RADAR

SYSTEMS

SYLLABUS

UNIT-I

Introduction, Nature of RADAR, Maximum Unambiguous range, Radar Waveforms, Block schematics of pulse radar and Operation, simple form of radar equation, RADAR frequencies, Applications of RADARS. Prediction of Range Performance, Minimum Detectable Signal, Receiver Noise, Modified radar range equation, Illustrative problems.

UNIT-II

Radar Equation: SNR, Envelop Detector, False alarm time and Probability, Integration of Radar Pulses, Radar Cross Section of Targets (simple Targets-sphere, cone-sphere),Transmitter Power, PRF and Range Ambiguities, System Losses(qualitative treatment), Illustrative Problems .

UNIT-III

CW and Frequency Modulated Radar: Doppler Effect, CW Radar – Block Diagram, Isolation between Transmitter and Receiver, Non-zero IF Receiver, Receiver bandwidth requirements, Applications of CW radar, Illustrative Problems.

UNIT-IV

FM-CW Radar, Range and Doppler measurement, Block Diagram and Characteristics (Approaching/Receding targets), FM-CW Altimeter, Measurement errors, Multiple Frequency CW Radar.

UNIT-V

MTI and Pulse Doppler radar: Introduction, Principle, MTI Radar with - Power Amplifier Transmitter and Power Oscillator Transmitter, Delay Line Cancellers – Filter Characteristics, Blind Speeds, Double Cancellation, Staggered PRFs. Range Gated Doppler Filters. MTI Radar Parameters, Limitations to MTI Performance, Non-coherent MTI, MTI versus Pulse Doppler radar.

UNIT-VI

Tracking Radar: Tracking with Radar, Sequential Lobing, Conical Scan, Monopulse Radar – Amplitude Comparison and Phase Comparison Monopulse. Low angle tracking, tracking in range, acquisition, ComparisonofTrackers.

UNIT-VII

Detection of Radar Signals in Noise: Introduction, Matched filter receiver-Response Characteristics and Derivation, Correlation Function and Cross-Correlation receiver, Efficiency of Non-matched Filters, matched Filter with Non-white Noise.

UNIT-VIII

Radar Receivers: Noise Figure and Noise temperature, Displays-types, Duplexers-Branch type and Balanced type, Circulators as Duplexers. Introduction to Phased Array Antennas-Basic Concepts, Radiation Pattern, Beam Steering and Beam Width Changes, Series versus Parallel Feeds, Applications, Advantages and Limitations.

UNIT-I

NATURE OF RADAR

INTRODUCTION:

The name Radar stands for Radio Detection and Ranging

Radar is a remote sensing technique: Capable of gathering information about objects located at remote distances from the sensing device.

Two distinguishing characteristics:

1. Employs EM waves that fall into the microwave portion of the electromagnetic spectrum

   $(1\ \text{mm} < l < 75\ \text{cm})$

2. Active technique: radiation is emitted by radar – radiation scattered by objects is detected by radar.

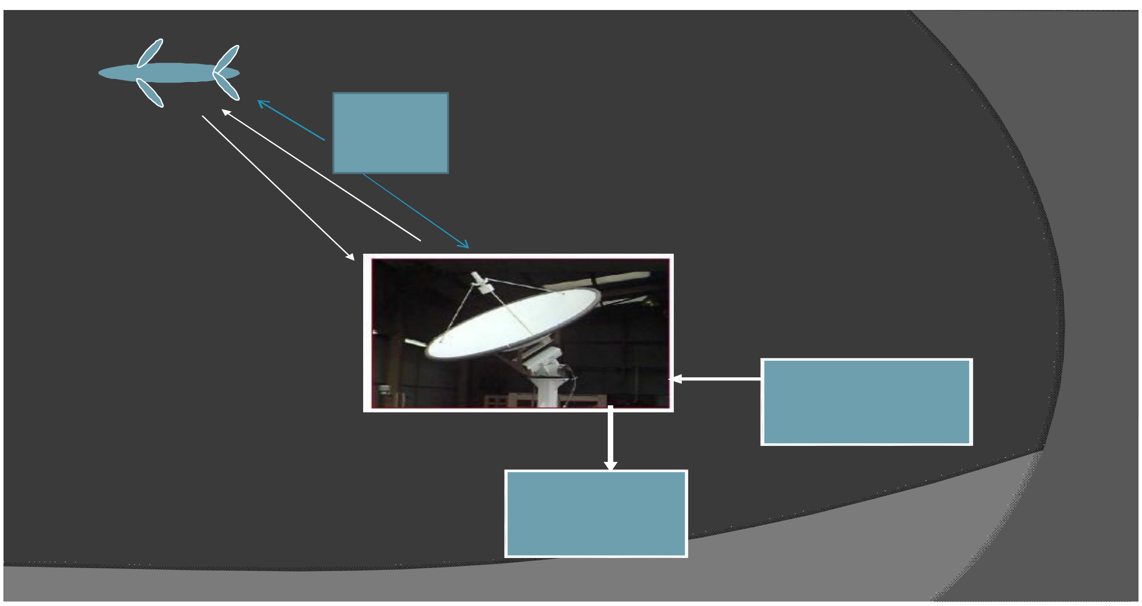
Radar is an electromagnetic system for the detection and location of objects (Radio Detection and Ranging). Radar operates by transmitting a particular type of waveform and detecting the nature of the signals reflected back from objects.

Radar can’t resolve detail or color as well as the human eye (an optical frequency passive scatter meter).

Radar can see in conditions which do not permit the eye to see such as darkness, haze, rain, smoke.

Radar can also measure the distances to objects. The elemental radar system consists of a transmitter unit, an antenna for emitting electromagnetic radiation and receiving the echo, an

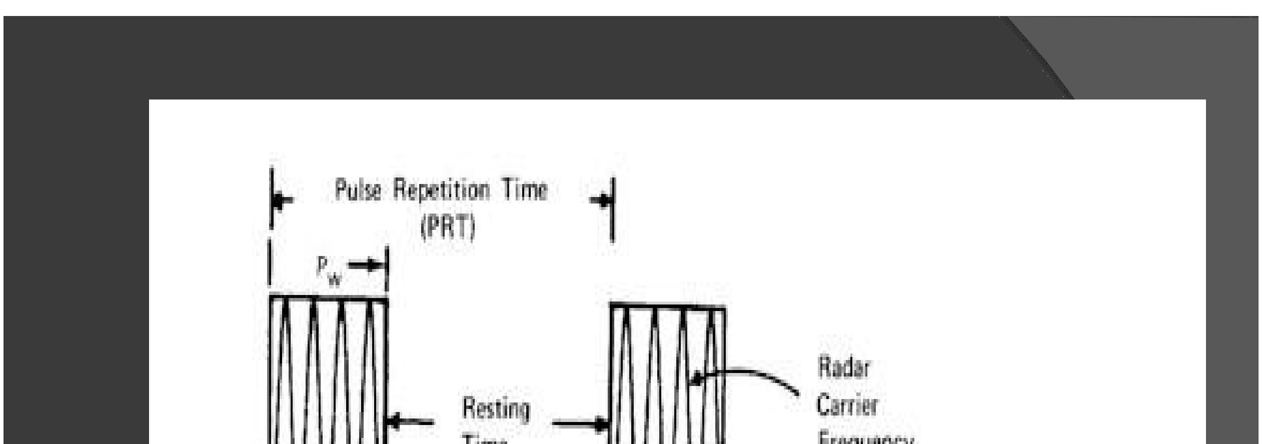
Energy detecting receiver and a processor.



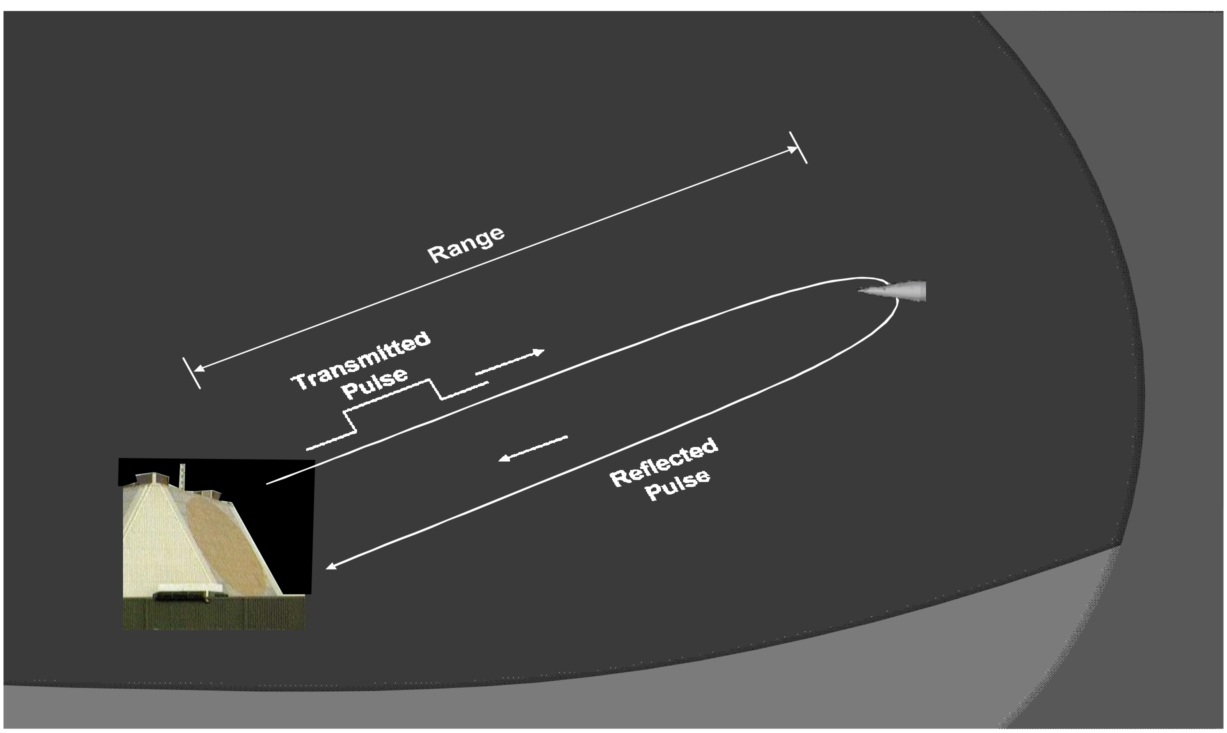
A portion of the transmitted signal is intercepted by a reflecting object (target) and is reradiated in all directions. The antenna collects the returned energy in the backscatter direction and delivers it to the receiver. The distance to the receiver is determined by measuring the time taken for the electromagnetic signal to travel to the target and back.

The direction of the target is determined by the angle of arrival (AOA) of the reflected signal. Also if there is relative motion between the radar and the target, there is a shift in frequency of the reflected signal (Doppler Effect) which is a measure of the radial component of the relative velocity. This can be used to distinguish between moving targets and stationary ones.

Radar was first developed to warn of the approach of hostile aircraft and for directing anti aircraft weapons. Modern radars can provide AOA, Doppler, and MTI etc.



RADAR RANGE MEASUREMENT


Range
Transmitted
Pulse
Reflected
Pulse

The simplest radar waveform is a train of narrow (0.1μs to 10μs) rectangular pulses modulating a sinusoidal carrier the distance to the target is determined from the time T_R taken by the pulse to travel to the target and return and from the knowledge that electromagnetic energy travels at the speed of light.

Since radio waves travel at the speed of light (v = c = 300,000 km/sec)

Range = c×time/2

The range or distance, $\mathbf{R = cT_R/2}$

$$\mathbf{R\ (in\ km) = 0.15T_R\ (\mu s)\ ;\ R\ (in\ nmi) = 0.081T_R\ (\mu s)}$$

NOTE:

- 1 nmi = 6076 feet =1852 meters.
- 1 Radar mile = 2000 yards = 6000 feet
- **Radar mile** is commonly used unit of distance.

NOTE:

Electromagnetic energy travels through air at approximately the speed of light:-

1. 300,000 kilometers per second.

2. 186,000 statute miles per second.

3. 162,000 nautical miles per second.

Once the pulse is transmitted by the radar a sufficient length of time must elapse before the next pulse to allow echoes from targets at the maximum range to be detected. Thus the maximum rate at which pulses can be transmitted is determined by the maximum range at which targets are expected. This rate is called the pulse repetition rate (PRF).

If the PRF is too high echo signals from some targets may arrive after the transmission of the next pulse. This leads to ambiguous range measurements. Such pulses are called second time around pulses.

The range beyond which second time around pulses occur is called the maximum unambiguous range.

$$\mathbf{R_{UNAMBIG} = c/2f_P}$$

Where f_P is the PRF in Hz.

More advanced signal waveforms then the above are often used, for example the carrier maybe frequency modulated (FM or chirp) or phase modulated (pseudorandom bi phase) too permit the echo signals to be compressed in time after reception. This achieves high range resolution without the need for short pulses and hence allows the use of the higher energy of longer pulses. This technique is called pulse compression. Also CW waveforms can be used by taking advantage of the Doppler shift to separate the received echo from the transmitted signal.

Note: unmodulated CW waveforms do not permit the measurement of range.

What is done by Radar?

Radar can see the objects in

- day or night
- rain or shine
- land or air
- cloud or clutter
- fog or frost
- earth or planets
- stationary or moving and
- Good or bad weather.

In brief, Radar can see the objects hidden any where in the globe or planets except hidden behind good conductors.

INFORMATION GIVEN BY THE RADAR:

Radar gives the following information:

- ✓ The position of the object
- ✓ The distance of objects from the location of radar
- ✓ The size of the object
- ✓ Whether the object is stationary or moving
- ✓ Velocity of the object
- ✓ Distinguish friendly and enemy aircrafts
- ✓ The images of scenes at long range in good and adverse weather conditions
- ✓ Target recognition
- ✓ Weather target is moving towards the radar or moving away
- ✓ The direction of movement of targets
- ✓ Classification of materials

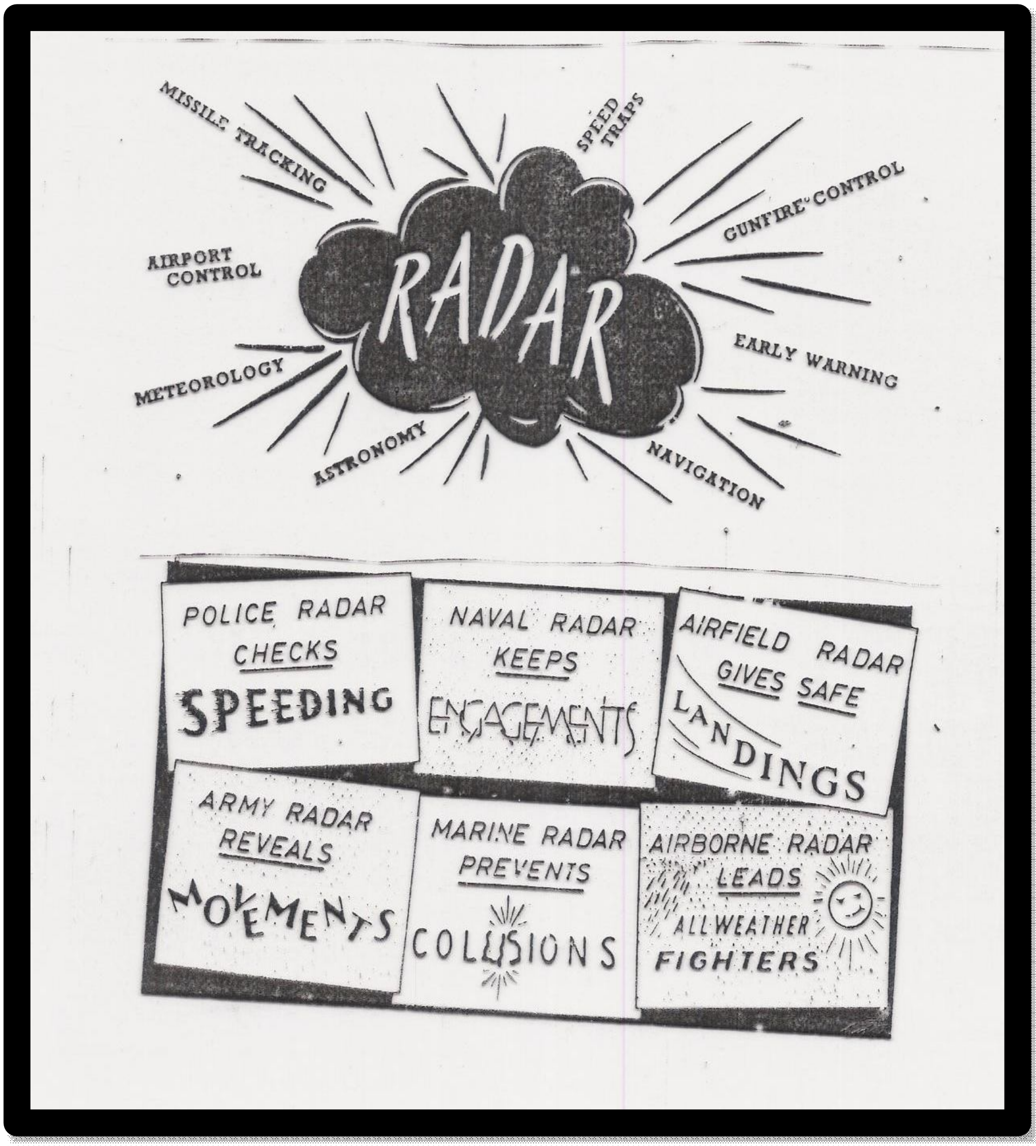
NATURE AND TYPES OF RADARS:

The common types of radars are:

- Speed trap Radars
- Missile tracking Radars
- Early warning Radars

- Airport control Radars
- Navigation Radars
- Ground mapping Radars
- Astronomy Radars
- Weather forecast Radars
- Gun fire control Radars
- Remote sensing Radars
- Tracking Radars
- Search Radars
- IFF (Identification Friend or Foe)
- Synthetic aperture Radars
- Missile control Radars
- MTI (Moving Target Indication) Radars
- Navy Radars
- Doppler Radars
- Mesosphere, Stratosphere and Troposphere (MST) Radars
- Over-The-Horizon (OTH) Radars
- Mono pulse Radars

- Phased array Radars
- Instrumentation Radars
- Gun direction Radars
- Airborne weather Radars



PULSE CHARACTERISTICS OF RADAR SYSTEMS:

There are different pulse characteristics and factors that govern them in a Radar system

- Carrier
- Pulse width
- Pulse Repetition Frequency (PRF)
- Unambiguous Range

NOTE: ECHO is a reflected EM wave from a target and it is received by a Radar receiver.

