect. For example, the curve for X-band sea clutter in Fig. 7.13 does not show the presence of a critical angle below which the clutter decreases rapidly.

One reason for the lack of a critical angle in some cases is that at very low grazing angles ducted propagation can occur.[42] Erroneous measurements of σ^0 can be made unless propagation effects are separated from sea-surface scattering. This requires that ducting propagation be accounted for using the proper propagation models.[43] The commonly encountered evaporation duct, which occurs a large portion of the time over most of the oceans of the world, probably is the dominant factor that gives rise to larger values of sea clutter than expected with normal (nonducting) refractive conditions.

It is difficult to obtain reliable empirical information on sea clutter at very low grazing angles that is correlated with environmental conditions. It is also difficult to obtain a theoretical understanding of the nature of the scatterers and the propagation medium at low angles. Fortunately, sea clutter is quite low at very low grazing angles and might not be a serious factor in most radar applications that have to detect targets in a sea clutter background. For example, at X band and sea state 3, σ^0 is less than -40 dB at 1.0°, less than -45 dB at 0.3°, and less than -50 dB at 0.1° grazing angle.[44] At lower frequencies, sea clutter is even less than at X band when the polarization is horizontal.

Sea Clutter at Vertical (Normal) Incidence Although sea clutter usually is much lower than land clutter at low grazing angles, it is higher than land clutter at perpendicular incidence. In the discussion of clutter at vertical (normal) incidence over a flat, perfectly reflecting land surface, it was mentioned that the value of σ^0 is approximately equal to the antenna gain G. The same is true over the sea. Thus when examining measured values of sea (or land) clutter at or near normal incidence, it needs to be kept in mind that the antenna has a significant effect on the value of the backscatter clutter power per unit area when the surface is flat or even when it is slightly rough.

Theory of Sea Clutter Many theoretical models have been proposed over the years to explain sea clutter. Most of the discussion here applies to moderate or low grazing angles. Scattering near vertical incidence generally requires a different theoretical model than at the lower grazing angles.

Past attempts to explain sea echo have been based on two different approaches. In one, the clutter is assumed to originate from scattering features at or near the sea surface. Examples include reflection from a corrugated surface,[45] backscatter from droplets of spray thrown into the air above the sea surface,[46] and backscatter from small facets, or patches, that lie on the sea surface.[41] Each of these can be applied to some limited aspect of sea echo, but none has provided an adequate explanation that accounts for the experimental evidence.

The other approach is to derive the scattering field as a boundary-value problem in which the sea surface is described by some kind of statistical process. One of the first attempts assumed that the surface disturbances could be described by a gaussian probability density function. It has been said[47] that gross observation of the sea characterized by large water wavelengths shows that the surface can be considered to be approximately gaussian, but observation of the sea's fine structure (which is what is of interest for radar backscatter) shows it is not so. Calculations of the scattering of the sea based on a gaussian surface[48] produce results that appear at first glance to be reasonable, but on close examination do not match experimental data. The conclusion that the sea cannot be represented as a gaussian statistical surface was also found by application of the theory of chaos in nonlinear dynamical systems to experimental sea clutter data.[49]

b. Derive the radar equation for detection of target in rain. (8)

Answer:

Radar Equation for Detection of Targets in Rain Rain can be a serious limitation to the detection of targets, especially at L-band and higher frequencies. The radar equation derived here indicates the important parameters that affect detection of targets in rain.

We derive the radar equation for detection of a target in rain by taking the ratio of the echo from the target and the echo from rain. It is assumed that the rain echo is much larger than receiver noise. The received signal power S from the target is (Sec. 1.2)

$$S = P_r = \frac{P_t G^2 \lambda^2 \sigma_t}{(4\pi)^3 R^4} \quad [7.35]$$

where σ_t = target radar cross section, and the other parameters are defined as in Eq. (7.24). The rain clutter C is similar to that of Eq. (7.33). and is written

$$C = \frac{K_1 P_t G \tau Z}{R^2 \lambda^2} \quad [7.36]$$

where K_1 is a constant = 12×10^{-10}. (The constant K_1 includes the velocity of propagation which has units of m/s.) The ratio of Eqs. (7.35) and (7.36) gives the signal-to-clutter ratio for a *single* pulse. If the maximum range R_{max} corresponds to the minimum detectable signal-to-clutter ratio $(S/C)_{min}$, then

$$R_{max}^2 = \frac{K_2 G \lambda^4 \sigma_t}{\tau Z (S/C)_{min}} \quad [7.37]$$

where $K_2 = 4.2 \times 10^6$. Attenuation has not been included. (For Z, one can substitute $200r^{1.6}$, or any other suitable ar^b relationship.) It should be noted that the statistics of rain echo can be different from those of receiver noise, and the value of $(S/C)_{min}$ might not be easy to determine. We see from Eq. (7.37) that for long-range detection of targets in rain, the radar wavelength should be large (low frequency), the pulse width small, and the beamwidths small (high antenna gain).

The radar equation derived above for detection of targets in rain applies for one pulse. When a number of pulses are available from a target, they may be added together (integrated) to get larger signal-to-clutter ratio if the echoes are not correlated. It was mentioned that for land and sea clutter the pulses might not be decorrelated pulse to pulse, and the use of an effective number of pulses n_{eff} must be done with caution. Rain clutter, however, is likely to be decorrelated quicker than other clutter echoes and have the statistics of the Rayleigh pdf if the radar resolution cell is not too small and the prf is not too high. Therefore, it might be appropriate to include an effective number of pulses in the numerator of Eq. (7.37) when the conditions for independent pulses apply. The decorrelation (or independence) time of rain in seconds has been said to be

$$T_t = 0.2\lambda/\sigma_v \quad [7.38]$$

where λ = radar wavelength in meters and σ_v = standard deviation of the velocity spectrum of the rain echo in m/s.[10] For example, at S band (λ = 10 cm) and σ_v = 1 m/s, the decorrelation time is 0.02 s, which means there are only 50 independent samples of rain echo available per second. (The value of σ_v depends on wind shear, turbulence, and the terminal fall velocities of the precipitation.[11] It varies from 0.5 m/s for snow to 1 m/s for rain. In convective storms it might reach 5 m/s.)

Q.7 a. Enlist all the important functions of radar antenna. (4)

Answer:

The radar antenna is a distinctive and important part of any radar. It serves the following functions:

- Acts as the transducer between propagation in space and guided-wave propagation in the transmission lines.
- Concentrates the radiated energy in the direction of the target (as measured by the antenna gain).
- Collects the echo energy scattered back to the radar from a target (as measured by the antenna effective aperture).
- Measures the angle of arrival of the received echo signal so as to provide the location of a target in azimuth, elevation, or both.
- Acts as a spatial filter to separate (resolve) targets in the angle (spatial) domain, and rejects undesired signals from directions other than the main beam.
- Provides the desired volumetric coverage of the radar.
- Usually establishes the time between radar observations of a target (revisit time).

In addition, the antenna is that part of a radar system that is most often portrayed when a picture of a radar is shown. (More can be learned about the nature of a radar from a picture of its antenna than from pictures of its equipment racks.)

With radar antennas, big is beautiful (within the limits of mechanical and electrical tolerances and the constraints imposed by the physical space available on the vehicle that carries the antenna). The larger the antenna, the better the radar performance, the smaller can be the transmitter, and the less can be the total amount of prime power needed for the radar system. The transmitting antenna gain and the receiving effective aperture are proportional to one another [as given by Eq. (1.8) or Eq. (9.9)] so that a large transmitting gain implies a *large* effective aperture, and vice-versa. As was mentioned in Chap. 1, in radar a common antenna generally has been used for both transmission and reception.