CARRIER: The carrier is used in a Radar system is an RF(radio frequency) signal with microwave frequencies.

Carrier is usually modulated to allow the system to capture the required data.

In simple ranging Radars, the carrier will be pulse modulated but in continuous wave systems such Doppler radar modulation is not required.

In pulse modulation, the carrier is simply switched ON & OFF in synchronization.

PULSE WIDTH: The pulse width of the transmitted signal determines the dead zone. When the Radar transmitter is active, the receiver input is blanked to avoid the damage of amplifiers. For example, a Radar echo will take approximately 10.8 µsec to return from 1 standard mile away target.

PULSE REPETITION FREQUENCY (PRF): PRF is the number of pulses transmitted per second. PRF is equal to the reciprocal of pulse repetition time (PRT). It is measured in Hertz

$$PRF = 1/PRT$$

Pulse Interval Time or Pulse Reset Time (PRT) is the time interval between two pulses. It is expressed in milliseconds.

Pulse Reset Time = Pulse Repetition Time – Pulse Width

UNAMBIGUOUS RANGE: In simple systems, echoes from targets must be detected and processed before the next transmitter pulse is generated if range ambiguity is to be avoided.

Range ambiguity occurs when the time taken for an echo to return from a target is greater than the pulse repetition period (T).

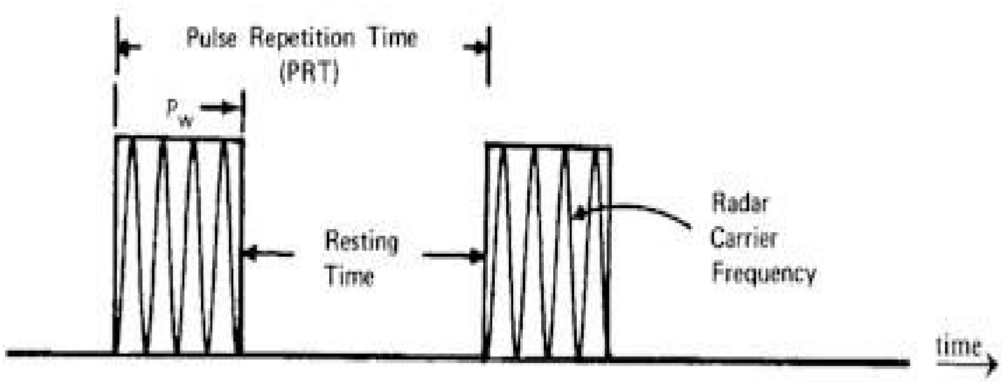


Figure 2–1. Pulse transmission.

Echoes that arrive after the transmission of the next pulse are called as second-time-around echoes.

The range beyond which targets appear as second-time-around echoes is called as the Maximum

Unambiguous Range and is given by

$$R_{UNAMBIG} = c/2f_P$$

c = velocity of propagation

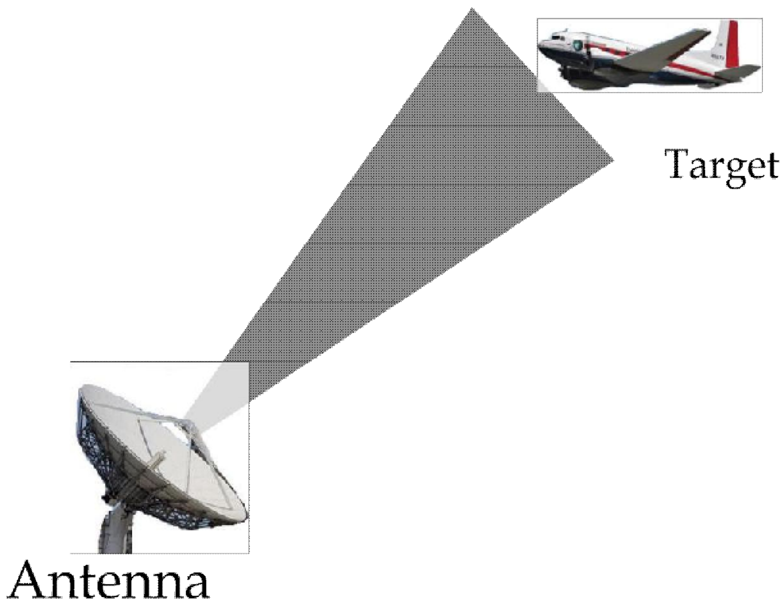
Where, $T_P = f_P$

f_P is the PRF(PULSE REPETITION FREQUENCY) in Hz

TYPES OF BASIC RADARS:

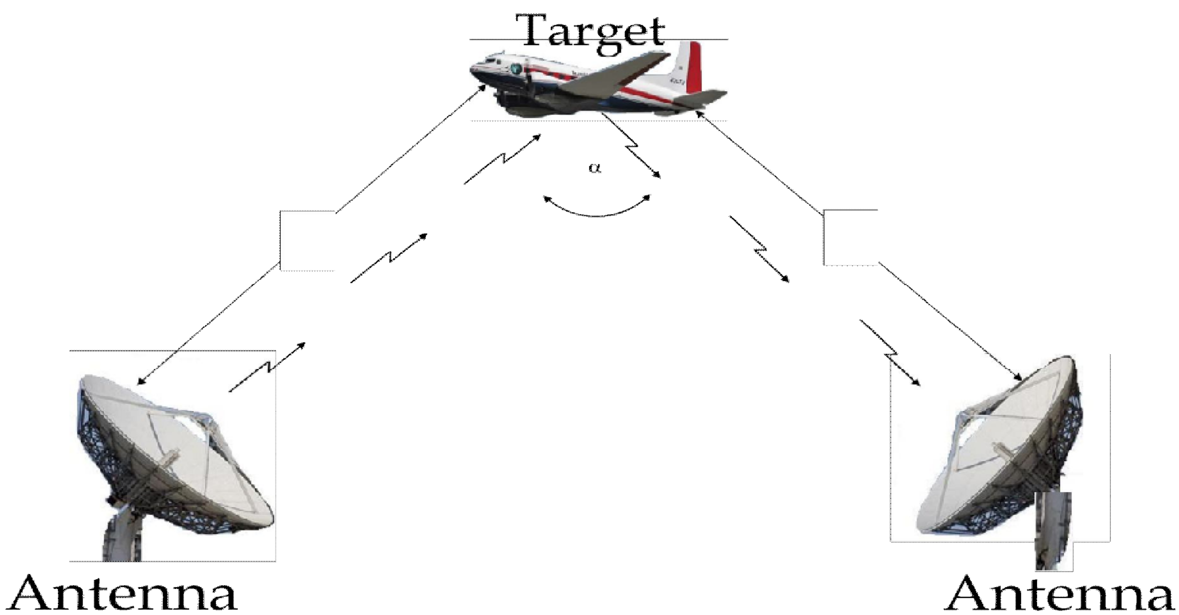
- Monostatic and Bistatic
- CW
- FM-CW
- Pulsed radar

Monostatic radar uses the same antenna for transmit and receive. Its typical geometry is shown in the below fig.



Bistatic radars use transmitting and receiving antennas placed in different locations.

CW radars, in which the two antennas are used, are not considered to be bistatic radars as the distance between the antennas is not considerable. The bistatic radar geometry is shown in below fig.


Target
α
Antenna
Antenna

RADAR WAVE FORMS:

The most common Radar waveform is a train of narrow, rectangular shape pulses modulating a sine-wave carrier.

The figure shows a pulse waveform, which can be utilized by the typical Radar.

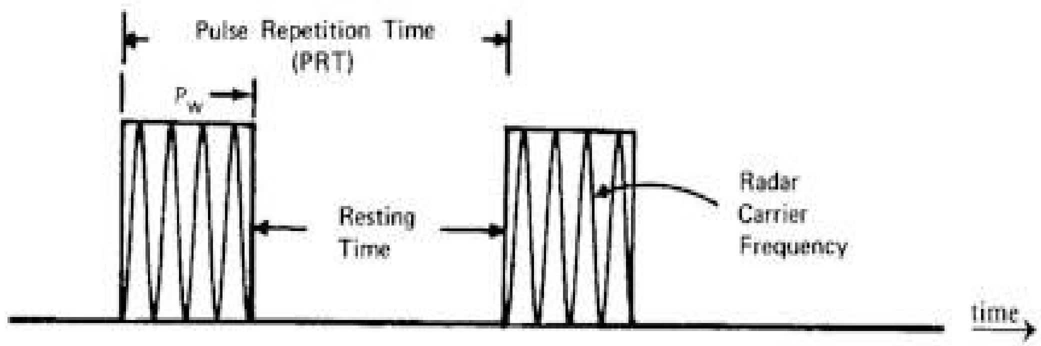
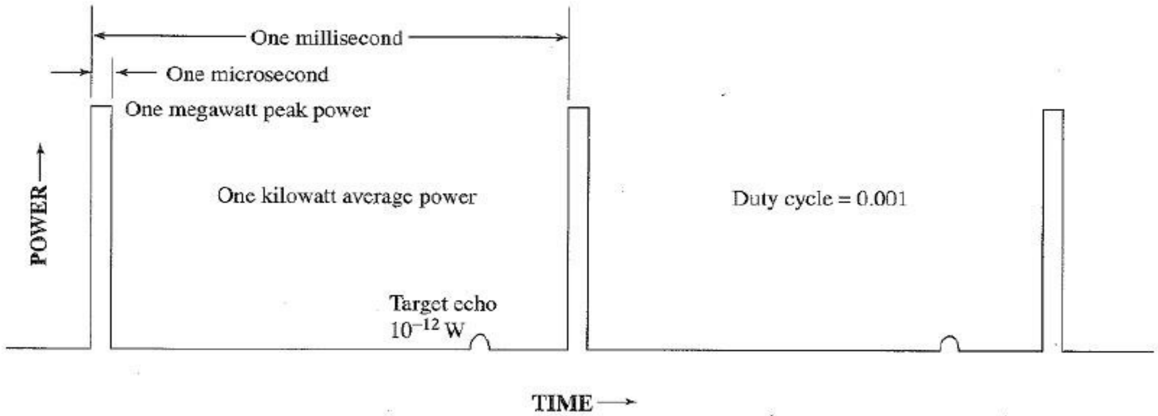


Figure 2–1. Pulse transmission.


One millisecond
One microsecond
One megawatt peak power
POWER
One kilowatt average power
Duty cycle = 0.001
Target echo
10^{-12} W
TIME

From the given Radar waveform:

Peak power $p_t = 1$ Mwatt

Pulse Width $\tau = 1$ μsec.

Pulse Repetition Period $T_P = 1$ msec.

A maximum unambiguous range of 150 km was provided by the PRF $f_P = 1000$ Hz.

$$R_{UNAMBIG} = c/2f_P \implies 150\times10^3 = 3\times10^8 / 2f_P$$

$$\implies f_P = 1000 \text{ Hz.}$$

Then, the average power P_{avg} of a repetitive pulse train wave form is given by $P_{avg} = p_t \tau / T_P \implies P_{avg} = p_t \tau f_P$

In this case, $P_{avg} = 1$ Kwatt

For a Radar wave form, the ratio of the total time that the Radar is radiating to the total time it could have radiated is known as duty cycle.

Duty Cycle = $\tau/T_P = \tau f_P = P_{avg}/p_t$

Duty Cycle = $\tau/T_P = 0.001$

The energy of the pulse is given by, $E = \tau p_t = 1$ Joule.

The Radar waveform can be extended in space over a distance of 300 meters using a pulse width of 1 μsec.

i.e., Distance = $c\,\tau$ = 300 m.

Half of the above distance (i.e. $c\,\tau/2$) can be used to recognize the two equal targets which are being resolved in range. In this case, a separation of 150m between two equal size targets can be used to resolve them.

Name	Symbol	Units	Typical values
Transmitted Frequency	f_t	MHz, G hz	1000-12500 Mhz
Wavelength	□	cm	3-10 cm
Pulse Duration	□	□ sec	1 □ sec
Pulse Length	h	m	150-300 m (h=c□□
Pulse Repetition Frequency	PRF	sec^{-1}	1000 sec^{-1}
Interpulse Period	T	Milli sec	1 milli sec
Peak Transmitted Power	P_t	MW	1 MW
Average Power	P_{avg}	kW	1 kW (P_{avg} = P_t □ PRF)
Received Power	P_r	mW	10^{-6} mW

The Radar Range Equation:

The radar range equation relates the range of the radar to the characteristics of the transmitter, receiver, antenna, target and the environment.

It is used as a tool to help in specifying radar subsystem specifications in the design phase of a program. If the transmitter delivers P_T Watts into an isotropic antenna, then the power density (w/m^2) at a distance R from the radar is

$$P_t/4\pi R^2$$

Here the $4\pi R^2$ represents the surface area of the sphere at distance R

Radars employ directional antennas to channel the radiated power Pt in a particular direction. The gain G of an antenna is the measure of the increased power radiated in the direction of the target, compared to the power that would have been radiated from an isotropic antenna

∴ Power density from a directional antenna = $P_tG/4\pi R^2$

The target intercepts a portion of the incident power and redirects it in various directions.

The measure of the amount of incident power by the target and redirected back in the direction of the radar is called the cross section σ.

Hence the Power density of the echo signals at the radar = $\frac{P_tG}{4\pi R^2}\frac{\sigma}{4\pi R^2}$

Note: the radar cross-section σ has the units of area. It can be thought of as the size of the target

as seen by the radar.

The receiving antenna effectively intercepts the power of the echo signal at the radar over a certain area called the effective area $A_{e.}$

Since the power density (Watts/m^2) is intercepted across an area A_e, the power delivered to the receiver is

$$P_r = \frac{P_tG}{4\pi R^2}\frac{\sigma}{4\pi R^2}A_e$$

$P_r = (P_tG\sigma A_e) /(4\pi R^2)^2 \implies R^4 = (P_tG\sigma A_e) /(4\pi)^2 P_r$

$$R = [(P_t G \sigma A_e) / (4\pi)^2 P_r]^{1/4}$$

Now the maximum range R_{max} is the distance beyond which the target cannot be detected due to insufficient received power P_r, the minimum power which the receiver can detect is called the minimum detectable signal S_{min}. Setting, $P_r = S_{min}$ and rearranging the above equation gives

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

Note here that we have both the antenna gain on transmit and its effective area on receive.

These are related by: $\frac{G}{4\pi} = \frac{A_e}{\lambda^2}$

As long as the radar uses the same antenna for transmission and reception we have

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

Example: Use the radar range equation to determine the required transmit power for the TRACS

radar given: Prmin $=10^{-13}$ Watts, G=2000, λ=0.23m, PRF=524, σ=2.0 m^2

Now, $R_{max} = \frac{c}{PRF}$

From

$$P_t = \frac{P_r (4\pi)^3 R^4}{G^2 \lambda^2 (PRF)^4 \sigma}$$

$$P_t = \frac{(10^{-13})(4\pi)^3 \left(\frac{3(10^8)}{2}\right)^4}{(2000)^2 (0.23)^2 (524)^4 (2.0)} \quad = 3.1 \text{ MW}$$

Note 1: these three forms of the equation for Rmax varywith different powers of λ. This results

from implicit assumptions about the independence of G or Ae from λ.