Almost all radar antennas are directive and have some means for steering the beam in angle. Directive antennas mean narrow beams, which result in accurate angular measurements and allow closely spaced targets to be resolved. An important advantage of microwave frequencies for radar is that directive antennas with narrow beamwidths can be achieved with apertures of relatively small physical size.

In this chapter, the radar antenna will be considered as either a transmitting or a receiving antenna, depending on which is more convenient for explaining a particular antenna property. Results obtained for one may be readily applied to the other because of the reciprocity theorem of antenna theory.[1]

Antenna designers have a variety of directive antenna types from which to choose including the reflector antenna in its various forms, phased arrays, endfire antennas, and lenses. They all have seen application in radar at one time or other. These antennas differ in how the radiated beam is formed and the method by which the beam is steered in angle. Steering the antenna beam can be done mechanically (by physically positioning the antenna) or electronically (by using phase shifters with a fixed phased array). The relatively simple *parabolic reflector,* similar to the automobile headlight or the searchlight, in one form or other has been a popular microwave antenna for conventional radars. As will be discussed later in this chapter, a parabolic reflector can be a paraboloid of revolution, a section of a paraboloid, a parabolic cylinder, Cassegrain configuration, parabolic torus, or a mirror scan (also called polarization-twist Cassegrain). There have also been applications of spherical reflectors, but only for special limited purposes.

The mechanically rotating array antenna was the *basis* for most of the lower frequency air-surveillance radars that saw service early in World War II. They were eventually replaced by parabolic reflector antennas when frequencies increased to the microwave region during and just after World War II. In the 1970s the mechanically scanned planar array antenna reappeared, but at microwave frequencies with slotted waveguide radiators or printed-circuit antennas rather than dipoles. The mechanically scanned planar array is found in almost all 3D radar antennas, low sidelobe antennas, and in airborne radars where the antenna is fitted behind a radome in the nose of the aircraft. (A planar aperture allows a larger antenna to be used inside the radome than is practical with a parabolic reflector.) An example of a very low sidelobe rotating planar array used in the AWACS radar is shown later in Fig. 9.49.

Starting in the mid-1960s the electronically steered phased array antenna began to be employed for some of the more demanding military radar applications. It is the most interesting and the most versatile of the various antennas, but it is also more costly and more complex.

b. When the beam of a phased array antenna is electronically steered to an angle θ_0 from broadside, show that its beam width varies inversely as cos θ_0. (8)

Answer:

Change of Beamwidth with Steering Angle As the beam of a phased array scans in angle θ_0 from broadside, its beamwidth increases as $1/(\cos\theta_0)$. This may be shown by assuming the sine in the denominator of Eq. (9.30) can be replaced by its argument, so that the radiation pattern is of the form $(\sin^2 u)/u^2$, where $u = N\pi(d/\lambda)(\sin\theta - \sin\theta_0)$. The $(\sin^2 u)/u^2$ antenna pattern is reduced to half its maximum value when $u = \pm 0.443\pi$. Denote by θ_+ the angle corresponding to the half-power point when $\theta > \theta_0$, and denote by θ_- the angle corresponding to the half-power point when $\theta < \theta_0$; that is, θ_+ corresponds to $u = +0.443\pi$ and θ_- to $u = -0.443\pi$. The $\sin\theta - \sin\theta_0$ term in the expression for u can be written[28]

$$\sin\theta - \sin\theta_0 = \sin(\theta - \theta_0)\cos\theta_0 - [1 - \cos(\theta - \theta_0)]\sin\theta_0 \qquad [9.32]$$

The second term on the right-hand side of this equation can be neglected when θ_0 is small (beam is near broadside), so that $\sin\theta - \sin\theta_0 \approx \sin(\theta - \theta_0)\cos\theta_0$. With this approximation, the two angles corresponding to the half-power (3 dB) point of the antenna pattern are

$$\theta_+ - \theta_0 = \sin^{-1}\frac{0.443\lambda}{Nd\cos\theta_0} \approx \frac{0.443\lambda}{Nd\cos\theta_0}$$

$$\theta_- - \theta_0 = \sin^{-1}\frac{-0.443\lambda}{Nd\cos\theta_0} \approx \frac{-0.443\lambda}{Nd\cos\theta_0}$$

The half-power beamwidth is

$$\theta_B = \theta_+ - \theta_- \approx \frac{0.886\lambda}{Nd\cos\theta_0} \qquad [9.33]$$

Thus when the beam is scanned an angle θ_0 from broadside, the beamwidth in the plane of scan increases as $(\cos\theta_0)^{-1}$. This expression, however, is not valid when θ_0 is large, and the array performance can be much worse. In addition to the approximation made in this derivation not being valid at large angles, mutual coupling effects can increase as the beam is scanned from broadside. At a scan angle of 60° from broadside, the beamwidth of a practical phased array antenna increases by more than the factor of 2 predicted from Eq. (9.33) and the sidelobe levels increase more than expected from simple theory.

Equation (9.33) applies for a uniform line-source distribution, which seldom is used in radar. With a cosine-on-a-pedestal aperture illumination of the form $a_0 + 2a_1\cos(2\pi n/N)$ for a linear array of N elements with spacing d, the beamwidth is approximately[29]

$$\theta_B \approx \frac{0.886\lambda}{Nd\cos\theta_0}[1 + 0.636(2a_1/a_0)^2] \qquad [9.34]$$

where a_0 and a_1 are constants, and the parameter n in the aperture illumination represents the position of the element. Since the illumination is assumed to be symmetrical about the center element, n takes on values of $0, \pm 1, \pm 2, \ldots, \pm(N-1)/2$. The antenna aperture illuminations cover the span from uniform illumination to a tapered illumination that drops to zero at the ends of the array. (The effect of the array is assumed to extend a distance $d/2$ beyond each end element.) Although the above applies to a linear array, similar results are obtained for a planar aperture; that is, the beamwidth varies approximately inversely as $\cos\theta_0$.

A consequence of the beamwidth increasing with scan angle is that the antenna gain also decreases with scan angle as $\cos\theta_0$.

c. Write a short note on the following antenna parameters: (Any two) (4)

(i) Effective aperture

(ii) Antenna radiation pattern

(iii) Power gain

Answer:

Effective Aperture The effective aperture of a receiving antenna is a measure of the effective area presented to the incident wave by the antenna. As was given previously as Eq. (1.8), the transmitting gain G and receiving effective area A_e of a lossless antenna are related by

$$G = \frac{4\pi A_e}{\lambda^2} = \frac{4\pi\rho_a A}{\lambda^2} \qquad [9.9]$$

where λ = wavelength, ρ_a = antenna aperture efficiency, A = physical area of the antenna, and $A_e = \rho_a A$. The aperture efficiency depends on the nature of the current illumination across the antenna aperture. With a uniform illumination, $\rho_a = 1$. The advantage of high efficiency obtained with a uniform illumination is tempered by the radiation pattern having a relatively high peak-sidelobe level. An aperture illumination that is maximum at the center of the aperture and tapers off in amplitude towards the edges has lower sidelobes but less efficiency than the uniform illumination.

Antenna Radiation Pattern In the above, antenna gain meant the maximum value. It is also common to speak of gain as a function of angle. Quite often the ordinate of a radiation pattern is given as the gain as a function of angle, normalized to unity. It is then known as *relative gain*. Unfortunately the term *gain* is used to denote both the maximum value and the gain as a function of angle. Uncertainty as to which usage is meant can usually be resolved from the context.