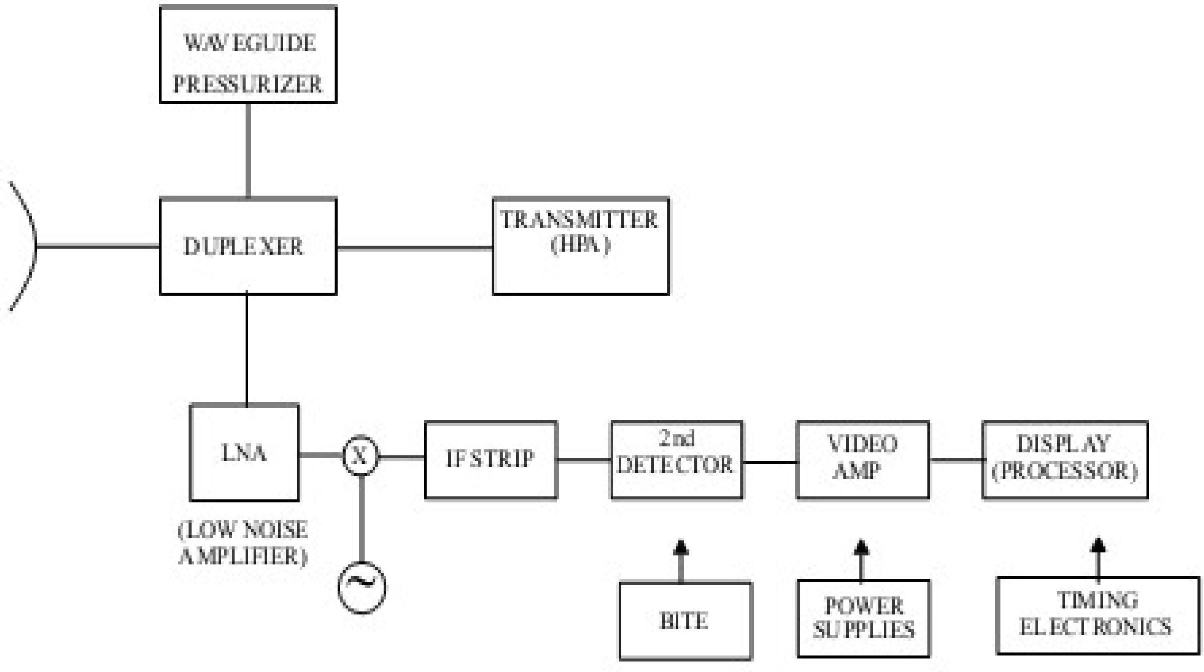
Note 2: the introduction of additional constraints (such as the requirement to scan a specific volume of space in a given time) can yield other λ dependence.

Note 3: The observed maximum range is often much smaller than that predicted from the above equation due to the exclusion of factors such as rainfall attenuation, clutter, noise figure etc.

RADAR BLOCK DIAGRAM AND OPERATION:

The Transmitter may be an oscillator (magnetron) that is pulsed on and off bya modulator to generate the pulse train.

- the magnetron is the most widely used oscillator
- typical power required to detect a target at 200 NM is MW peak power and several kW average power
- typical pulse lengths are several μs
- typical PRFs are several hundreds of pulses per second



The waveform travels to the antenna where it is radiated. The receiver must be protected from damage resulting from the high power of the transmitter. This is done by the duplexer.

- duplexer also channels the return echo signals to the receiver and not to the transmitter

- duplexer consists of 2 gas discharge tubes called the TR (transmit/receive) and the and an

  ATR (anti transmit/receive) cell

- The TR protects the receiver during transmission and the ATR directs the echo to the receiver during reception.
- solid state ferrite circulators and receiver protectors with gas plasma (radioactive keep alive) tubes are also used in duplexers

The receiver is usually a superheterodyne type. The LNA is not always desirable. Although it provides better sensitivity, it reduces the dy namic range of operation of the mix er. A receiver with just a mixer front end has greater dynamic range, is less susceptible to overload and is less vulnerable to electronic interference.

The mixer and Local Oscillator (LO) convert the RF frequency to the IF frequency.

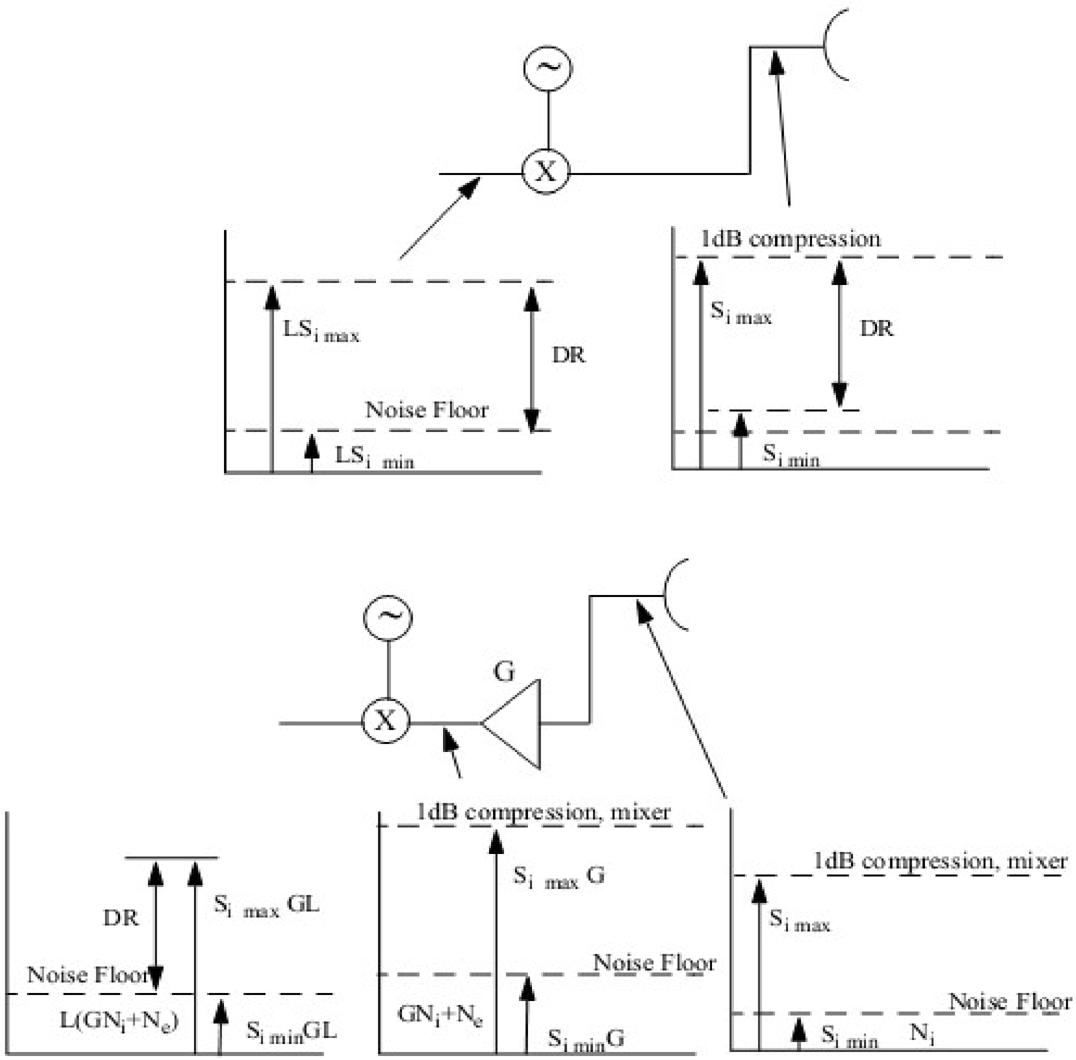
- The IF is typically 300MHz, 140Mz, 60 MHz, 30 MHz with bandwidths of 1 MHz to 10 MHz.

- The IF strip should be designed to give a matched filter output. This requires its H(f) to maximize the signal to noise power ratio at the output.
- This occurs if the |H(f)| (magnitude of the frequency response of the IF strip is equal to the signal spectrum of the echo signal |S(f)|, and the ARG(H(f)) (phase of the frequency response) is the negative of the ARG(S(f)).

i.e. H(f) and S(f) should be complex conjugates

- For radar with rectangular pulses, a conventional IF filter characteristic approximates a matched filter if its bandwidth B and the pulse width τ satisfy the relationship

EFFECT OF LNA ON DYNAMIC RANGE



The pulse modulation is extracted by the second detector and amplified by video amplifiers to levels at which they can be displayed (or A to D'd to a digital processor). The display is usually a CRT; timing signals are applied to the display to provide zero range information. Angle information is supplied from the pointing direction of the antenna.

- The most common type of CRT display is the plan position indicator (PPI) which maps the location of the target in azimuth and range in polar coordinates
- The PPI is intensitymodulated bythe amplitude of the receiver output and the CRT

  electron beam sweeps outward from the centre corresponding to range.

- Also the beam rotates in angle in synchronization with the antenna pointing angle.
- A B scope display uses rectangular coordinates to display range vs angle i.e. the x axis is angle and the y axis is range.
- Since both the PPI and B scopes use intensity modulation the dynamic range is limited
- An A scope plots target echo amplitude vs range on rectangular coordinates for some fixed direction. It is used primarily for tracking radar applications than for surveillance radar.

The simple diagram has left out many details such as

- AFC to compensate the receiver automatically for changes in the transmitter
- AGC
- Circuits in the receiver to reduce interference from other radars

Rotary joints in the transmission lines to allow for movement of the antenna

- MTI (moving target indicator) circuits to discriminate between moving targets and unwanted stationary targets
- Pulse compression to achieve the resolution benefits of a short pulse but with the energy benefits of a long pulse.
- Monopulse tracking circuits for sensing the angular location of a moving target and allowing the antenna to lock on and track the target automatically
- Monitoring devices to monitor transmitter pulse shape, power load and receiver sensitivity

- Built in test equipment (BITE) for locating equipment failures so that faulty circuits can be replaced quickly

Instead of displaying the raw video output directly on the CRT, it might be digitized and processed and then displayed. This consists of:

- Quantizing the echo level at range-azimuth resolution cells
- Adding (integrating) the echo level in each cell
- Establishing a threshold level that permits only the strong outputs due to target echoes to pass while rejecting noise
- Maintaining the tracks (trajectories) of each target
- Displaying the processed information

This process is called automatic tracking and detection (ATD) in surveillance radar

Antennas:

- The most common form of radar antenna is a reflector with parabolic shape, fed from a point source (horn) at its focus

- The beam is scanned in space by mechanically pointing the antenna
- Phased array antennas are sometimes used. Her the beam is scanned by varying the phase of the array elements electrically

Radar Frequencies:

- Most Radar operates between 220 MHz and 35 GHz.
- Special purpose radars operate out side of this range.
  - Skywave HF-OTH (over the horizon) can operate as low as 4 MHz
  - Groundwave HF radars operate as low as 2 MHz
  - Millimeter radars operate up to 95 GHz
  - Laser radars (lidars) operate in IR and visible spectrum

The radar frequencyletter -band nomenclature is shown in the table. Note that the frequencyassignment to the latter band radar (e.g. L band radar) is much smaller than the complete range of frequencies assigned to the letter band

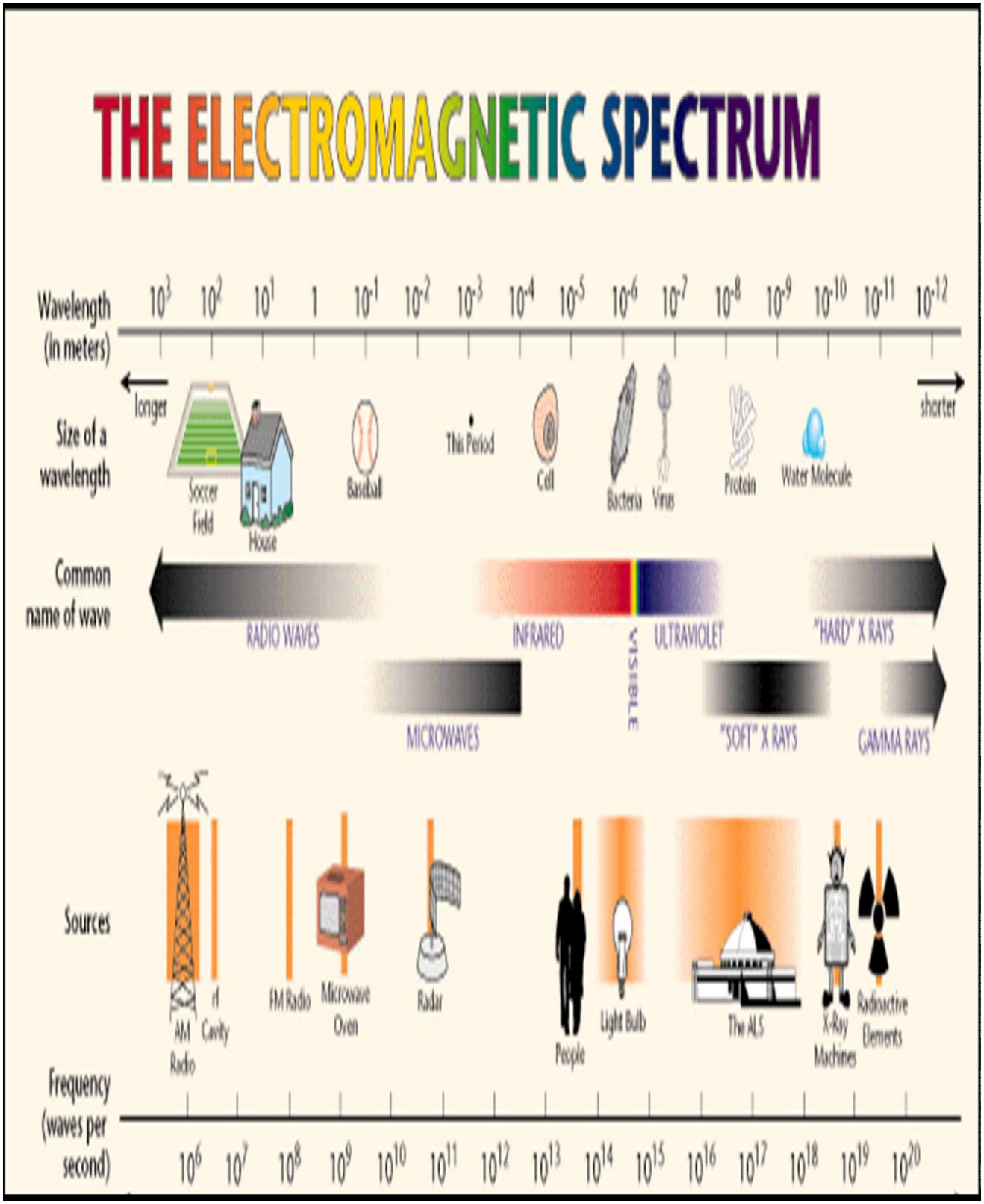


Table 1:

Band Designation	Nominal Frequency Range	Specific radar bands based on ITU assignments for region 2
HF	3-30 MHz	
VHF	30-300 MHz	138-144 MHz 216-225 MHz
UHF	300-1000 MHz	420-450 MHz 890-942 MHz
L	1000 - 2000 MHz	1215-1400 MHz
S	2000 - 4000 MHz	2300 - 2500 MHz 2700 - 3700 MHz
K_u	12 - 18 GHz	13.4 - 14.0 GHz 15.7 - 17.7 GHz
K	18 - 27 GHz	24.05 - 24.25 GHz
K_a	40 - 300 GHz	33.4 - 36.0 GHz
mm	40 - 300 GHz	

Applications of Radar

General

i. Ground-based radar is applied chiefly to the detection, location and tracking of aircraft of space targets

ii. Shipborne radar is used as a navigation aid and safety device to locate buoys, shorelines and other ships. It is also used to observe aircraft

iii. Airborne radar is used to detect other aircraft, ships and land vehicles. It is also used for mapping of terrain and avoidance of thunderstorms and terrain.

iv. Spaceborne radar is used for the remote sensing of terrain and sea, and for rendezvous/docking.

Major Applications

1. Air Traffic Control

- Used to provide air traffic controllers with position and other information on aircraft flying within their area of responsibility (airways and in the vicinity of airports)
- High resolution radar is used at large airports to monitor aircraft and ground vehicles on the runways, taxiways and ramps.
- GCA (ground controlled approach) or PAR (precision approach radar) provides an operator with high accuracy aircraft position information in both the vertical and horizontal. The operator uses this information to guide the aircraft to a landing in bad weather.
- MLS (microwave landing system) and ATC radar beacon systems are based on radar technology

2. Air Navigation

- Weather avoidance radar is used on aircraft to detect and display areas of heavy precipitation and turbulence.
- Terrain avoidance and terrain following radar (primarily military)
- Radio altimeter (FM/CW or pulse)
- Doppler navigator
- Ground mapping radar of moderate resolution sometimes used for navigation

3. Ship Safety

- These are one of the least expensive, most reliable and largest applications of radar
- Detecting other craft and buoys to avoid collision
- Automatic detection and tracking equipment (also called plot extractors) are available with these radars for collision avoidance
- Shore based radars of moderate resolution are used from harbour surveillance and

  as an aid to navigation

4. Space

- Radars are used for rendezvous and docking and was used for landing on the moon

- Large ground based radars are used for detection and tracking of satellites
- Satellite-borne radars are used for remote sensing (SAR, synthetic aperture radar)

5. Remote Sensing

- Used for sensing geophysical objects (the environment)
- Radar astronomy - to probe the moon and planets
- Ionospheric sounder (used to determine the best frequency to use for HF communications)
- Earth resources monitoring radars measure and map sea conditions, water resources, ice cover, agricultural land use, forest conditions, geological formations, environmental pollution (Synthetic Aperture Radar, SAR and Side Looking Airborne Radar SLAR)

6. Law Enforcement

- Automobile speed radars
- Intrusion alarm systems

7. Military

- Surveillance
- Navigation, Fire control and guidance of weapons

ADVANTAGES OF BASIC RADAR:

- It acts as a powerful eye.
- It can see through: fog, rain, snow, darkness, haze, clouds and any insulators.
- It can find out the range, angular position, location and velocity of targets.

LIMITATIONS:

- Radar can not recognize the color of the targets.
- It can not resolve the targets at short distances like human eye.
- It can not see targets placed behind the conducting sheets.
- It can not see targets hidden in water at long ranges.

- It is difficult to identify short range objects.

- The duplexer in radar provides switching between the transmitter and receiver alternatively when a common antenna is used for transmission and reception.

- The switching time of duplexer is critical in the operation of radar and it affects the minimum range. A reflected pulse is not received during

- the transmit pulse

- subsequent receiver recovery time

- The reflected pulses from close targets are not detected as they return before the receiver is connected to the antenna by the duplexer.

Other Forms of the Radar Equation:-

FIRST EQUATION:-

If the transmit and receive antennas are not the same and have different gains, the radar equation will

$$\frac{S}{N} = \frac{P G_t G_r \lambda^2 \sigma}{(4\pi)^3 R^4 FKTB_n}$$

where G_t is the gain of transmit antenna and G_r is the gain of receive antenna .

SECOND EQUATION:-

If the target ranges are different for transmit and receive antennas. The equation will be :

$$\frac{S}{N} = \frac{P G_t G_r \lambda^2 \sigma}{(4\pi)^3 R_t^2 R_r^2 F K T B_n}$$

Where R_t and R_r are ranges between the target and the transmit antenna and the target and the receive antenna respectively .

THIRD EQUATION:-

The first radar equation we discussed was derived without incorporating losses of energy which accompany transmission, reception, and the processing of electromagnetic radiation. It is sufficient to incorporate all of these losses in one term and write equation as follows :-

$$\frac{S}{N} = \frac{P G^2 \lambda^2 \sigma}{(4\pi)^3 R^4 F K T B_n L}$$

Where L is the total loss term.

FOURTH EQUATION:-

If we know that the signal power equals the noise power S/N = 1 the equation will be : $R_0^4 = \dfrac{PG^2\lambda^2\sigma}{(4\pi)^3 FKTB_n L}$ .Note that all of

The terms appearing in the R_o equation, with the exception of the target cross section, are a characteristic of the radar system.

Once a design is established, R_o can be determined for a given target from the fourth size. Using the value of R_o equation in the third equation we get $\dfrac{S}{N} = \left[\dfrac{R_o}{R}\right]^4$. From this equation, it is noted that S/N is inversely proportional to the fourth power of Range, R.

Parameters Affecting the Radar Range Equation:-

The radar equation was derived in the previous section and is below for reference:-

$$\frac{S}{N} = \frac{PG^2 \lambda^2 \sigma}{(4\pi)^3 R^4 FKTB_n L} = \left[\frac{R_o}{R}\right]^4$$

The terms of this equation, which depend on the:-

1) Physical structure of antenna.

2) Radar transmitter.

3) Processing of received signal.

4) System losses.

5) Characteristics of the target.

Type of Transmission:-

- Passive: - there is no transmission.
- Active: - there is transmission.

RADAR PARAMETERS AND DEFINITIONS:

RADAR: Radio means Radio Detection and Ranging. It is a device useful for detecting and ranging, tracking and searching. It is useful for remote sensing, weather forecasting, speed traping, fire control and astronomical abbrivations.

Echo: Echo is a reflected electromagnetic wave from a target and it is received by radar receiver. The echo signal power is captured by the effective area of the receiving space antenna.

Duplexer: It is a microwave switch which connects the transmitter and receiver to the antenna alternatively. It protects the receiver from high power output of the transmitter. It allows the use of the single antenna for both radar transmistion and reception. It balnks the receiver during the transmitting period.

Antenna: It is a device which acts as atransducer between transmitter and free space and between free space and receiver. It converts electromagnetic energy into electrical energy at receiving side and converts the electrical energy into electromagnetic energy at the transmitting side. Antenna is a source and a sensor of electromagnetic waves. It is also an impedence matching device and a radiator of electromagnetic waves.

Transmitter: It conditions the signals interest and connects them to the antenna. The transmitter generates high power RF energy. It consists of magnetron or klystron or travelling wave tube or cross field amplifier.

Receiver: It receives the signals from the receiving antenna and connects them to display. The receiver amplifies weak return pulses and separates noise and clutter.

Synchronizer: It synchronizes and coordinates the timing for range determination. It regulates PRF and resets for each pulse. Synchronizer connects the signals simultaneously to transmitter and display. It maintains timing of transmitted pulses. It ensures that all components and devices operate in a fixed time relationship.

Display: It isa device to present the received information for the operator to interpret. It provides visual presentation of echoes.

Bearing or Azimuth Angle: It is an angle measured from true north in a horizontal plane. In other words, it is the antenna beams angle on the local horizontal plane from some reference. The reference is usually true north.

Elevation Angle: It is an angle measured between the horizontal plane and line of sight. In other words, it is an angle between the radar beam antenna axis and the local horizontal.

Resolution: It is the ability to separate and detect multiple targets or multiple features on the same target. In other words, it is the ability of radar to distinguish targets that are very close in either range or bearing. The targets can be resolved in four dimensions range, horizontal cross- range, vertical cross-range and Doppler shift.

Range Resolution (RS): It is the ability of Radar to distinguish two or more targets at different ranges but at the same bearing. It has the units of distance.

$$RS = v_o \times (PM/2) \text{ in meters}$$

Bearing Resolution: It is the ability of Radar to distinguish objects which are in different bearing but at the same range. It is expressed in degrees.

Range of Radar: It is the distance of object from the location of radar, $R = v_o \Delta t/2$

Where, v_o = velocity of EM wave, Δt = The time taken to receiver echo from the object.

Radar Pulse: It is a modulated radiated frequency carrier wave. The carrier frequency is the transmitter oscillator frequency and it influences antenna size and beam width.

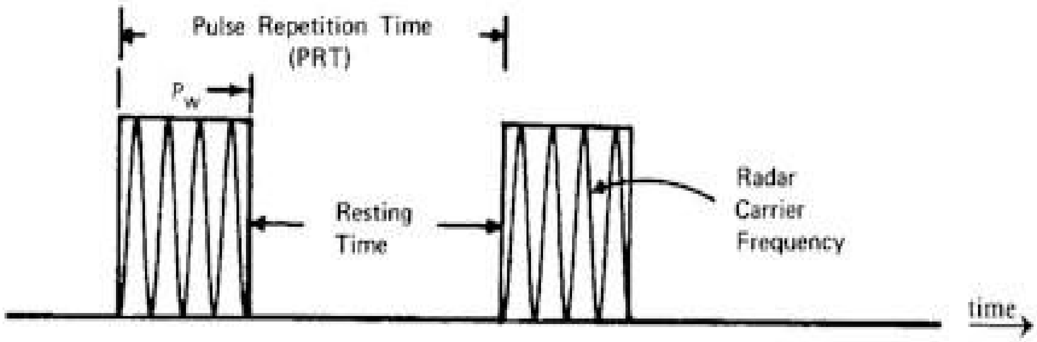


Figure 2–1. Pulse transmission.

Cross-Range Resolution of Radar: It is the ability of Radar to distinguish multiple targets at the same range. It has linear dimension perpendicular to the axis of the Radar antenna. It is of two types:

- Azimuth (Horizontal) cross-range
- Elevation (Vertical) cross-range

Narrow beam of radar antennas resolve closed spaced targets. The cross-range resolution Δx is given by, $\Delta x = R\lambda/L_{eff}$

Where R = Target range in meters

L_{eff} = Effective length of the antenna in the direction of the beam width is estimated.

λ = Wavelength in meters

Doppler Resolution: It is the ability to distinguish targets at the same range, but moving at

Different radial velocities. The Doppler resolution Δf_d is given by, $\Delta f_d = 1/T_d$ in Hz

Here T_d = The look time in seconds.

The Doppler resolution is possible if Doppler frequencies differ by at least one cycle over the time of observation. It depends on the time over which signal is gathered for processing.

Radar Signal: Radar signal is an alternating electrical quantity which conveys information. It can be voltage or current. The different types of radar signals are:

- Echoes from desired targets
- Echoes from undesired targets
- Noise signals in the receiver
- Jamming signals
- Signals from hostile sources

Radar Beam: It is the main beam of radar antenna. It represents the variation of a field

strength or radiated power as a function of θ in free space.

ECM: ECM represents Electronic Counter Measure. It is also known as jamming. It is an electronic technique which distrup radar or communication.

Radar Beam Width: It is the width of the main beam of radar antenna between two half power points or between two first nulls. It is expressed in degrees.

Search Radar: These are used for searching the targets and they scan the beam a few times per minute. These are used to detect targets and find their range, angular velocity and some times velocity. The different types of search radars are:

- Surface Search Radar
- Air Search Radar
- Two-dimentional Search Radar
- Three-dimentional Search Radar

Pulse Width: It is the duration of the radar pulse. It is expressed in milli seconds. The pulse width influences the total pulse energy. It determines minimum range and range resolution. In, fact it represents the transmitter 'ON' time.

Pulse Interval Time or Pulse Reset Time (PRT): It is the time interval between two pulses. It is expressed in milli seconds.

Pulse Reset Time (PRT) = Pulse Repetetion Time (PRT) – Pulse Width (PW)

Pulse Repetetion Frequency (PRF): It is the number of pulses transmitted per second. It is equal to the reciprocal of pulse repetition time. It is measured in hertzs.

PRF = 1/PRT

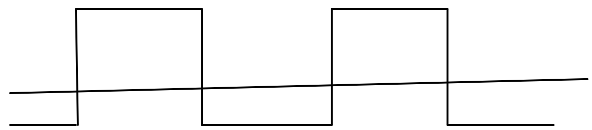
Pulse Repetetion Time (PRT): It is the time interval between the start of one pulse and the start of next pulse. It is the sum of pulse width and pulse reset time (PRT). In other words it is the time. It is measured in microseconds.

PRT = PW+PRT

Duty Cycle (D_c): It is the ratio of average power to the peak power. It is also defined as the produt of pulse width and PRF. It has no units.

$$\text{Duty Cycle, } D_c = PW \times PRF = PW/PRT = P_{avg}/P_{peak}$$

Average Power (P_{avg}): It is the average transmitted power over the pulse repetition period.



Two-Dimentional Radars: These are the radars which determine:

- Range
- Bearing of targets

Three-Dimentional Radars: These are the radars which determine:

- Altitude
- Range
- Bearing of object

Target resolution of Radar: It is the ability of Radar to distinguish targets that are very close in either range or bearing.

Navigational Radars: They are similar to search radars. They basically transmit short waves which can be reflected from earth, stones and other obstacles. These are either ship borne or airborne.

Weather Radars: These are similar to search radars. They radiate EM waves with circular polarization or horizontal or vertical polarization.

Radar Altimeter: It is radar which is used to determine the height of the aircraft from the ground.

Air Traffic Control Radars: This consists of primary and secondary radars to control the traffic in air.

Primary Radars: It is radar which receives all types of echoes including clouds and aircrafts. It receives its own signals as echoes.

Secondary Radars: It transmits the pulses and receives digital data coming from aircraft transponder. The data like altitude, call signs interms of codes are transmitted by the transponders. In military applications, these transponders are used to establish flight identity etc. Example of secondary radar is IFF radar.

Pulsed Radar: It is radar which transmits high power and frequency pulse. After transmitting one pulse, it receives echoes and then transmits another pulse. It determines direction, distance and altitude of an object.

CW Radar: It is radar which transmits high frequency signal continuously. The echo is a received and processed.

Un modulated CW Radar: It is radar in which the transmitted signal has constant amplitude and frequency. It useful to measure velocity of the object but not the speed.

Modulated CW Radar: It is radar in which the transmitted signal has constant amplitude with modulated frequency.

MTI Radar: It is pulsed radar which uses the Doppler frequency shift for discriminating moving targets from fixed ones, appearing as clutter.

Local Oscillator: It is an oscillator which generates a frequency signal which is used to convert the received signal frequency into a fixed intermediate frequency.

Mixer: It is a unit which mixes or heterodynes the frequency of the received echo signal and the frequency of local oscillator and then produces a signal of fixed frequency known as intermediate frequency. This unit is useful to increase the signal-to-noise ratio.

Doppler Frequency: Is the change in the frequency of a signal that occurs when the source and the observer are in relative motion, or when the signal is reflected by a moving object, there is an increase in frequency as the source and the observer (or the reflecting object) approach, and a decrease in frequency as they separate

Doppler Effect: Doppler Effect is discovered by Doppler. It is a shift in frequency and the wavelength of the wave as perceived by the source when the source or the target is in motion.

Astronomy Radar: It is radar which is used to probe the celestial objects.

OTH Radar: It represents Over-The-Horizon radar. It is radar which can look beyond the radio horizon. It uses ground wave and sky wave propagation modes between 2MHz and 30MHz.

MST Radar: It represents Mesosphere, Stratosphere and Troposphere radar. Mesosphere exists between 50km and 100km above the earth. Stratosphere exits between 10km and 50 km above the earth. Troposphere exists between 0 and 10km above the earth. MST Radar is used to observe wind velocity, turbulence etc.

PPI: It represents Plan position Indicator. It is a circular display with an intensity modulated map. It gives the location of a target in polar coordinates.

A-Scope: It is a radar display and represents an oscilloscope. Its horizontal coordinate represents the range and its vertical coordinate represents the target echo amplitude. It is the most popular radar display.

B-Scope: It is a radar display and it is an intensity modulated radar display. Its horizontal axis represents azimuth angle and its vertical axis represents the range of the target. The lower edge of the display represents the radar location.

Tracking Radar: It is radar which tracks the target and it is usually ground borne. It provides range tracking and angle tracking. It follows the motion of a target in azimuth and elevation.

Monostatic Radar: It is radar which contains transmitter and receiver at the same location with common antenna.

Bistatic Radar: In this radar transmitting and receiving antennas are located at different locations. The receiver receives the signals both from the transmitter and the target.

Laser Radar: It is radar which uses laser beam instead of microwave beam. Its frequency of operation is in between 30 THz and 300 THz.

Remote Sensing Radar: It provides the data about the remote places and uses the shaped beam antenna. The angle subtended at the radar antenna is much smaller than the angular width of the antenna beam.

Phased Array Radar: It is radar which uses phased array antenna in which the beam is scanned by changing the phase distribution of array. It is possible to scan the beam with this radar at a fraction of microseconds.

Clutter: The clutter is an unwanted echo from the objects other than the targets.

LIDAR: It represents Light Detection and Ranging. it is sometimes called as LADAR or Laser Radar.