An example of an antenna radiation pattern for a paraboloidal reflector antenna is shown in Fig. 9.1.[5] This particular pattern might not be representative of a well-designed modern high-gain antenna, but it does illustrate the various features that a simple reflector-antenna radiation pattern might have. The *main beam* is shown at zero degrees. The remainder of the pattern outside the main beam is the *sidelobe* region. As the angle increases from the direction of maximum gain, there is an irregularity in this particular radiation pattern at about 22 dB below the peak. This is called a *vestigial lobe* or "shoulder" on the side of the main beam. It does not appear in all radiation patterns and is not desired since it is indicative of phase errors in the aperture illumination. Normally, when errors in the aperture illumination are small, the first sidelobe appears near where this vestigial lobe is indicated rather than where the first sidelobe is indicated in the figure.

The near-in sidelobes generally decrease in magnitude as the angle increases. The decrease is determined by the shape of the aperture illumination (as described in the next

*In some types of phased array antennas, the losses in the phase shifters and the power dividing networks can be quite high so that the difference between the power gain and the directive gain can be significant. In such cases, the directive gain is what is usually quoted and the losses are accounted for elsewhere.

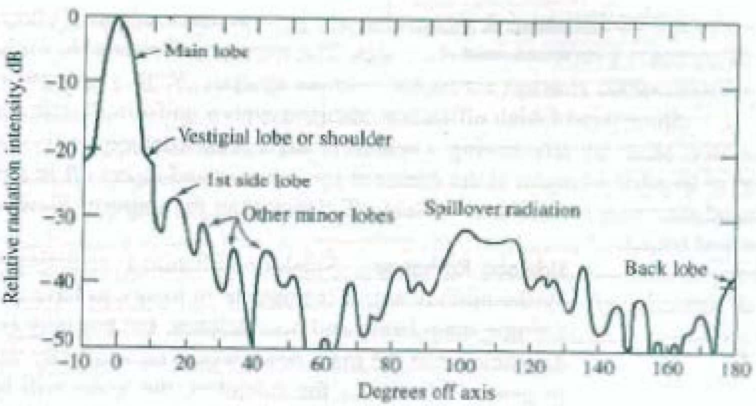


Figure 9.1 Radiation pattern for a particular paraboloid reflector antenna illustrating the main beam and sidelobe radiation. (After Cutler et al.,[5] Proc. IRE.)

section). Eventually, the sidelobes due to the aperture illumination are masked by sidelobes due to the random errors in the aperture [Sec. (9.12)]. With a conventional reflector antenna, there usually will be spillover radiation from that part of the feed radiation pattern that is not intercepted by the reflector (in the example of Fig. 9.1, this appears from about 100 to 115°). This radiation pattern also has a pronounced lobe in the backward direction (180°) due to diffraction around the edges of the reflector as well as direct leakage through the mesh reflector surface (if the surface is not solid).

The radiation pattern shown in Fig. 9.1 is plotted as a function of one angular coordinate, but the actual pattern is a plot of the radiation intensity $P(\theta,\phi)$ as a function of two angles. The two angle coordinates commonly employed with a ground-based radar antenna are azimuth and elevation, but other appropriate angle coordinates also can be used.

A complete three-dimensional plot of the radiation pattern can be complicated to display and interpret, and is not always necessary. For example, an antenna with a symmetrical pencil-beam pattern can be represented by a single plot in one angle coordinate because of its circular symmetry. The radiation intensity pattern for rectangular or rectangular-like apertures can often be written as the product of the radiation-intensity patterns in the two coordinate planes; for instance,

$$P(\theta,\phi) = P(\theta,0)\,P(0,\phi) \qquad [9.8]$$

Thus when the pattern can be expressed in this manner, the complete radiation pattern in two coordinates can be determined from the two single-coordinate patterns in the θ and in the ϕ planes.

Power Gain The power gain, which we denote by G, is similar to the directive gain except that it takes account of dissipative losses in the antenna. (It does not include loss arising from mismatch of impedances or loss due to polarization mismatch.) It can be defined similarly to the definition of directive gain, Eq. (9.2), if the denominator is the net power accepted by the antenna from the connected transmitter, or

$$G = \frac{4\pi(\text{maximum power radiated per unit solid angle})}{\text{net power accepted by the antenna}} \qquad [9.6a]$$

An equivalent definition is

$$G = \frac{\text{maximum radiation intensity from subject antenna}}{\text{radiation intensity from a lossless isotropic radiator with the same power input}} \qquad [9.6b]$$

Whenever there is a choice, the power gain should be used in the radar equation since it includes the dissipative losses introduced by the antenna. The directive gain, which is always greater than the power gain, is more closely related to the antenna beamwidth. The difference between the two antenna gains is usually small for reflector antennas. The power gain and the directive gain are related by the radiation efficiency ρ_r, as follows

$$G = \rho_r G_D \qquad [9.7]$$

The radiation efficiency is also the ratio of the total power radiated by the antenna to the net power accepted by the antenna at its terminals. The distinction between the two definitions of gain often can be ignored in practice, especially when the dissipative loss in the antenna is small.*

The definitions of power gain and directive gain described in the above were given in terms of a transmitting antenna. Because of reciprocity the pattern of a receiving antenna is the same as the pattern of a transmitting antenna, so the receiving antenna can be described by a gain just as can the transmitting antenna. This is why one can talk of a receiving gain even though gain was defined in terms of a transmitting antenna. The effective aperture of a receiving antenna, on the other hand, has no similar attribute in a transmitting antenna.

It should be kept in mind that the accuracy with which the gain of a radar antenna can be measured is usually about ± 0.5 dB.[4] Thus one should not specify or quote antenna gains to an accuracy much better than this unless there is a reason to be more accurate.

Q.8 a. What are the advantages of duplexer and receiver protector? Also, explain the working of balanced duplexer. (10)

Answer:

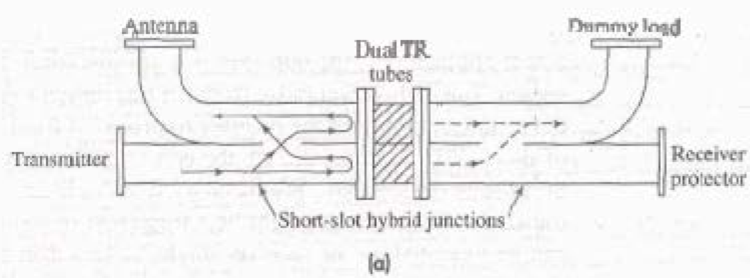
A pulse radar can time share a single antenna between the transmitter and receiver by employing a fast-acting switching device called a *duplexer*. On transmission the duplexer must protect the receiver from damage or burnout, and on reception it must channel the echo signal to the receiver and not to the transmitter. Furthermore it must accomplish the switching rapidly, in microseconds or nanoseconds, and it should be of low loss. For high-power applications, the duplexer is a gas-discharge device called a TR (transmit-receive) switch. The high-power pulse from the transmitter causes the gas-discharge device to break down and short circuit the receiver to protect it from damage. On receive, the RF circuitry of the "cold" duplexer directs the echo signal to the receiver rather than the transmitter. Solid-state devices have also been used in duplexers. In a typical duplexer application, the transmitter peak power might be a megawatt or more, and the maximum safe power that can be tolerated by the receiver might be less than a watt. The duplexer, therefore, must provide more than 60 to 70 dB of isolation between the transmitter and recovery with negligible loss on transmit and receive.