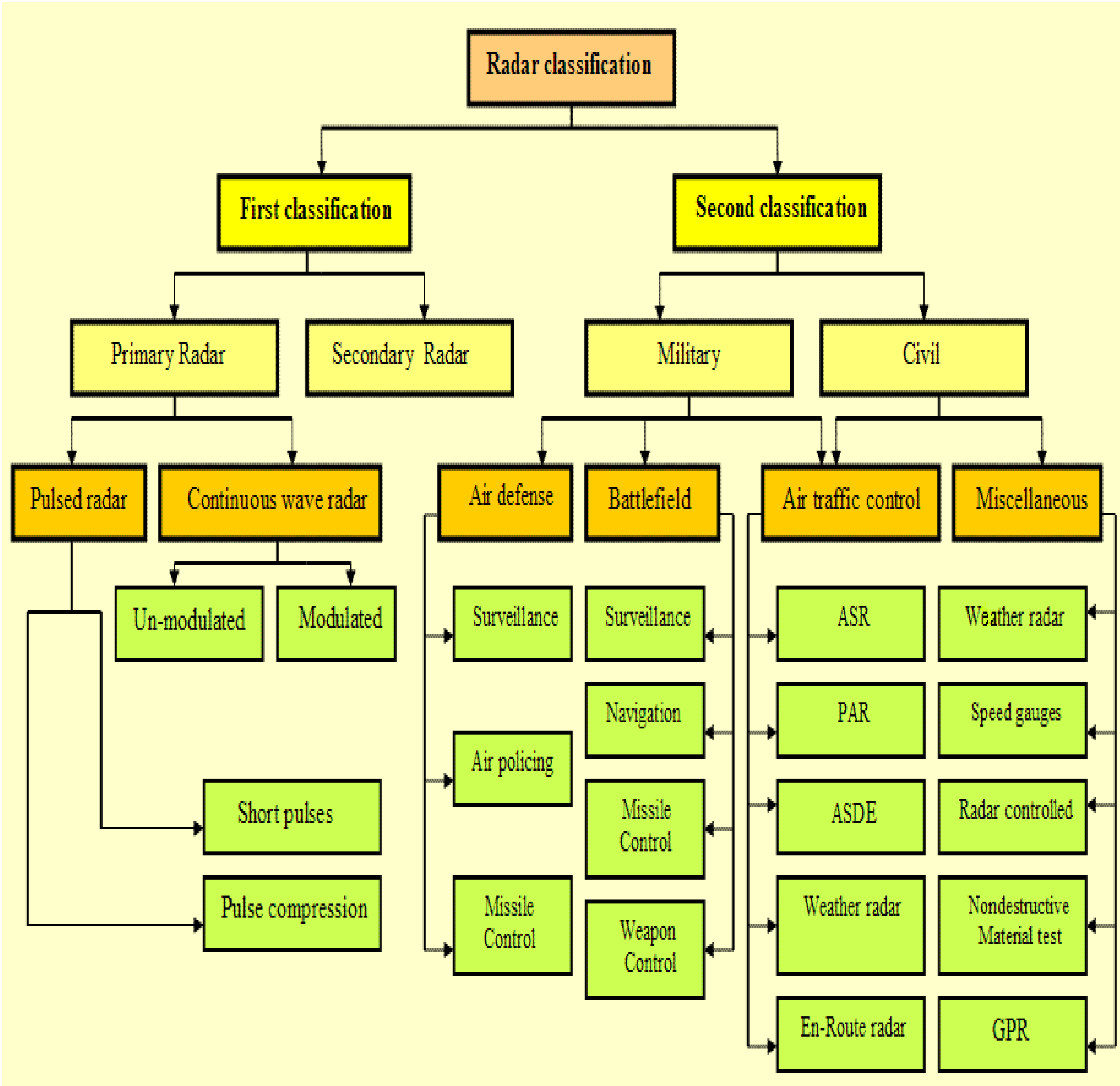
Pulse Doppler radar: It is radar that uses series of pulses to obtain velocity content.

Radar Signature: It is the identification of patterns in a target radar cross-section.

Range Tracking Radar: It is radar which tracks the targets in range.

TWS Radar: It represents Tract-While-Scan Radar. This radar scans and tracks the targets simultaneously.

Blind Range: is a range corresponding to the time delay of an integral multiple of the inter pulse period plus a time less than or equal to the transmitted pulse length. Radar usually cannot detect targets at a blind range because of interference by subsequent transmitted pulses. The problem of blind ranges can be solved or largely mitigated by employing multiple PRFs.



Radar Display:

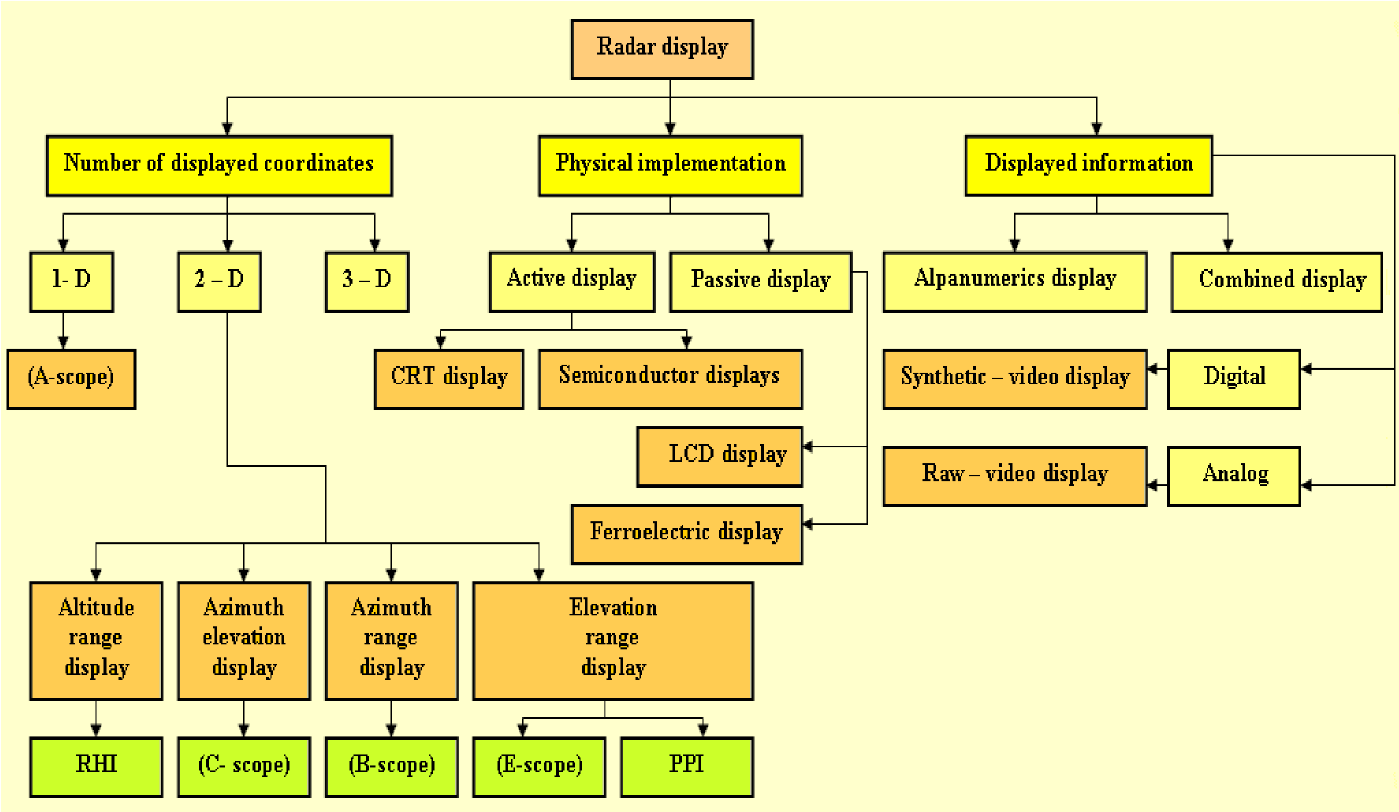
A radar display is an electronic instrument for visual representation of radar data. Radar displays can be classified from the standpoint of their functions, the physical principles of their implementation, type of information displayed, and so forth. From the viewpoint of function, they can be detection displays, measurement displays, or special displays. From the viewpoint of number of displayed coordinates, they can be one dimensional (1D), two dimensional (2D), or three dimensional (3D).

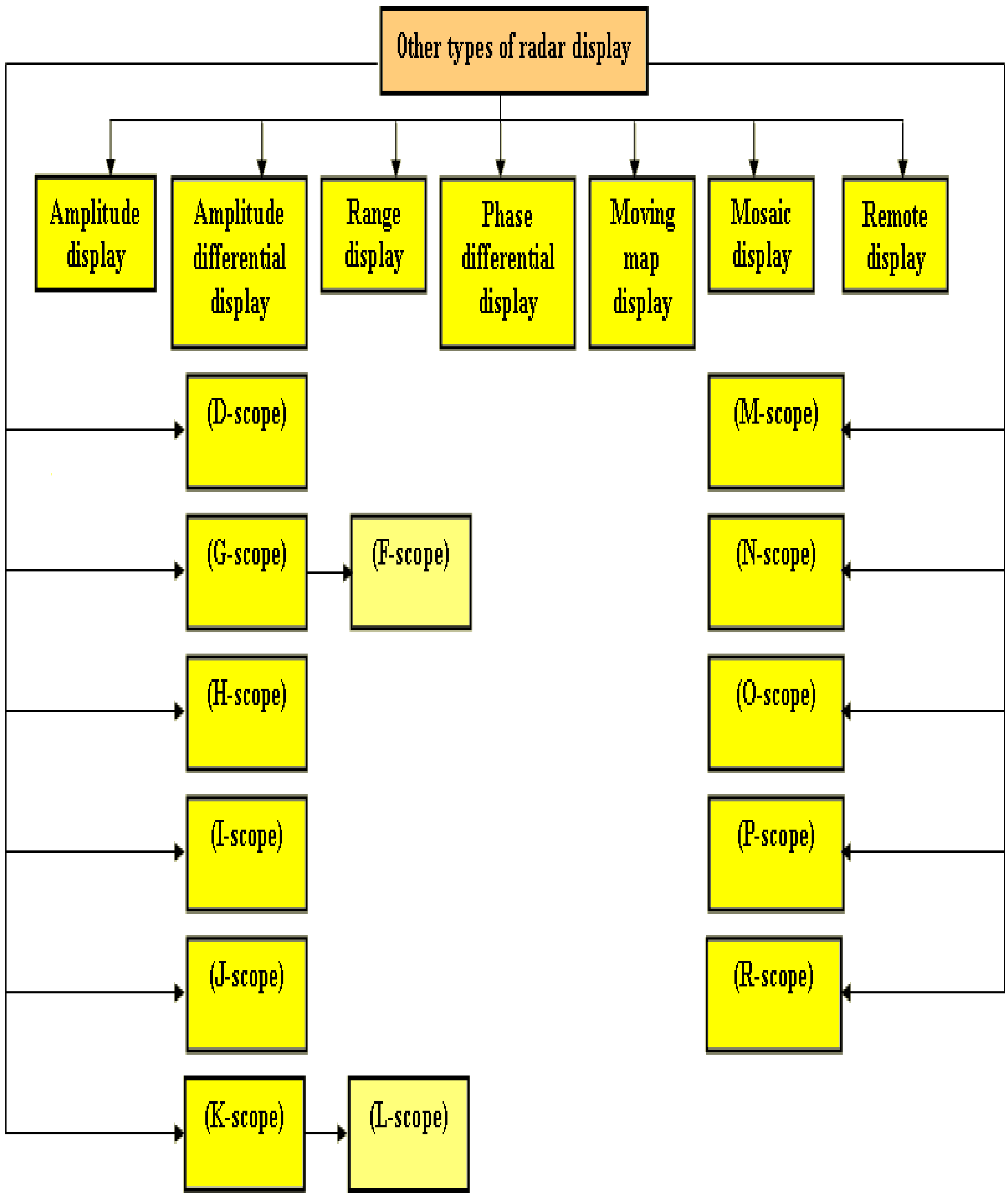
An example of a 1D display is the range display (A-scope). Most widely used are 2D displays, represented by the altitude range display (range-height indicator, or RHI), azimuth elevation display (C-scope), azimuth range display (B-scope), elevation range display (E-scope), and plan position indicator (PPI). These letter descriptions date back to World War II, and many of them are obsolete. From the viewpoint of physical implementation, active and passive displays are distinguished. The former are represented mainly by cathode ray tube (CRT) displays and semiconductor displays. Passive displays can be of liquid crystal or ferroelectric types. In most radar applications CRT displays remain the best choice because of their good performance and low cost.

From the viewpoint of displayed information, displays can be classified as presenting radar signal data, alpha numeric's, or combined displays. These can be driven by analog data (analog or raw video displays) or digital data (digital or synthetic video displays). Displays in modern radar are typically synthetic video combined displays, often using the monitors of computer

based work stations.

Now we will discuss the classifications of radar display from this figure.


Radar display
Number of displayed coordinates
Physical implementation
Displayed information
1- D
2 – D
3 – D
Active display
Passive display
Alpanumerics display
Combined display
(A-scope)
CRT display
Semiconductor displays
Synthetic – video display
Digital
LCD display
Raw – video display
Analog
Ferroelectric display
Altitude range display
Azimuth elevation display
Azimuth range display
Elevation range display
RHI
(C- scope)
(B-scope)
(E-scope)
PPI


Other types of radar display
Amplitude display
Amplitude differential display
Range display
Phase differential display
Moving map display
Mosaic display
Remote display
(D-scope)
(G-scope)
(F-scope)
(H-scope)
(I-scope)
(J-scope)
(K-scope)
(L-scope)
(M-scope)
(N-scope)
(O-scope)
(P-scope)
(R-scope)

OBJECTIVE TYPE QUESTIONS

1.The Doppler shift Df is given by ________ {]

a. 2Vr / k b. Vr / 2 k c. 2k / Vr d. k/ Vr

2. Magnetrons are commonly sued as radar transmitters because ________ [ ]

a. high power can be generated and transmitted to aerial directly from oscillator

b. it is easily cooled c. it is a cumbersome device d. it has least distortion.

3. A simple CW radar does not give range information because _________ [ ]

a. it uses the principle of Doppler shift

b. continuous echo cannot be associated with any specific part of the transmitted wave c. CW wave do not reflect from a target d. multi echoes distort the information

4. Increasing the pulse width in a pulse radar - ________ [ ]

a. increases resolution b. decreases resolution

c. has no effect on resolution d. increase the power gain

5. COHO in MTI radar operates ------- [ ]

a. at supply frequency b. at intermediate frequency c. pulse repetition frequency d. station frequency.

6. A high noise figure in a receiver means ________ [ ]

a. poor minimum detectable signal b. good detectable signal c. receiver bandwidth is reduced d. high power loss.

7. Which of the following will be the best scanning system for tracking after a target has been

acquired ________ [ ]

a. Conical b. Spiral c. Helical d. Nodding

40

8. A RADAR IS used for measuring the height of an aircraft is known as _________ [ ]

a. radar altimeter b. radar elevator c. radar speedometer d. radar

latitude

9. VOR stands for__________ [ ]

a. VHF omni range b. visually operated radar

c. voltage output of regulator. d. visual optical radar

10. The COHO in MTI radar operates at the__________ [ ]

a. received frequency b. pulse repetition frequency

c. transmitted frequency d. intermediate frequency.

11. Radar transmits pulsed electromagnetic energy because__________ [ ]

a. it is easy to measure the direction of the target. b. it provides a very ready measure*ment of range

c. it is very easy to identify the targets d. it is easy to measure the velocity of target

12. A scope displays ____________ [ ]

a. neither target range nor position, but only target velocity.

b. the target position, but not range c. the target position and range d. the target range but not position.

13. Which of the following is the remedy for blind speed problem__________ [ ]

a. change in Doppler frequency b. use of MTI

c. use of Monopulse d. variation of PRF.

14. Which of the following statement is incorrect? Flat topped rectangular pulses must be

transmitted in radar to ______________ [ ]

a. allow accurate range measurements b. allow a good minimum range. c. prevent frequency changes in the magnetron.

d. make the returned echoes easier to distinguish from noise.

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15. In case the cross section of a target is changing, the tracking is generally done by [ ]

a. duplex switching b. duplex scanning c. mono pulse d. cw radar

16. Which of the following is the biggest disadvantage of the CW Doppler radar ? [ ]

a. it does not give the target velocity b. it does not give the target position c. a transponder is required at the target d. it does not give the target range.

17. The sensitivity of a radar receiver is ultimately set by _______ [ ]

a. high S/N ratio b. lower limit of signal input

c. over all noise temperature d. higher figure of merit

18. A rectangular wave guide behaves like a ______ [ ]

a. band pass filter b. high pass filter

c. low pass filter d. m - derived filter

19. Non linearity in display sweep circuit results in _________ [ ]

a. accuracy in range b. deflection of focus

c. loss of time base trace. d. undamped indications

20. The function of the quartz delay line in a MTI radar is to _________ [ ]

a. help in subtracting a complete scan from the previous scan

b. match the phase of the Coho and the output oscillator. c. match the phase of the Coho and the stalo

d. delay a sweep so that the next sweep can be subtracted from it,

Answers:

1.a	2.a	3.b	4.b	5.b	6.a	7.a	8.a	9.a	10.d
11.b	12.d	13.d	14.d	15.c	16.d	17.c	18.b	19.a	20.a

ESSAY TYPE QUESTIONS

1. Discuss the parameters on which maximum detectable range of a radar system depends.

2. What are the specific bands assigned by the ITU for the radar? What the corresponding frequencies?
3. What are the different range frequencies that radar can operate and give their applications?

4. What are the basic functions of radar? In indicating the position of a target, what is the difference between azimuth and elevation?
5. Derive fundamental radar range equation governed by minimum receivable echo power s_{min}.
6. Modify the range equation for an antenna with a transmitting gain G and operating at a wavelength.

7. Draw the functional block diagram of simple pulse radar and explain the purpose and functioning of each block in it.
8. List major applications of radar in civil and military systems.
9. With the help of a suitable block diagram explain the operation of a pulse radar

10. Explain how the Radar is used to measure the range of a target?

11. Draw the block diagram of the pulse radar and explain the function of each block

12. Explain how the Radar is used to measure the direction and position of target?

13. What are the peak power and duty cycle of a radar whose average transmitter power is 200W, pulse width of 1µs and a pulse repetition frequency of 1000Hz?

14. What is the different range of frequencies that radar can operate and give their applications?
15. What are the basic functions of radar? In indicating the position of a target, what is the difference between azimuth and elevation?
16. Determine the probability of detection of the Radar for a process of threshold

17. Draw the block diagram of Basic radar and explain how it works?

18. Write the simplifier version of radar range equation and explain how this equation does not adequately describe the performance of practical radar?
19. Derive the simple form of the Radar equation.

20. Compute the maximum detectable range of a radar system specified below:

    a. Operating wavelength = 3.2 cm

    b. Peak pulse transmitted power = 500 kW. c. Minimum detectable power = 10^{-3} W
    c. Capture area of the antenna = 5 sq.m.

    d. e. Radar cross-sectional area of the targe t = 20 sq.m.

UNIT-II

RADAR

EQUATION

The Radar Range Equation:

We know that,

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

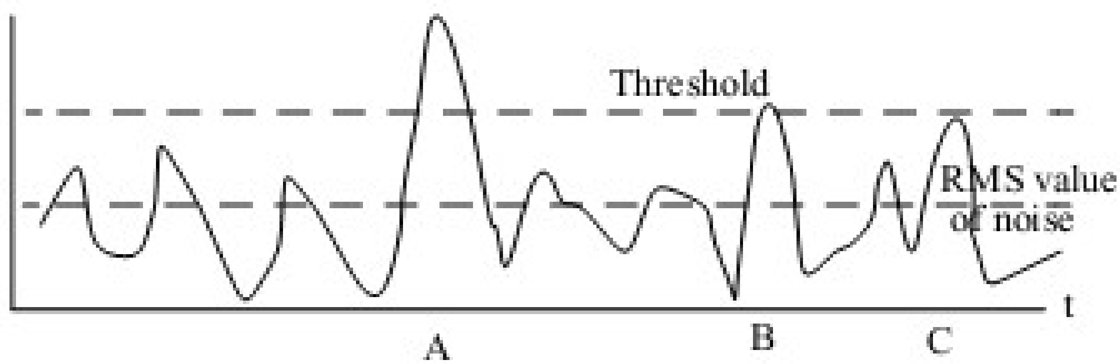
- All of the parameters are controllable by the radar designer except for the target cross section σ.
- In practice the simple range equation does not predict range performance accurately. The actual range may be only half of that predicted.
- This due, in part, to the failure to include various losses
- It is also due to the statistical nature of several parameters such as Smin, σ, and propagation losses
- Because of the statistical nature of these parameters, the range is described by the probability that the radar will detect a certain type of target at a certain distance.

Minimum detectable Signal:

- The ability of the radar receiver to detect a weak echo is limited by the noise energy that occupies the same spectrum as the signal
- Detection is based on establishing a threshold level at the output of the receiver.
- If the receiver output exceeds the threshold, a signal is assumed to be present

A sample detected envelope is show below, a large signal is detected at A. The threshold must be adjusted so that weak signals are detected, but not so low that noise peaks cross the threshold and give a false target.

The voltage envelope in the figure is usually from a matched filter receiver. A matched filter maximizes the output peak signal to average noise power level.



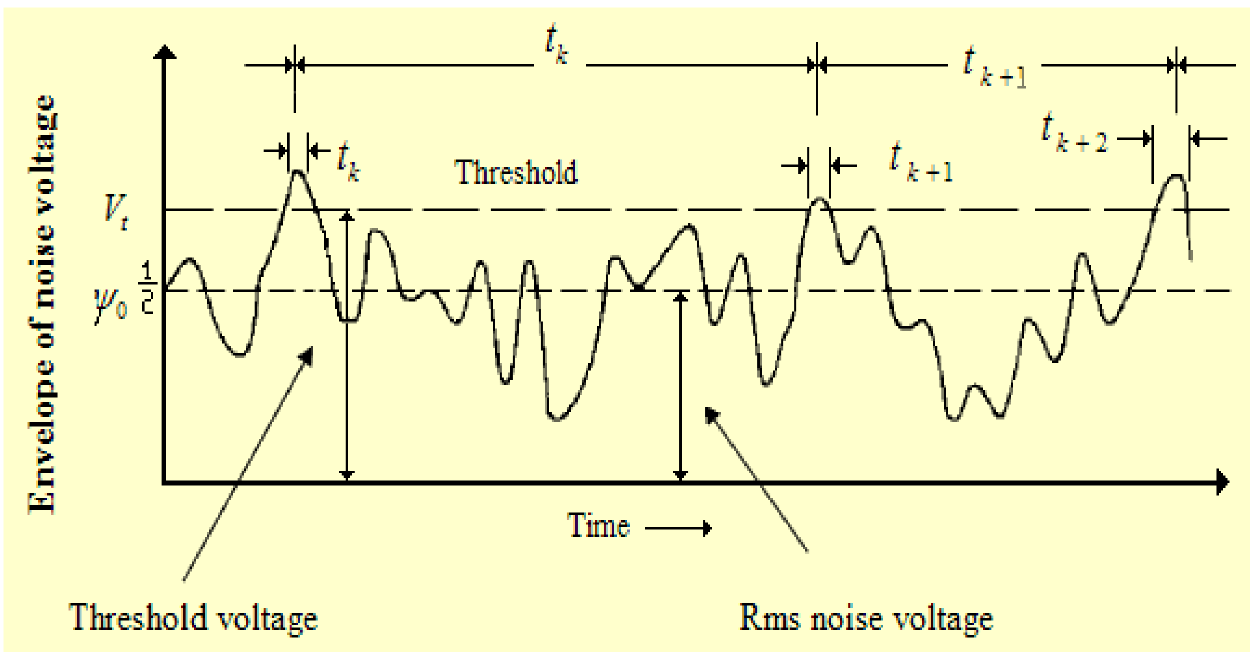


Fig: Envelope of receiver output showing false alarms due to noise.

A matched filter has a frequency response which is proportional to the complex conjugate of the signal spectrum. The output of a matched filter is the cross correlation between the received waveform and the replica of the transmitted waveform. The shape of the input waveform to the matched filter is not preserved.

In the figure, two signals are present at point B and C. The noise voltage at point B is large enough so that the combined signal and noise cross the threshold. The presence of noise sometimes enhances the detection of weak signals.

At point C the noise is not large enough and the signal is lost.

The selection of the proper threshold is a compromise which depends on how important it is if a mistake is made by (1) failing to recognize a signal (probability of a miss) or by (2) falsely indicating the presence of a signal (probability of a false alarm)

Note: threshold selection can be made byan operator viewing a CRT display. Here the threshold is difficult to predict and may not remain fixed in time.

The SNR necessary to provide adequate detection must be determined before the minimum detectable signal Smin can be computed.

Although detection decision is done at the video output, it is easier to consider maximizing the SNR at the output of the IF strip (before detection). This is because the receiver is linear up to this point.

It has been shown that maximizing SNR at the output of the IF is equivalent to maximizing the video output.

False Alarm Rate

A false alarm is „an erroneous radar target detection decision caused by noise or other interfering signals exceeding the detection threshold". In general, it is an indication of the presence of a radar target when there is no valid target. The False Alarm Rate (FAR) is calculated using the following formula:

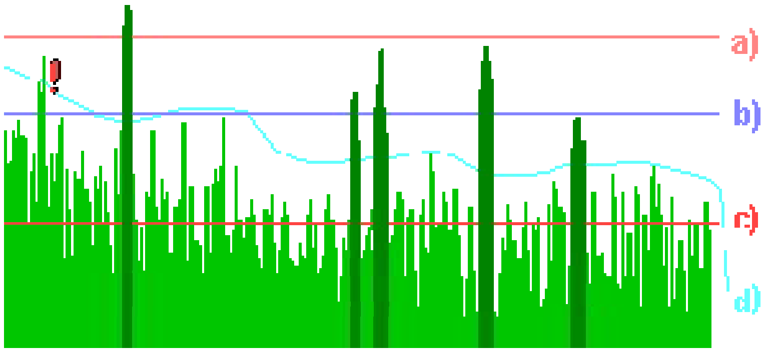


Figure 1: Different threshold levels

$$FAR = \frac{\text{False targets per PRT}}{\text{Number of range cells}} \quad \text{..........} \quad (1)$$

False alarms are generated when thermal noise exceeds a pre-set threshold level, by the presence of spurious signals (either internal to the radar receiver or from sources external to the radar), or by equipment malfunction. A false alarm may be manifested as a momentary blip on a cathode ray tube (CRT) display, a digital signal processor output, an audio signal, or by all of these means. If the detection threshold is set too high, there will be very few false alarms, but the signal-to-noise ratio required will inhibit detection of valid targets. If the threshold is set too low, the large number of false alarms will mask detection of valid targets.

- Threshold is set too high: Probability of Detection = 20%
- Threshold is set optimal: Probability of Detection = 80% But one false alarm arises!
  False alarm rate = 1 / 666 = $1{,}5 \cdot 10^{-3}$
- Threshold is set too low: a large number of false alarms arise!
- Threshold is set variabel: constant false-alarm rate

Receiver Noise:

Noise is unwanted EM energy which interferes with the abilityof the receiver to detect wanted signals. Noise may be generated in the receiver or may enter the receiver via the antenna.

One component of noise which is generated in the receiver is thermal (or Johnson) noise.

Noise power (Watts) = kTB_n

Where k = Boltzmann's constant =1.38×10^{-23} J/deg

T = degrees Kelvin and B_n = noise bandwidth

Note: B_n is not the 3 dB bandwidth but is given by:

$$B_n = \frac{\int_{-\infty}^{\infty} |H(f)|^2 df}{|H(f_0)|^2}$$

Here f_0 is the frequency of maximum response

i.e. B_n is the width of an ideal rectangular filter whose response has the same area as the filter or amplifier in question.

Note: For many types of radar Bn is approximately equal to the 3 dB bandwidth (which is easier to determine).

Note: A receiver with a reactive input (e.g. a parametric amplifier) need not have any ohmic loss and hence all thermal noise is due to the antenna and transmission line preceding the antenna.

The noise power in a practical receiver is often greater than can be accounted for by thermal noise. This additional noise is created by other mechanisms than thermal agitation.

The total noise can be considered to be equal to thermal noise power from an ideal receiver multiplied by a factor called the noise figure F_n (sometimes NF)

$$F_n = \frac{N_0}{(kT_0B_n)G_a}$$

= Noise out of a practical receiver/Noise out of an ideal receiver at T_0

Here G_a is the gain of the receiver

Note: the receiver bandwidth B_n is that of the IF amplifier in most

receivers. Since,

$$G_a = \frac{S_o}{S_i} \quad \text{and} \quad N_i = kT_0B_n$$

We have,

$$F_n = \frac{S_i/N_i}{S_0/N_0}$$

Rearranging gives:

$$S_i = \frac{kT_0B_nF_nS_o}{N_0}$$

Now S_{min} is that value of S_i corresponding to the minimum output SNR: (S_o/N_o) necessary for detection. Hence

$$S_{min} = kT_0B_nF_n\left(\frac{S_0}{N_0}\right)_{min}$$

Substituting the above equation into the radar range equation, we get,

$$R_{max}^4 = \frac{P_tGA_e\sigma}{(4\pi)^2kT_0B_nF_n(S_0/N_0)_{min}}$$

Probability Density Function (PDF):

Consider the variable x as representing a typical measured value of a random process such as a noise voltage. Divide the continuous range of values of x into small equal segments of length Δx, and count the number of times that x falls into each interval. The PDF p(x) is than defined as:

$$p(x) = \lim_{\substack{\Delta x \to 0 \\ N \to \infty}} \frac{(\text{No of values in range } \Delta x \text{ at } x)}{N}$$

Where N is the total number of values

The probability that a particular measured value lies within width dx centred at x is p(x) dx, also the probability that a value lies between x_1 and x_2 is

$$P(x_1 < x < x_2) = \int_{x_1}^{x_2} p(x)dx$$

Note: PDF is always positive by definition $\int_{-\infty}^{\infty} p(x)dx = 1$

The average value of a variable function Φ(x) of a random variable x is:

$$\langle \Phi(x) \rangle_{ave} = \int_{-\infty}^{\infty} \Phi(x)p(x)dx$$

Hence the average value or mean of x is $\langle x \rangle_{ave} = \int_{-\infty}^{\infty} xp(x)dx = m_1$

Also the mean square value is $\langle x^2 \rangle_{ave} = \int_{-\infty}^{\infty} x^2 p(x) dx = m_2$

Where, m_1 and m_2 are called the first and second moments of the random variable x.

Note: If x represents current, then m1 is the DC component and m2 multiplied by the resistance gives the mean power.

Variance is defined as,

$$\mu_2 = \sigma^2 = \langle (x - m_1)^2 \rangle_{ave} = \int_{-\infty}^{\infty} (x - m_1)^2 p(x) dx$$

$$= m_2 - m^2{}_1$$

Variance is also called the second central moment. If x represents current, μ_2 multiplied bythe resistance gives the mean power of the AC component. Standard deviation, σ is defined as the square root of the variance. This is the RMS value of the AC component.

In RADAR systems, there are different types of PDF:

- Uniform Probability Density Function
- Gaussian (Normal) Probability Density Function

- Rayleigh Probability Density Function
- Exponential Probability Density Function

Uniform Probability Density Function:

The Uniform Probability Density Function is defined as,

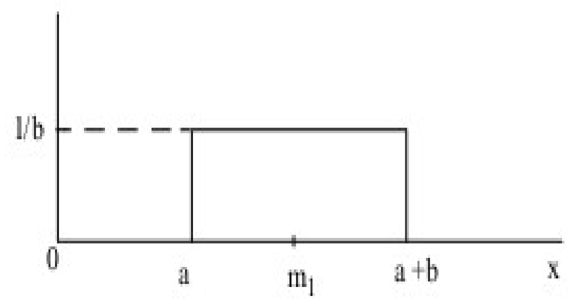
$$p(x) = \begin{cases} K, & a < x < a + b \\ 0 & x < a,\ x > a+b \end{cases}$$

Example of a uniform probability distribution is the phase of a random sine wave relative to a particular origin of time.

The constant K is found from the following $\int_{-\infty}^{\infty} p(x)dx = \int_{a}^{(a+b)} Kdx = 1 \Rightarrow K = \frac{1}{b}$

Hence for the phase of a random sine wave $K = \frac{1}{2\pi}$

The average value for a uniform PDF $m_1 = \int_a^{(a+b)} \left(\frac{1}{b}\right) x\,dx = a + \frac{b}{2}$



The mean squared value is

$$m_2 = \int_a^{(a+b)} \left(\frac{1}{b}\right) x^2\,dx = a^2 + ab + \frac{b^2}{3}$$

The variance is

$$m_2 - m_1^2 = \frac{b^2}{12}$$

The standard deviation is $\sigma = \frac{b}{2\sqrt{3}}$

Gaussian (Normal) PDF:

The Gaussian (Normal) Probability Density Function is defined as, $p(x) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{\frac{-(x-x_0)^2}{2\sigma^2}\right\}$

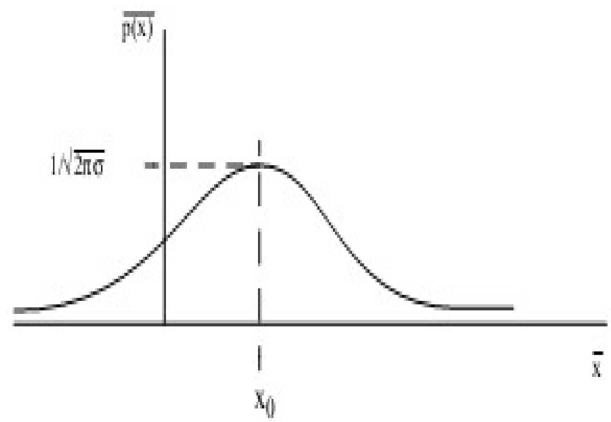
An example of normal PDF is thermal noise

We have for the Normal PDF

$$m_1 = x_0$$

$$m_2 = x^2_0 + \sigma^2$$

$$\sigma^2 = m_2 - m_1^2$$



***Central Limit Theorem*:**

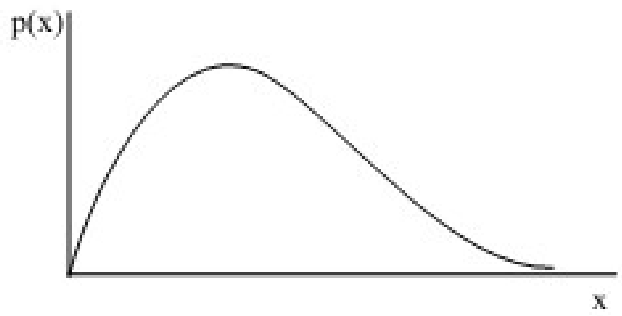
The PDF of the sum of a large number of independent, identically distributed randomquantities approaches the Normal PDF regardless of what the individual distribution might be, provided that the contribution of anyone quantityis not comparable with the resultant of all the others.

For the Normal distribution, no matter how large a value of x we may choose, there is always a finite probability of finding a greater value.

Hence if noise at the input to a threshold detector is normally distributed there is always a chance for a false alarm.