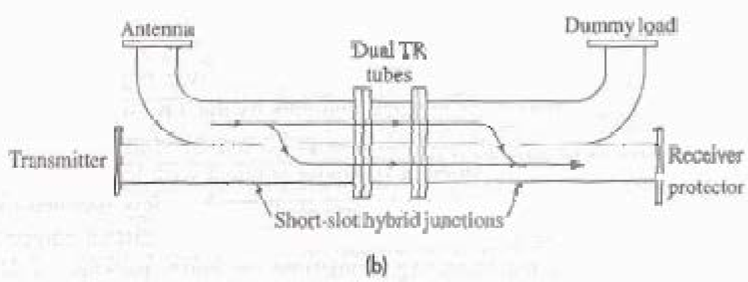
The duplexer cannot always do the entire job of protecting the receiver. In addition to the gaseous TR switch, a receiver might require diode or ferrite limiters to limit the amount of leakage that gets by the TR switch. These limiters, which have been called *receiver protectors*, also provide protection from the high-power radiation of other radars that might enter the radar antenna with less power than necessary to activate the duplexer, but with greater power than can be safely handled by the receiver. There might also be a mechanically actuated shutter to short-circuit and protect the receiver whenever the radar is not operating. Sometime the entire package of devices has been known as a *receiver* *protector.*[38] The term is ambiguous, since receiver protector is also the name for the diode limiter or similar device that follows the duplexer for the purpose of reducing the leakage power passed by the duplexer. In this text the term receiver protector is used to denote a limiter that follows the duplexer. The duplexer, receiver protector, and other devices for preventing receiver damage are better known as the *duplexer system,* so as to prevent confusion by the same term (receiver protector) being used to describe the entire receiver protection system as well as one part of it.

Balanced Duplexer The balanced duplexer, shown in Fig. 11.3, 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 wall to provide coupling between the two.[39] (The short-slot hybrid junction may be thought of as a broadband directional coupler with a coupling ratio of 3 dB.) Two TR tubes are used, one in each section of waveguide. In the transmit condition, Fig. 11.3a, power is divided equally into each waveguide by the first hybrid junction (on the left). Both gas-discharge TR tubes break down and reflect the incident power out the antenna arm as shown. The short-slot hybrid junction has the property that each time power passes through the slot in either direction, its phase is advanced by 90°. The power travels as indicated by the solid lines. Any power that 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 rubes, the hybrid junctions provide an additional 20 to 30 dB of isolation.

On reception the TR tubes do not fire and the echo signals pass through the duplexer and into the receiver as shown in Fig. 11.3b. The power splits equally at the first junction and because of the 90° phase advance on passing through the slot, the signal recombines in the receiving arm and not in the arm with the dummy load.

Figure 11.3 Balanced duplexer using dual TR tubes and two short-slot hybrid junctions. (a) Transmit condition and (b) receive condition.





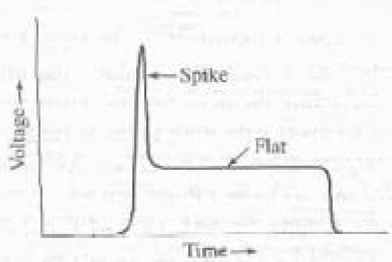
The balanced duplexer is a popular form of duplexer with good power handling capability and wide bandwidth.

TR Tube The TR tube is a gas-discharge device designed to break down and ionize quickly at the onset of high RF power, and to deionize quickly once the power is removed. One construction of a TR consists of a section of waveguide containing one or more resonant filters and two glass-to-metal windows to seal in the gas at low pressure. A noble gas like argon in the TR tube has a low breakdown voltage, and offers good receiver protection and relatively long life. TR tubes filled only with pure argon, however, have relatively long deionization times (long recovery times) and are not suitable for short-range applications. Adding water vapor or a halogen gas to the tube speeds up the deionization time, but such tubes have shorter lifetimes than tubes filled only with a noble gas. Thus a compromise must usually be accepted between fast recovery time and long life.

To insure reliable and rapid breakdown of the TR tube on application of high power, an auxiliary source of electrons is supplied to the tube to help initiate the discharge. This may be accomplished with a "keep-alive," which produces a weak d-c discharge that generates electrons that diffuse into the TR where they assist in triggering the breakdown once RF power is applied by the transmitter. An alternative is to include a small source of radioactivity, such as tritium (a radioactive isotope of hydrogen), which produces low-energy-level beta rays to generate a supply of electrons.[40] The tritium is in compounded form as a tritide film. The radioactive source, sometimes called a *tritiated ignitor*, has the advantage of not increasing the wideband noise level as does a keep-alive discharge (by about 50 K) and has longer life (by an order of magnitude), but it passes more leakage energy so that it requires one or more cascaded PIN diode limiter stages to further attenuate the leakage.[41] The tritium ignitor needs no active voltages, so it allows the receiver protector to function with the radar off without the need for a mechanical shutter to protect the radar from nearby transmissions. Being a radioactive device, however, does cause concern about its handling and disposal. The combination of the tritium-activated TR followed by a diode limiter has been called a *passive TR-limiter*.

The TR is not a perfect switch; some transmitter power always leaks through to the receiver. The envelope of the RF leakage might be similar to that shown in Fig. 11.4. The short-duration, large-amplitude *spike* at the leading edge of the leakage pulse is the result of the finite time required for the TR to ionize and break down. Typically, this time is of the order of 10 nanoseconds. After the gas in the TR tube is ionized, the power leaking through the tube is considerably reduced from the peak value of the spike. This portion

ure 11.4 Leakage pulse ough a TR tube.



of the leakage pulse is called the *flat*. Damage to the receiver front-end may result when either the energy contained within the spike or the power in the flat portion of the pulse is too large. The spike leakage of TR tubes varies with frequency and power and whether or not the tube is primed with electrons, but might be "typically" about one erg. The attenuation of the incident transmitter power might be of the order of 70 to 90 dB.

A fraction of the transmitter power incident on the TR tube is absorbed by the discharge. This is called *arc loss*. It might be 0.5 to 1 dB in tubes with water vapor and 0.1 dB or less with argon filling. On reception, the TR tube introduces an insertion loss of about 0.5 to 1 dB. The life of a TR tube is determined more by the amount of leakage power it allows to pass or when its recovery time becomes excessive rather than by its physical destruction or wear.

Solid-State Receiver Protectors, Diode Limiters Improvements in receiver sensitivity sometimes are obtained with front-ends and mixers that are more sensitive to damage from RF leakage. Such sensitive devices require better protection from the RF leakage of conventional duplexers. A PIN diode limiter placed in front of the receiver helps reduce the leakage and act as a *receiver protector*. A diode limiter passes low power with negligible attenuation, but above some threshold it attenuates the signal so as to maintain the output power constant. This property can be used for the protection of radar receivers in two different implementations depending whether the diodes are operated unbiased (self-actuated) or with a d-c forward-bias current. Unbiased operation without the use of an external current supply is also known as *passive*. It has the advantage of almost unlimited operating life, fast recovery time, no radioactive priming, and versatility to perform multiple roles.[38] Its chief limitation is its low power handling. A passive solid-state limiter for X-band WR90 waveguide with 7 percent bandwidth and a 1-μs pulse width had a peak power capability of 10 kW and a CW power capability of 10 W.[42] Its insertion loss was 0.6 dB, leakage power was 10 mW, and a 1.0-μs recovery time. When used with a 40-μs pulse width, this limiter could withstand 2 kW of peak power and 300 W CW, with a loss of 0.8 dB.

Biasing of the diodes during the high-power pulse, also known as *active*, is capable of handling a great deal more power than when operated passively. The diode is biased into its low impedance mode prior to the onset of the transmitter pulse. Although the active diode-limiter offers many advantages for use with duplexers, it does not protect the receiver when the bias is off. It thus offers poor protection against nearby asynchronous transmissions that arrive at the radar during the interpulse period or when the radar is shut down.