Rayleigh PDF:

$$p(x) = \frac{x}{\langle x^2 \rangle_{ave}} \exp\left(-\frac{x^2}{2\langle x^2 \rangle_{ave}}\right) \qquad x \geq 0$$



Examples of a Rayleigh PDF are the envelope of noise output from a narrowband band pass filter (IF filter in superheterodyne receiver), also the cross section fluctuations of certain

Here $\sigma = m_1\sqrt{\frac{4}{\pi} - 1}$

Exponential PDF:

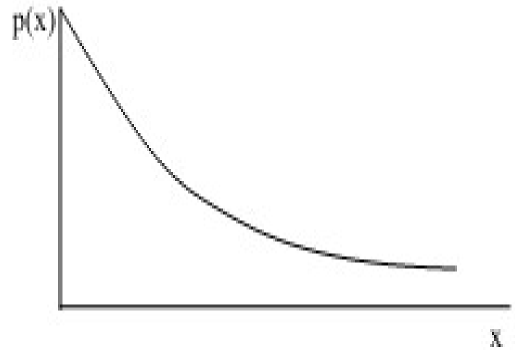
If x^2 is replaced by w where w represents power. And $\langle x^2\rangle_{avg}$ is replaced by w_0 where w_0

represents average power

Then

$p(w) = \frac{1}{w_0}\exp\left(-\frac{w}{w_0}\right)$, for $w \geq 0$

This is called the exponential PDF or the Rayleigh Power PDF



Here $\sigma = w_0$

The Probability Distribution Function is defined as, P(x) = probability (X≤x)

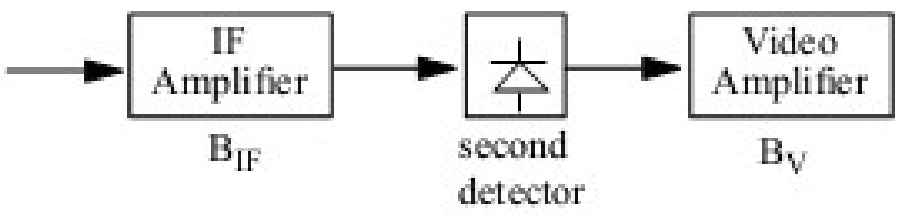
$$p(x) = \int_{-\infty}^{x} p(x)dx$$

In some cases the distribution function is easier to obtain from experiments.

Signal to Noise Ratio:

Here we will obtain the SNR at the output of the IF amplifier necessary to achieve a specific probability of detection without exceeding a specified probability of false alarm.

The output SNR is then substituted into maximu radar range equation to obtain S_{min}, the minimum detectable signal at the receiver input.



Here $B_V > B_{IF}/2$ in order to pass all video modulation.

The envelope detector may be either a square law or linear detector. The noise entering the IF amplifier is Gaussian.

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

Here ψ_0 is the variance, the mean value is zero.

When this Gaussian noise is passed through the narrow band IF strip, the PDF of the envelope of the noise is Rayleigh PDF.

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

Here R is the amplitude of the envelope of the filter output.

Now the probability that the noise voltage envelope will exceed a voltage threshold V_T (false alarm) is:

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

The average time interval between crossings of the threshold by noise alone is the false alarm time T_{fa}.

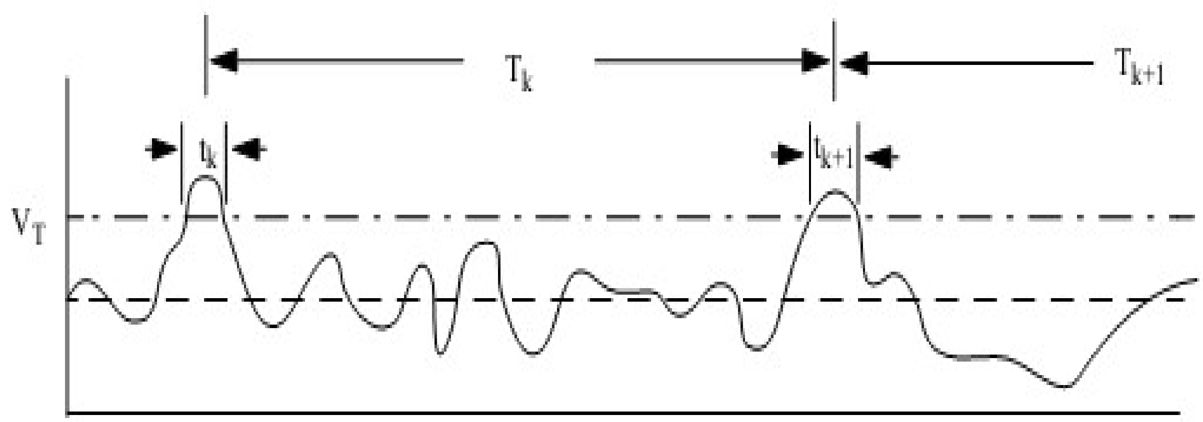
$$T_{fa} = \lim_{N \to \infty} \frac{1}{N} \sum_{k=1}^{N} T_k$$

Here T_k is the time between crossings of the threshold by noise when the slope of the crossing is Positive.

Now the false alarm probability Pfa is also given by the ratio of the time that the envelope is above the threshold to the total time.

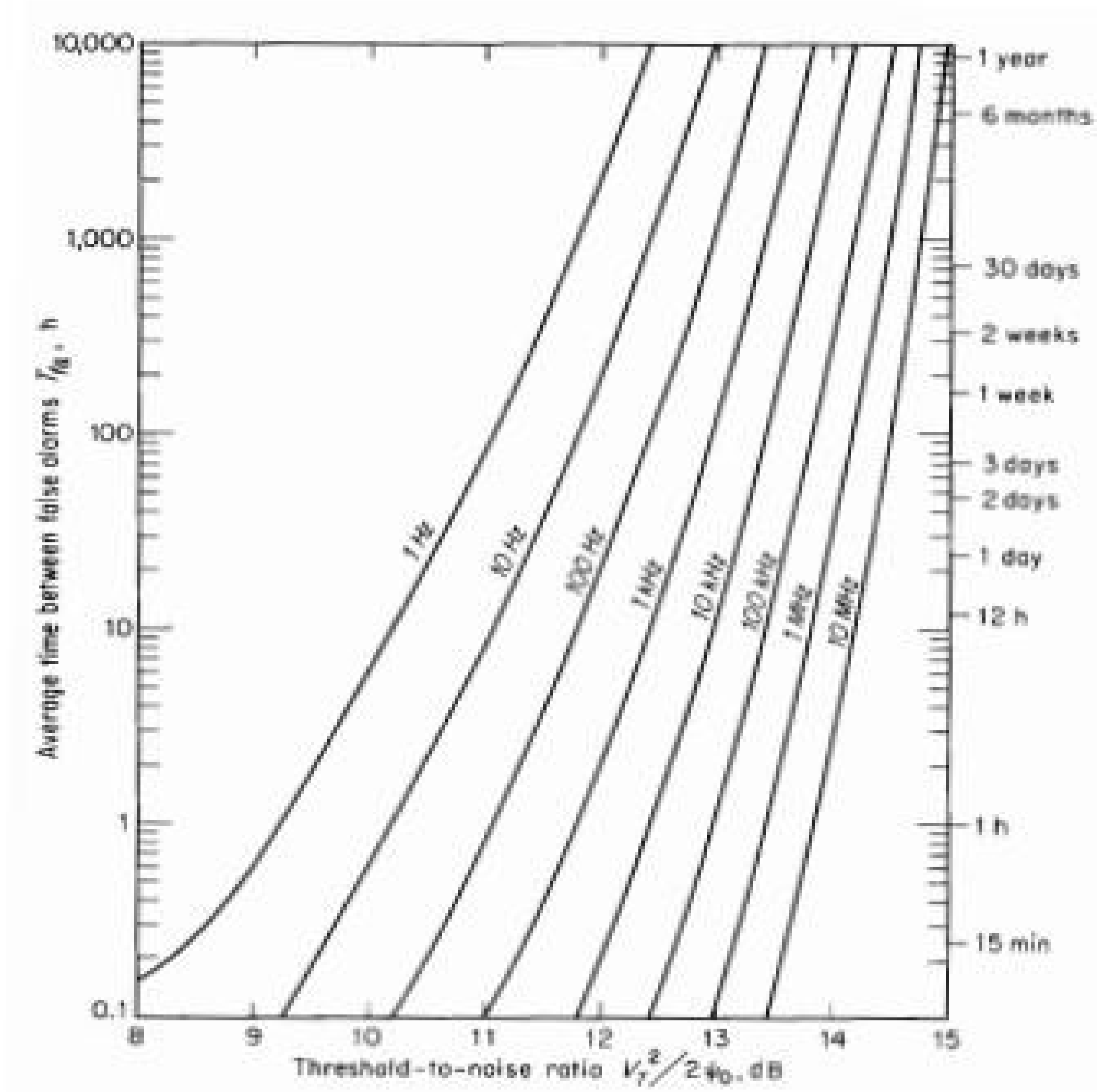
$$P_{fa} = \frac{\sum_{k=1}^{N} t_k}{\sum_{k=1}^{N} T_k} = \frac{\langle t_k \rangle_{ave}}{\langle T_k \rangle_{ave}} = \frac{1}{T_{fa} B_{IF}}$$

Where $\langle t_k \rangle \approx \dfrac{1}{B_{IF}}$



Since the average duration of a noise pulse is approximatelythe reciprocal of the bandwidth. From the above two palse alaram probabilities, the resultant equation we get,

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



Example: For B_{IF} = 1 MHz and required false alarm rate of 15 minutes.

$$P_{fa} = \frac{\sum_{k=1}^{N} t_k}{\sum_{k=1}^{N} T_k} = \frac{\langle t_k \rangle_{ave}}{\langle T_k \rangle_{ave}} = \frac{1}{T_{fa} B_{IF}}$$

$$P_{fa} = \frac{1}{(15) \cdot (60) \cdot 10^6} = 1.11x10^{-9}$$

Note: the false alarm probabilities of practical radars are quite small. This is due to their narrow bandwidth.

Note: False alarm time Tfa is very sensitive to variations in the threshold level VT due to the exponential relationship.

Example: For BIF = 1 MHz we have the following:

$V_T^2/2\psi_0$	T_{fa}
12.95 dB	6 min
14.72 dB	10,000 hours

Note: If the receiver is gated off for part of the time (e.g. during transmission interval) the P_{fa} will be increased by the fraction of the time that the receiver is not on. This assumes that T_{fa} remains constant. The effect is usually negligible.

We now consider a sine wave signal of amplitude A present along with the noise at the input to the IF strip.

Here the output of the envelope detector has a Rice PDF which is given by:

$$p(R) = \frac{R}{\psi_0}\exp\left(-\frac{R^2 + A^2}{2\psi_0}\right)I_0\left(\frac{RA}{\psi_0}\right)$$

Where $I_0(Z)$ is the modified Bessel function of zero order and argument Z

Now, $I_0(Z) \approx \frac{e^Z}{\sqrt{2\pi Z}}\left(1 + \frac{1}{8Z} + \ldots\right)$ for Z large

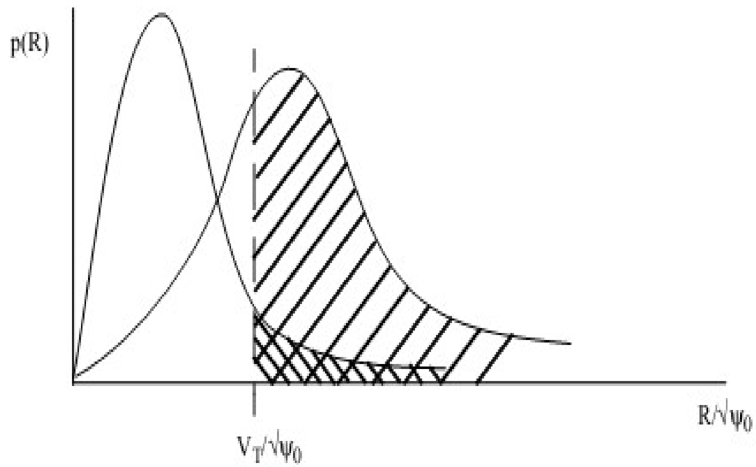
Note: when A = 0, the above equation reduces to the PDF from noise alone.

The probability of detection P_d is the probability that the envelope will exceed V_T.

$$P_d = \int_{V_T}^{\infty} p(R)dR$$

For the conditions $RA/\psi_0 >> 1$ and $A >> |R-A|$

$$P_d = \frac{1}{2}\left[1 - erf\left(\frac{V_T - A}{\sqrt{2\psi_0}}\right)\right] + \left[\exp\left(-\frac{(V_T - A)^2/(2\psi_0)}{2\sqrt{2\pi}A/(\sqrt{\psi_0})}\right)\right]\left[1 - \left(\frac{V_T}{4A} + \frac{1 + (V_T - A)^2/\psi_0}{(8A^2)/\psi_0}\right)\right]$$



Note: 1. the area represents the probability of detection.

2. The area represents the probability of false alarm. If P_{fa} is decreased by moving V_T then P_d is also decreased.

The above P_d may be converted to power by replacing the signal-r.m.s.-noise-voltatge ratio. The signal-r.m.s.-noise-voltatge ratio is given by

$\frac{A}{\sqrt{\psi_0}}$ = [Signal amplitude/RMS noise voltage] = √2[RMS signal voltage/ RMS noise voltage]

= [Signal power/noise power]$^{1/2}$ = $(2S/N)^{1/2}$

$$\frac{A}{\sqrt{\psi_0}} = \sqrt{\frac{2S}{N}} \quad \text{and}$$

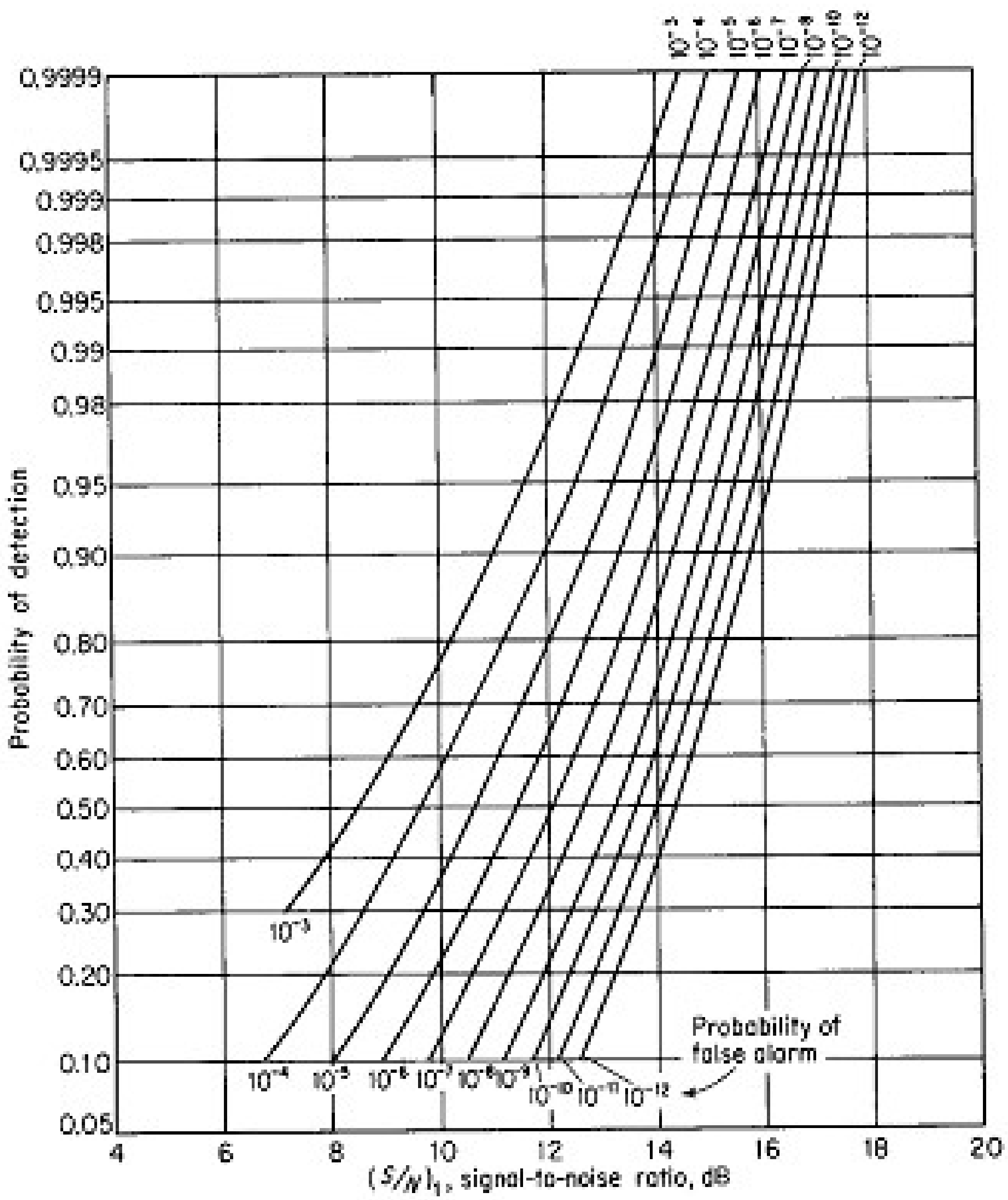
$$\frac{V_T^2}{2\psi_0} = \ln\frac{1}{P_{fa}}$$

The performance specification is Pfa and Pd and used to determine the S/N at the receiver output and the Smin at the receiver input.