Incorporation of Sensitivity Time Control[43,44] In Sec. 7.8 the use of sensitivity time control (STC) was described as a method to reduce the effects of nearby large clutter echoes without seriously degrading the detection of desired targets at short range. STC is the programmed change of receiver gain with time, or range. At short ranges the receiver gain is lowered to reduce large nearby clutter echoes. As the pulse travels out in range the gain is increased until there are no more clutter echoes.

There are advantages for having the STC in the RF portion of the radar just ahead of the receiver. STC can be applied by biasing the diodes of a receiver protector to provide a time-varying attenuation without adding to the receiver noise figure. There is no increase in insertion loss to obtain the STC action, above that inherent in the design of the receiver protector. The PIN diode stages provide the self-limiting action during transmit and the STC function during receive. The nonlinear nature of the diode requires a linearizing circuit to achieve the desired variation of attenuation with time. The STC variation depends on the nature of the terrain seen by the radar. A digitally controlled STC drive with random access memory allows the radar designer to employ different STC response profiles according to the various types of terrain that might be seen by the radar.

Varactor Receiver Protectors With fast-rise-time, high-power RF sources, the receiver protector may be required to self-limit in less than one nanosecond. This can be achieved with fast-acting PN (varactor) diodes. A number of diode stages, preceded by plasma limiters, might be employed. In one design, an *X*-band passive receiver protector was capable of limiting 1-ns rise time, multikilowatt RF pulses to 1-W spike levels.[45]

Ferrite Limiters The ferrite limiter has very fast recovery time (can be as low as several tens of nanoseconds), and if the power rating is not exceeded, it should have long life. The spike and flat leakage are low and it has been able to support a peak power of 100 kW;[46] but the insertion loss is usually higher (1.5 dB) and the package is generally longer, heavier, and more expensive than other receiver protectors. Except for the initial spike, the ferrite limiter is an absorptive device rather than a reflective device (as is a gas-tube TR) so that the average power capability of these devices can be a problem. Air or liquid cooling might be required. A diode limiter usually follows the ferrite limiter to reduce the leakage at high peak power.

Pre-TR Limiter[47] A pre-TR is a gaseous tube placed in front of a solid-state limiter. The function of the pre-TR is to reduce the power that has to be handled by the diode limiter (it is similar to what was called a passive TR limiter earlier in this section. The pre-TR gas tube has high power handling capability, can operate with long pulses, has very fast recovery time, contains a radioactive priming source, but has limited operating life. Very high average power levels may require liquid cooling of the pre-TR mount. End of life for a pre-TR tube usually is caused by the increased recovery times that result from the cleanup of the gas within the tube.

The pre-TR tube can be a quartz cylinder filled with chlorine or a mixture of chlorine and an inert gas. Chlorine, a halogen gas, has a very rapid recovery time; typically a fraction of a microsecond for pulse widths up to 10 μs. The tube is mounted in a waveguide iris. In some cases, the quartz pre-TR tube can be designed to be field replaceable once it reaches end of life.

Mutipactor[46,48] The recovery times of high-power duplexers discussed thus far are from a fraction of a microsecond to several tens of microseconds. By employing the principle of multipacting, a recovery time as short as 5 or 10 ns is possible. Fast recovery time is important for high prf and high duty cycle radars. The multipactor is a vacuum tube and does not have the long recovery characteristics of a gas-filled tube. It contains surfaces capable of large secondary electron emission upon impact by electrons. The secondary emission surfaces are biased will-a d-c potential. The presence of RF energy causes elec-Irons to make multiple impacts that generates by secondary emission a large electron cloud. The electron cloud moves in phase with the oscillations of the applied RF electric field to absorb energy from the RF field. RF power is dissipated thermally at the secondary emission surfaces. and the device requires liquid cooling to remove the absorbed power. Since it is a vacuum device, the recovery time of the multipactor is extremely fast. The flat-leakage power passed by the multipactor is often high enough to require a passive diode limiter to follow it. The multipactor offers no protection when the power is turned off. It has the disadvantage of being complex in that it requires liquid cooling, an ignitor electrode to ensure that multipacting starts quickly, an oxygen source to maintain the magnesium oxide surface that provides the secondary emission electrons, and a pump to maintain a good vacuum.

Solid-State Duplexers. There has always been a desire to replace the gas-discharge duplexer with an all-solid-state duplexer because of the potential fur long life, fast recovery time, no radioactive priming, and versatility. Although passive operation is desired, it is limited in power. The lowest loss and highest power handling are obtained with active circuit!, in which the PIN diodes are switched in synchronism with the transmitter pulses. Generally, diodes that can handle high power will have longer recovery times and tend to have higher leakage power—so that they might require additional stages uf lower level limiters with increased loss and increased cost. A failure of the active drive circuit, however, could cause destruction of the diode switches as well as the receiver.

Several examples of all solid-state duplexers have been described in the literature. An *L*-band self-switching duplexer design used four PIN diodes that were biased by four fast-acting decoupled varactor detector diodes.[49] These detector diodes bias the PIN diodes into conduction in a time considerably shorter than the rise lime of the RF power pulse. The device could handle 100-kW peak power with 100-W average power and a 3-μs pulse width. Its insertion loss was 0.5 dB The duplexer was followed by a low-power multiple stage varactor limiter that reduced the spike and flat leakage of 2.8 kW and 32 W peak respectively to levels low enough that low-noise amplifiers were adequately protected. The recovery time was about 15 μs. A UHF solid-state duplexer also using four diodes was reported to have 300-kW peak power, 5-kW average power, 60-μs pulse width, and an insertion loss of 0.75 dB. A *C*-band solid-state duplexer with 16 PIN diodes was capable ul' 1-MW peak power with a 14-μs pulse width, 0.01 duty cycle, and insertion less of less than 1 dB.[50] This device was followed by an additional low-power diode switch with an insertion loss of about 0.6 dB. It provided an isolation of 60 dB, making the total isolation of the duplexer system over 100 dB.

Circulators as Duplexers The ferrite circulator is a three or four-port device that can, in principle, offer isolation of the transmitter and receiver. In the three-port circulator, the transmitter may be connected to port 1. It radiates out of the antenna connected to port 2. The received echo signal from the antenna is directed to port 3 which connects to the receiver. The isolation between the various ports might be from 20 to 30 dB. but the limitation in isolation is determined by thr reflection (due to impedance mismatch) of the transmitter signal from the antenna that is then returned directly to the receiver. For

Table 11.1 Comparison of various types of duplexing devices

Device	Recovery Time	Average Power	Peak Power
TR tube	$<1\ \mu s$ to $100\ \mu s$		1 MW
Pre-TR	50 ns to $1\ \mu s$	50 kW	5 MW
Diode limiter	50 ns to $10\ \mu s$	1 kW	100 kW
Ferrite limiter	20 ns to 120 ns	10 W	100 kW
Multipactor	1 ns to 20 ns	500 W	80 kW
Electrostatic amplifier	20 ns	300 W, or higher	10 kW, perhaps as high as 500 kW

example, if the VSWR (voltage standing wave ratio) of the antenna were 1.5 (a pretty good value), about 4 percent of the transmitter power will be reflected by the antenna and return to the receiver. This corresponds to an isolation of 14 dB. If the VSWR were 2.0, the effective isolation is only 10 dB. To limit damage, a good receiver protector needs to be included. Circulators can be made to withstand high peak and average power; but large power capability generally comes with large size and weight. For example, an *S*-band differential phase-shift waveguide circulator that weighs 80 pounds has essentially the same insertion loss, isolation, and bandwidth of an *S*-band miniature coaxial Y-junction circulator that weighs 1.5 oz.[51] The larger circulator, however, can handle 50 kW of average power while the smaller circulator is rated at 50 W. (The ratio of powers exceeds the ratio of weights.)

Small-size circulators, usually in conjunction with a receiver protector, often are used as the duplexer in solid-state TR modules for active aperture phased arrays. (Note that, unfortunately, the term "TR" has been used for both T/R modules and TR duplexer gas-discharge tubes.)

b. Enlist different types of mixers used in radar receiver. Explain how mixer works in superheterodyne receiver? (6)

Answer:

Types of Mixers[8,9] An ideal mixer is one whose output is proportional to the product of the RF echo signal and the local oscillator (LO) signal. The mixer provides two output frequencies that are the sum and difference of the two input frequencies, or $f_{RF} \pm f_{LO}$, assuming $f_{RF} > f_{LO}$. The difference frequency $f_{RF} - f_{LO}$ is the desired IF frequency. The sum frequency $f_{RF} + f_{LO}$ is rejected by filtering. There are however, two possible difference frequency signals at the IF when a signal appears at the RF. One is $f_{IF} = f_{RF} - f_{LO}$, assuming the input RF signal is of greater frequency than the LO frequency. The other possible difference frequency occurs when the RF signal is at a lower frequency than the LO frequency such that $f_{IF} = f_{LO} - f_{RF}$. If one of these is at the desired signal frequency, the other is the *image frequency*. Signals and receiver noise that appear at the image frequency need to be rejected using either an RF filter or an image-reject mixer described later in this subsection.

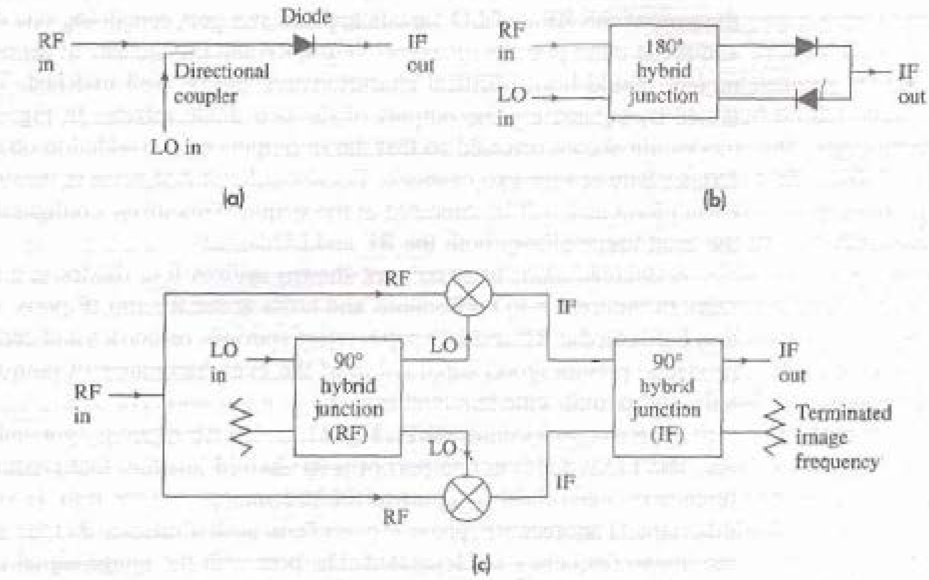
A relatively simple mixer is the *single-ended mixer*, which uses a single diode, as in Fig. 11.2a. The diode terminates a transmission line and the LO is inserted via a directional coupler. A low-pass filter, not shown, following the diode allows the IF to pass while rejecting the RF and LO signals. In a single-ended mixer the image frequency is short-circuited or open-circuited so as to avoid having the noise from the image frequency affect the mixer output.

The diode of a mixer is a nonlinear device and, in theory, can produce intermodulation products at other frequencies, called *spurious responses*. These occur for any RF signal that satisfies the relation[10]

$$mf_{RF} + nf_{LO} = f_{IF} \qquad [11.12]$$

where m and n are integers such that $m,n = \ldots, -2, -1, 0, 1, 2, \ldots$. These are unwanted since they appear within the radar receiver bandwidth. Spurious responses that are

Figure 11.2 Types of mixers: (a) single-ended mixer, (b) balanced mixer, (c) image-rejection mixer.



IF outputs due to the action of the mixer should not be confused with spurious signals, or spurs, that are due to the LO or the receiver power supply and can occur even in the absence of an RF signal. Taylor[7] describes the so-called *mixer chart*, which allows one to determine the combinations of the RF and LO frequencies that are free of strong spurious components. Such a chart indicates the bandwidth available for the mixer as a function of the ratio of the RF and LO frequencies. Taylor points out that the nature of the spurious responses are such that single-conversion receivers generally provide better suppression of spurious responses than double-conversion receivers. The third-order intermodulation product generally affects the dynamic range of the receiver, and is mentioned later under the discussion of dynamic range. There also can be other spurious, or intermodulation, responses from a mixer when two or more RF signals are present at the mixer's input and produce responses within the IF bandwidth.

Noise that accompanies the local oscillator (LO) signal in a single-ended mixer can appear at the IF frequency because of the nonlinear action of the mixer. This noise can be eliminated by inserting a narrowband RF filter between the LO and the mixer. It also has to be a tunable filter if the LO frequency is also tunable. A method to eliminate LO noise that doesn't have these disadvantages is a *balanced mixer*. The balanced mixer also can remove much of the mixer intermodulation products.

A diagram of a balanced mixer is shown in Fig. 11.2b. It can be thought of as two single-ended mixers in parallel and 180° out of phase. At the left of the figure is a four-port junction such as a magic-T, hybrid junction, 3-dB coupler, or equivalent. (Either a 90° or a 180° hybrid can be employed; here it is 180°.) In Fig. 11.2b the LO is applied to one port and the RF is applied to a second port. The signals inserted at these two ports appear in the third port as their sum and in the four port as their difference. A diode mixer is at the output of each of the other two ports. The hybrid junction has the property that the sum of the RF and LO signals appears at a port containing one of the diode mixers, and at the other port the difference of the RF and LO appears at the diode. The two diode mixers should have identical characteristics and be well matched. The IF signal is obtained by subtracting the outputs of the two diode mixers. In Fig. 11.2b, the balanced diodes are shown reversed so that the IF outputs can be added to obtain the required difference between the two channels. Local-oscillator AM noise at the two diode mixers will be in phase and will be canceled at the output. This mixer configuration also suppresses the even harmonics of both the RF and LO signals.

A *double-balanced mixer* (not shown) utilizes four diodes in a ring, or bridge, network to reduce the LO reflections and noise at the RF and IF ports, achieve better isolation between the RF and LO ports, reject spurious responses and certain intermodulation products, provide good suppression of the even harmonics of both the RF and LO signals, and permit wide bandwidth.[11]

In an *image-rejection mixer*, Fig. 11.2c, the RF signal is split and fed to the two mixers. The LO is fed into one port of a 90° hybrid junction that produces a 90° phase difference between the LO inputs to the two mixers. On the right is an IF hybrid junction that imparts another 90° phase difference in such a manner that the signal frequency and the image frequency are separated. The port with the image signal can be terminated in a matched load. According to Maas,[12] to reduce the image frequency by 20 dB requires that the phase error of the image-rejection mixer be less than 10° and the gain imbalance to be less than 1 dB. Dixon states[13] that the image-rejection mixer provides only about 30 dB of image rejection, which might not be sufficient for some applications. The image-rejection mixer is capable of wide bandwidth, and is restricted only by the frequency sensitivity of the structure of the microwave circuit. It is attractive because of its high dynamic range, good VSWR, low intermodulation products, and less susceptibility to burnout. The noise figure of the image-rejection mixer as well as the balanced mixer will be higher than that of a single-ended mixer because of the loss associated with the hybrid junctions.

The *image-recovery mixer* is an image-rejection mixer designed to reduce the mixer conversion loss by properly terminating the diode in a reactance at the image frequency. Sometimes the lower conversion loss is offset by an increase in noise temperature, a mismatch at the IF, and higher intermodulation products. The improvement using image enhancement is about 1 or 2 dB; hence, the miser needs to be of low loss so as not to negate the benefit.[14]

SUPERHETERODYNE RECEIVER

The discussion of the superheterodyne receiver in this section does not include all aspects of the receiver, but only with those component parts that have an effect on the radar system design. This includes the low-noise RF amplifier, the mixer, receiver dynamic range, the $1/f$ noise at IF, oscillator noise, and the detector.

Low-Noise Front-End The first stage of a superheterodyne receiver for radar application can be a transistor amplifier. At the lower radar frequencies the silicon bipolar transistor has been used. Gallium-arsenide field-effect transistors (FET) are found at the higher frequencies. Other types of transistors also can be used, depending on the trade-off between the desired noise figure and the ability of the transistor to withstand burnout. An *X*-band transistor can provide a noise figure of about one dB and can withstand a leakage peak power of 0.2 W.[5] With a diode limiter ahead of the transistor, the peak power can be as great as 50 W before burnout. The diode limiter increases the noise figure about 0.5 dB at *X* band and 0.2 dB at *C* band. The lower the frequency the lower can be the transistor noise figure. At *C* band the noise figure might be around 0.6 dB. These values are more than adequate for radar. (Early microwave radars had noise figures of 12 to 15 dB and radars in the 1960s had noise figures of 7 to 8 dB.) It is not necessary for the radar systems engineer to have extremely low noise figures in most radar applications, especially when the unavoidable losses in the transmission line between receiver and antenna are considered. If improved radar system performance is of concern, it is probably more fruitful to try to reduce some of the many system losses that occur elsewhere in a radar rather than try to reduce further the noise figure of the low-noise amplifier (LNA). It is usually good enough.

Prior to the low-noise transistor amplifier, the parametric amplifier and the maser were available as low-noise receiver front-ends. Although their noise figures were low (lower than those of transistors, which came later), they were seldom used operationally for radar. They were expensive, of large size, and often did not have sufficient dynamic range. Until low noise transistor amplifiers were developed, the radar receiver seldom employed an RF amplifier stage except perhaps at UHF or lower frequencies. Before the low-noise transistor, the mixer was the receiver front-end. As already mentioned, a mixer as the front-end without a low-noise amplifier preceding it is a valid option for some radar applications in spite of its higher receiver noise figure.

Achieving low receiver noise is no longer the problem it once was, and designers of high-performance radar receivers usually are more concerned with obtaining large dynamic range and low oscillator noise.

Q.9 a. What are the benefits of tracking radar? How many types of radar that can track the target? Explain in brief. (10)

Answer:

Types of Tracking Radar Systems Thus far we have considered radar mainly as a surveillance sensor that detects targets over a region of space. A radar not only recognizes the presence of a target, but it determines the target's location in range and in one or two angle coordinates. As it continues to observe a target over time, the radar can provide the target's trajectory, or *track*, and predict where it will be in the future. There are at least four types of radars that can provide the tracks of targets:

- *Single-target tracker* (STT). This tracker is designed to continuously track a single target at a relatively rapid data rate. The data rate, of course, depends on the application, but 10 observations per second might be "typical" of a military guided-missile weapon-control radar. The antenna beam of a single-target tracker follows the target by obtaining an angle-error signal and employing a closed-loop servo system to keep the error signal small. (A small angle-error signal means that the radar is accurately tracking the target.) Most of this chapter will be concerned with this type of tracker. The *C*-band AN/FPQ-6, shown in Fig. 4.1a is an example of a long-range precision tracking radar that was used at missile instrumentation ranges. The major application for continuous tracking radars has been for the tracking of aircraft and/or missile targets in support of a military weapon-control system.

- *Automatic detection and track* (ADT). This performs tracking as part of an air-surveillance radar. It is found in almost all modern civil air-traffic control radars as well as military air-surveillance radars. The rate at which observations are made depends on the time for the antenna to make one rotation (which might vary from a few seconds to as much as 12 seconds). The ADT, therefore, has a lower data rate than that of the STT, but its advantage is that it can simultaneously track a large number of targets (which might be many hundreds or a few thousands of aircraft). Tracking is done open loop in that the antenna position is not controlled by the processed tracking data as it is in the STT. This type of tracking is discussed in Sec. 4.9.
- *Phased array radar tracking.* A large number of targets can be held in track with a high data rate by an electronically steered phased array radar. Multiple targets are tracked on a time-shared basis under computer control since the beam of an electronically scanned array can be rapidly switched from one angular direction to another, sometimes in a few microseconds. It combines the rapid update rate of a single-target tracker with the ability of the ADT to hold many targets in track. This is the basis for such air-defense weapon systems as Aegis and Patriot. An example of a phased array for multiple-target tracking is the *C*-band multiple-target tracking range instrumentation radar called MOTR which is shown in Fig. 4.1b.
- *Track while scan* (TWS). This radar rapidly scans a limited angular sector to maintain tracks, with a moderate data rate, on more than one target within the coverage of the antenna. It has been used in the past for air-defense radars, aircraft landing radars, and in some airborne intercept radars to hold multiple targets in track. It is briefly mentioned in Sec. 4.7. Unfortunately, the same name *track while scan* was also applied in the past to what is now usually called ADT.

A radar can track targets in range as well as angle. Sometimes tracking of the doppler frequency shift, or the radial velocity, is also performed. Most of the discussion in this chapter, however, will be on angle tracking.

Angle-Tracking In a simple pencil-beam radar the detection of a target provides its location in angle as being somewhere within the antenna beamwidth; but more information is needed to determine the direction the antenna should be moved to maintain the target within its beam. Consider the angle measurement in a single angular coordinate. In order to determine the direction in which the antenna beam needs to be moved, a measurement has to be made at two different beam positions. Figure 4.2 shows two beam positions A and B at two different angles. The two beams are said to be *squinted*, with a squint angle $\pm\theta_q$ relative to the boresight direction. These may be two simultaneous beams, or one beam that is rapidly switched between the two angular positions. The crossover of the two beams determines the *boresight* direction. The tracking radar has to position the two beams so that the boresight is always maintained in the direction of the target; that is, the angle θ_0 is in the direction of the target angle θ_T. In this example, the relative amplitudes a_A and a_B of the echo signals received from a target measured in the two positions determine how far the target is from boresight and in what direction the two beams have to be repositioned to maintain the target on boresight. This applies for one angle coordinate. Two additional beam positions are needed in the orthogonal plane to obtain angle

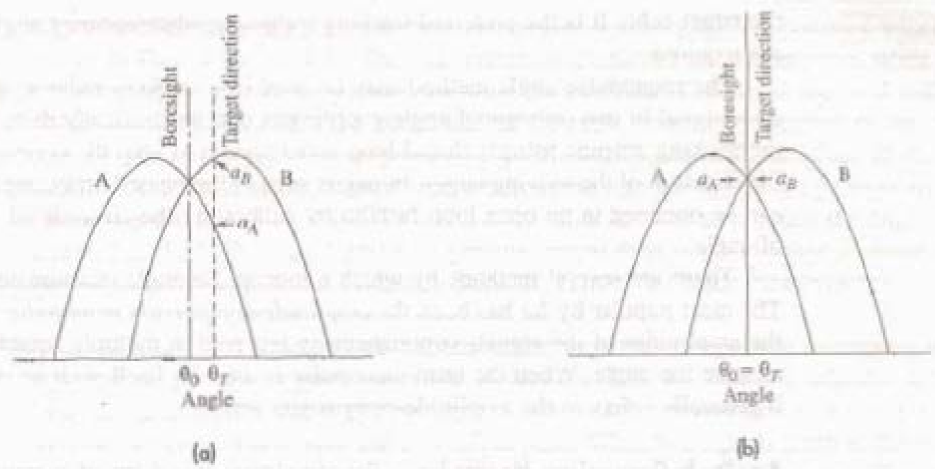


Figure 4.2 Basic principle of continuous angle tracking. (a) Two overlapping antenna patterns that cross over at the boresight direction θ_0. A target is located in this example to the right of the boresight at the angle θ_T. The amplitude a_B of the target echo in beam B is larger than the amplitude a_A in beam A, which indicates that the two beams should be moved to the right to bring the target to the boresight position. (b) Boresight position θ_0 is shown located in the direction of the target θ_T when $a_A = a_B$.

tracking in the orthogonal angle coordinate. Three beam positions are the minimum needed to obtain an angle measurement in two coordinates; but, almost always, four beams have to be used.

Early tracking radars used a single time-shared beam to track in two angles. These trackers which time share a single beam are known as either *conical scan* or *sequential lobing* trackers. (Both will be discussed later in this chapter.) Modern, high-precision tracking radars, however, use the equivalent of four simultaneous beams to perform two-dimensional tracking. They are called *simultaneous lobing* trackers, of which the most popular is *monopulse*, which is described next.

b. What is conical scan and sequential lobing? (6)

Answer:

4.3 CONICAL SCAN AND SEQUENTIAL LOBING

The monopulse tracker described in the previous section utilized multiple fixed beams to obtain the angle measurement. It is also possible to time share a single antenna beam to obtain the angle measurement in a sequential manner, as was done in early tracking radars. Time sharing a single antenna beam is simpler and uses less equipment than simultaneous beams, but it is not as accurate.

Sequential Lobing The first U.S. Army angle-tracking air-defense radar in the 1930s (SCR-268) switched a single beam between two squinted angular positions to obtain an angle measurement. This is called *lobe switching*, *sequential switching*, or *sequential lobing*. Figure 4.10a is a polar representation of the antenna beam in the two switched positions. The same in rectangular coordinates is in Fig. 4.10b. The error signal obtained

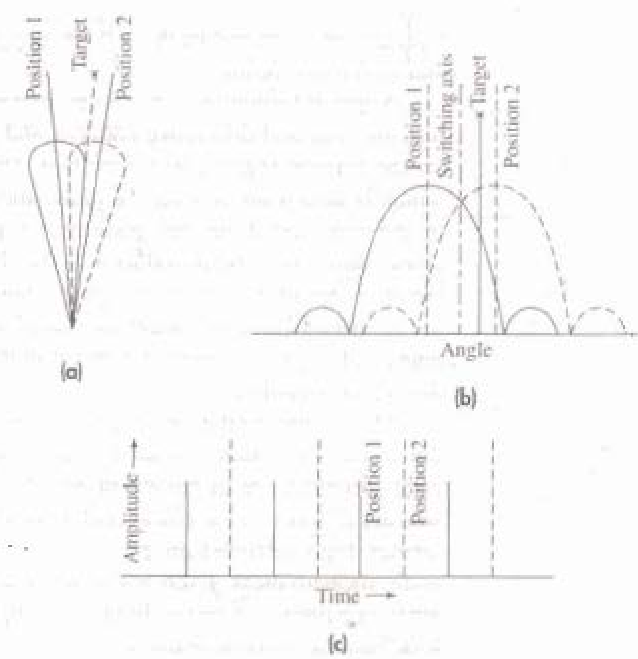


Figure 4.10 Lobe-switching antenna patterns and the error signal (for one angle coordinate). (a) Polar representation of the switched antenna pattern; (b) rectangular representation; (c) error signal.

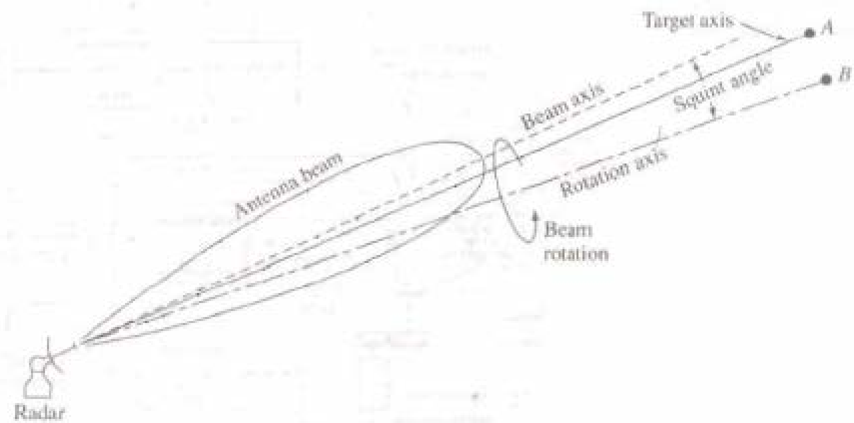
from a target not located on the switching axis (boresight) is shown in Fig. 4.10c. The difference in amplitude between the voltages obtained in the two switched positions is a measure of the angular displacement of the target from the switching axis. The direction in which to move the beam to bring the target on boresight is found by observing which beam position has the larger signal. When the echo signals in the two beam positions are equal, the target is on axis and its direction is that of the switching axis.

Two additional switching positions are needed to obtain the angle measurement in the orthogonal coordinate. Thus a two-dimensional sequentially lobing radar might consist of a cluster of four feed horns illuminating a single reflector antenna, arranged so that the right–left, up–down sectors are covered by successive antenna positions. A cluster of five feed horns might also be used, with a central feed used for transmission and four outer feeds used for reception on a sequential basis.

In a sequential lobing system, a pulse might be transmitted and received when the beam is squinted to the right, again when the beam is squinted up, when the beam is squinted to the left, and when the beam is squinted down. Thus the beam might be switched right, up, left, down, right, and so forth. After living with this type of scanning for a while, it must have become obvious that the four horns and RF switches could be replaced by a single feed that radiated a single beam squinted off axis. The squinted feed could then be continuously rotated to obtain angle measurements in two coordinates. This is a *conical-scan* radar.

Conical Scan The basic concept of conical scan, or *con-scan*, is shown in Fig. 4.11. The angle between the axis of rotation and the axis of the antenna beam is the squint angle. Consider a target located at position *A*. Because of the rotation of the squinted beam and the target's offset from the rotation axis, the amplitude of the echo signal will be modulated at a frequency equal to the beam rotation frequency (also called the conical-scan frequency). The amplitude of the modulation depends on the angular distance between the target direction and the rotation axis. The location of the target in two angle coordinates determines the phase of the conical-scan modulation relative to the conical-scan beam rotation. The conical-scan modulation is extracted from the echo signal and applied

Figure 4.11 Conical-scan tracking.



TEXT BOOK

I Introduction to Radar Systems, Merrill I. Skolnik, 3e, TMH, 2001
II Electronic and Radio Engineering, F.E. Terman, McGraw Hill Publications