Note: This S/N is for a single radar pulse.

The above figure shows the probability of detection for a sine wave in noise as a function of the signal-to-noise (power) ratio and the probability of false alaram.

Note: S/N required is high even for P_d = 0.5. This is due to the requirement for the P_{fa} to be small. A change in S/N of 3.4 dB can change the P_d from 0.999 to 0.5. When a target cross section fluctuates, the change in S/N is much greater than this S/N required for detection is not a sensitive function of false alarm time.



Integration of Radar pulses:

The above figure applies for a single pulse only. However many pulses are usually returned from any particular target and can be used to improve detection. The number of pulses n_B as the antenna scans is

$$n_B = \frac{\theta_B f_P}{\dot{\theta}_S} = \frac{\theta_B f_P}{6\omega_m}$$

Where θ_B = antenna beam width (deg) and f_P = PRF (Hz)

$\dot{\theta}_S$ = antenna scan rate (deg/sec)

ω_m = antenna scan rate (rpm)

Example: For a ground based search radar having

$\theta_B = 1.5^\circ$, $f_P = 300$ Hz, $\dot{\theta}_S = 30^\circ/\mathrm{s}$ $(\omega_m = 5\ \mathrm{rpm})$

Determine the number of hits from a point target in each scan n_B = 15

The process of summing radar echoes to improve detection is called

integration. All integration techniques employ a storage device

- The simplest integration method is the CRT displaycombined with the integrating properties of the eye and brain of the operator.
- For electronic integration, the function can be accomplished in the receiver either before the second detector (in the IF) or after the second detector (in the video).
- Integration before detection is called predetection or coherent detection.
- Integration after detection is called postdetection or noncoherent integration.
- Predetection integration requires the phase of the echo signal to be preserved.
- Postdetection integration can not preserve RF phase.
- For predetection SNR integrated = n SNRi or $(SNR)_n = n(SNR)_1$

Where SNRi is the SNR for a single pulse and n is the number of pulses integrated.

- For postdetection, the integrated SNR is less than the above since some of the energy is converted to noise in the nonlinear second detector.
- Postdetection integration, however, is easier to implement

Integration efficiency is defined as

$$E_i(n) = \frac{(S/N)_1}{n(S/N)_n} \qquad \text{---------- (1)}$$

Where $(S/N)_1$ = value of SNR of a single pulse required to produce a given probability of detection and

$(S/N)_n$ = value of SNR per pulse required to produce the same probability of detection. When n pulses are integrated.

For postdetection integration, the integration improvement factor is $I_i = n\,E_i(n)$

For ideal postdetection, $E_i(n) = 1$ and hence the integration improvement factor is n

Examples of I_i are given in Fig from data by Marcum

Note that I_i is not sensitive to either P_d or P_{fa}.

We can also develop the integration loss as $L_i = 10\log\left[\frac{1}{E_i(n)}\right]$

This is shown in Fig.

The parameter n_f in Fig. is called the false alarm number which is defined as the average number of possible decisions between false alarms

n_f = [no. of range intervals/pulse][no. of pulse periods/sec][false alarm rate]

= $[T_P/\tau][f_P][T_{fa}]$

Here T_P = PRI (pulse repetition interval) and f_P = PRF

Thus $n_f = T_{fa}/\tau = \approx T_{fa}B \approx 1/P_{fa}$

Note: for a radar with pulse width τ, there are $B = 1/\tau$ possible decisions per second on the presence of a target

If n pulses are integrated before a target decision is made, then there is B/n possible decisions/sea.

Hence the false alarm probability is n times as great.

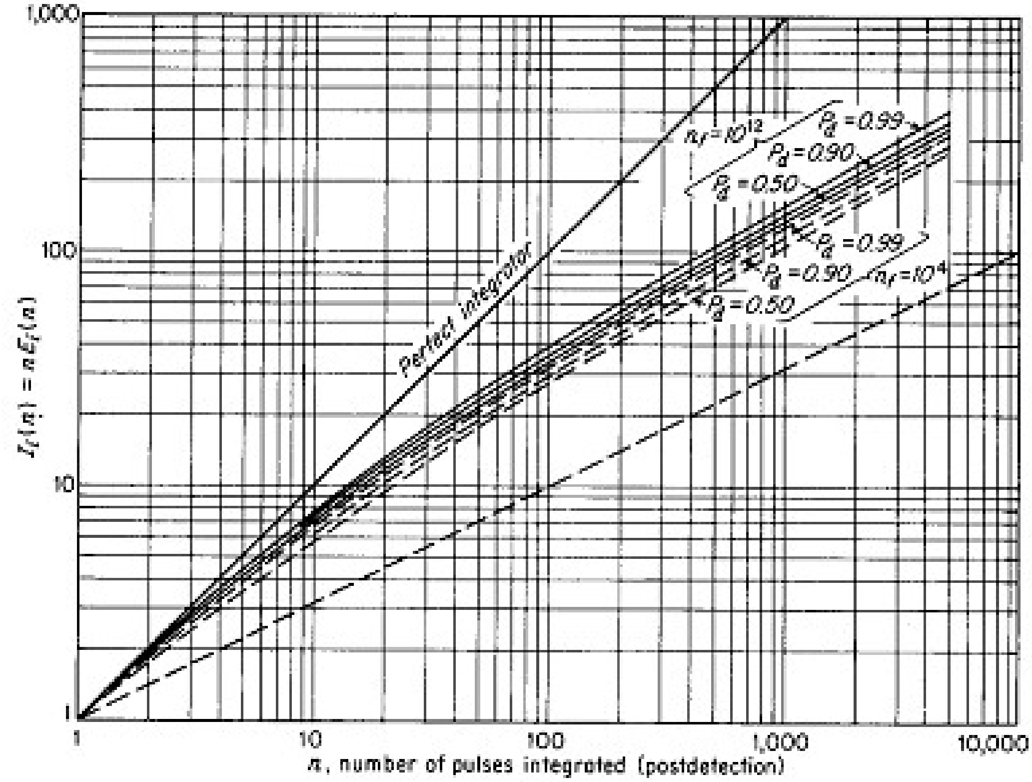
Note: This does not mean that there will be more false alarms since it is the rate of detection- decisions is reduced, not the average time between false alarms.

Hence T_{fa} is more meaningful than P_{fa}

Note: some authors use a false alarm number $n_f' = n_f/n$

Caution should be used in computations for SNR as a function of P_{fa} and P_d

Fig. shows that for a few pulses integrated post detection, there is not much difference from a perfect predetection integrator.


Perfect integrator
$n_f = 10^{12}$
$P_d = 0.99$
$P_d = 0.90$
$P_d = 0.50$
$P_d = 0.99$
$P_d = 0.90$
$P_d = 0.50$
$n_f = 10^4$
$I_i(n) = n E_i(n)$
n, number of pulses integrated (postdetection)
1,000
100
10
1
1
10
100
1,000
10,000

(a)

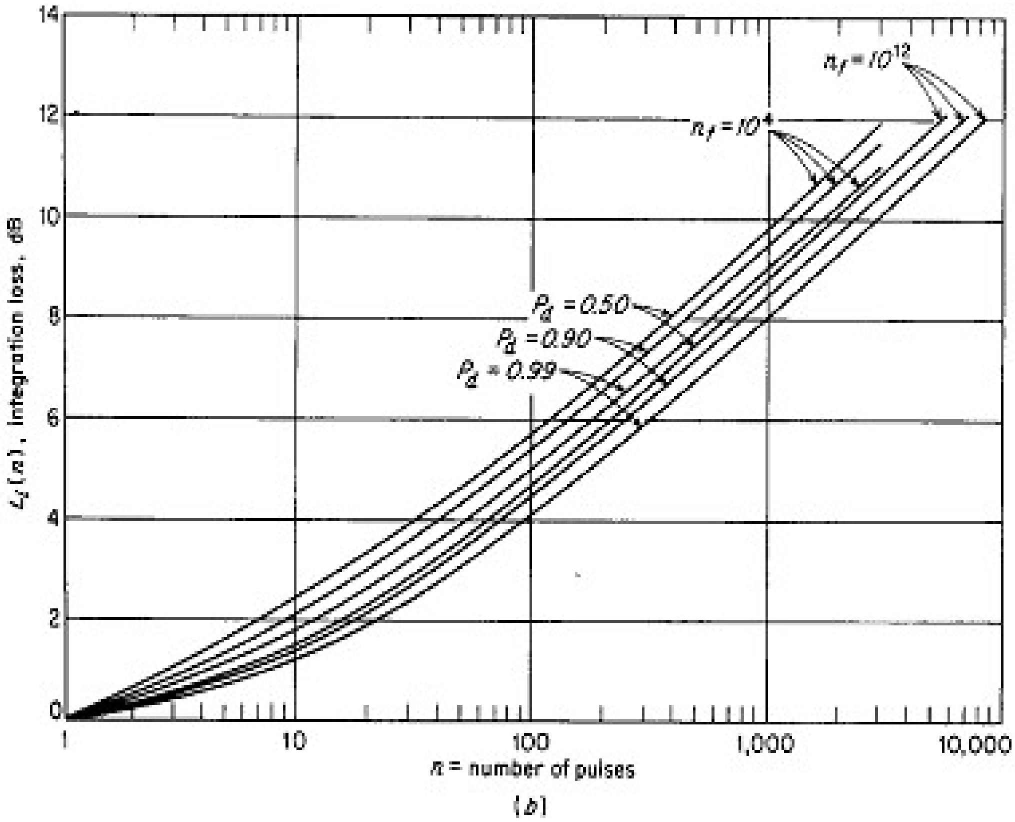


Figure 2.8 (a) Integration-improvement factor, square law detector, P_d = probability of detection, $n_f = T_{fa}B$ = false alarm number, T_{fa} = average time between false alarms, B = bandwidth; (b) integration loss as a function of n, the number of pulses integrated, P_d, and n_f. (*After Marcum,*[10] *courtesy IRE Trans.*)

When there are many pulses integrated (small S/N per pulse) the difference is pronounced. The radar equation with n pulses integrated is

$$R^4_{max} = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_n F_n (S/N)_n}$$

Here $(S/N)_n$ is the SNR of one of n equal pulses that are integrated to produce the required P_d for a specified P_{fa}.

Using equation 1 into above equation, we get,

$$R_{max}^4 = \frac{P_t G A_e \sigma n E_i(n)}{(4\pi)^2 k T_0 B_n F_n (S/N)_1}$$

Here $(S/N)_1$ is found from Fig. and $nE_i(n)$ is found from Fig .

Some postdetection integrators use a weighting of the integrated pulses. These integrators include the recirculating delay line, the LPF, the storage tube and some algorithms in digital integration.

If an "exponential" weighting of the integrated pulses is used then the voltage out of the

integrator is

$$V = \sum_{i=1}^{N} V_i \exp[-(i-1)\gamma]$$

Here V_i is the voltage amplitude of the ith pulse and $\exp(-\gamma)$ is the attenuation per pulse.

For this weighting, an efficiencyf actor ρ can be calculated which is the ratio of the average S/N

for the exponential integrator to the average S/N for the uniform integrator:

$$\rho = \frac{\tanh\left(\frac{n\gamma}{2}\right)}{n\tanh\left(\frac{\gamma}{2}\right)} \qquad \text{for a dumped integrator}$$

also

$$\rho = \frac{[1-\exp(-n\gamma)]^2}{n\tanh\left(\frac{\gamma}{2}\right)} \qquad \text{for a continuous integrator}$$

Note: Maximum efficiency for a dumped integrator corresponds to $\gamma = 0$

Maximum efficiency for a continuous integrator corresponds to $n\gamma = 1.257$

Radar Cross Section of Targets:

Cross-section: The fictional area intercepting that amount of power which, when scattered equally in all directions, produces an echo at the radar that is equal to that actually received.

$$\sigma = \frac{\text{power reflected towards the source/unit solid angle}}{\text{incident power density}/4\pi}$$

$$= \lim_{R \to \infty} 4\pi R^2 \left|\frac{E_r}{E_i}\right|^2$$

Where R = range

Er= reflected field strength at radar

Ei = incident field strength at target

Note: for most targets such as aircraft. Ships and terrain, the σ does not bear a simple relationship to the physical area.

EM scattered field: is the difference between the total field in the presence of an object and the field that would exist if the object were absent. EM diffracted field: is the total field in the presence of the object

Note: for radar backscatter, the two fields are the same (since the transmitted field has disappeared by the time the received field appears).

The σ can be calculated using Maxwell's equations onlyfor simple targets such as the sphere (Fig.2.9).

When $\frac{2\pi a}{\lambda} \ll 1$ (the Rayleigh region), the scattering from a sphere can be used for modelling raindrops. Since σ varies as λ^{-4} in the Rayleigh region, rain and clouds are invisible for long wavelength Radars.

The usual radar targets are much larger than raindrops and hence the long λ operation

does not reduce the target σ.

When $\frac{2\pi a}{\lambda} \gg 1$ the σ approaches the optical cross section πa^2

Note: in the Mie (resonance region) σ can actually be 5.6 dB greater than the optical value or 5.6

dB smaller.

Note: For a sphere the σ is not aspect sensitive as it is for all other objects, and hence can be

used fro calibrating a radar system.

Backscatter of a long thin rod (missile) is shown. Where the length is 39λ and the diameter $\lambda/4$, the material is silver.

Here $\theta = 0^\circ$ is the end on view and σ is small since the projected area is small.

However at near end on ($\theta \approx 5^\circ$) waves couple onto the rod, travel the length of the rod

and reflect from the discontinuity at the far end ⇒ large σ.

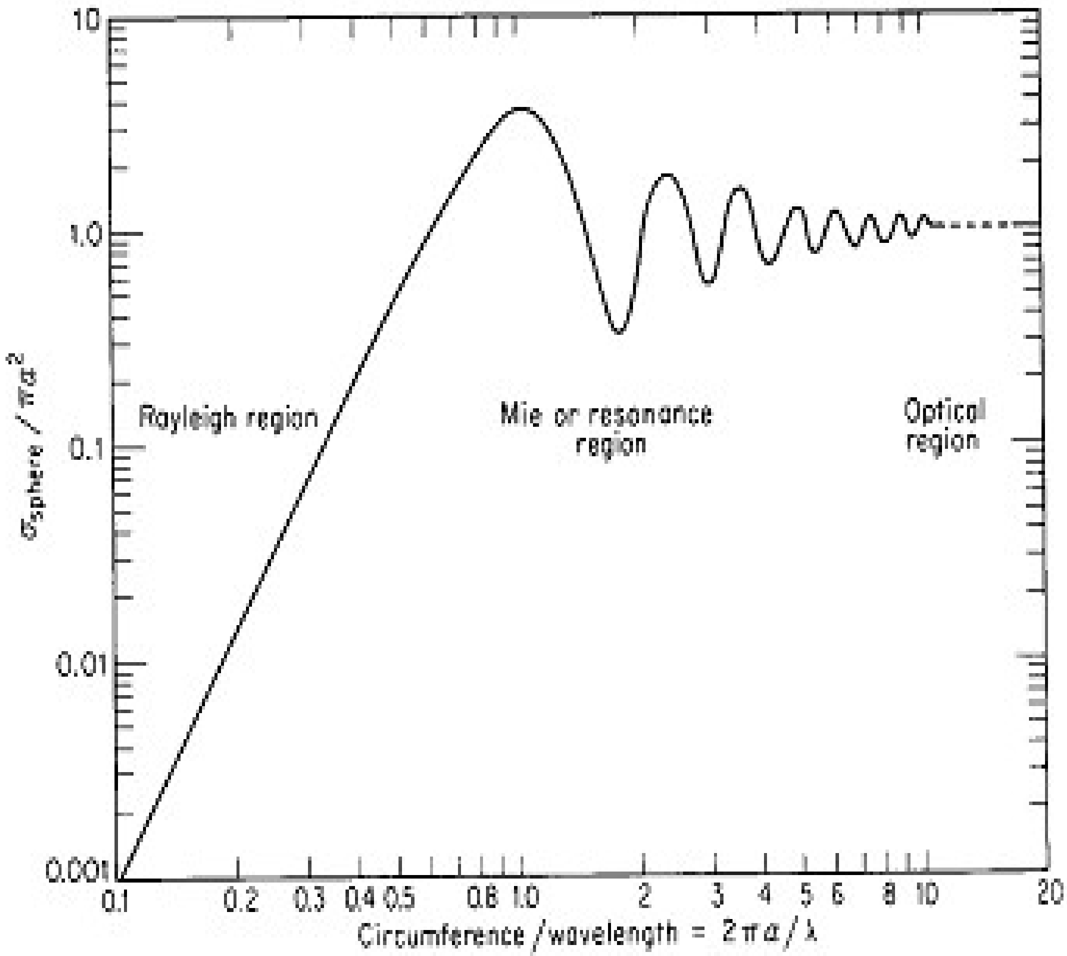


Figure 2.9 Radar cross section of the sphere. a = radius; λ = wavelength.

The Cone Sphere

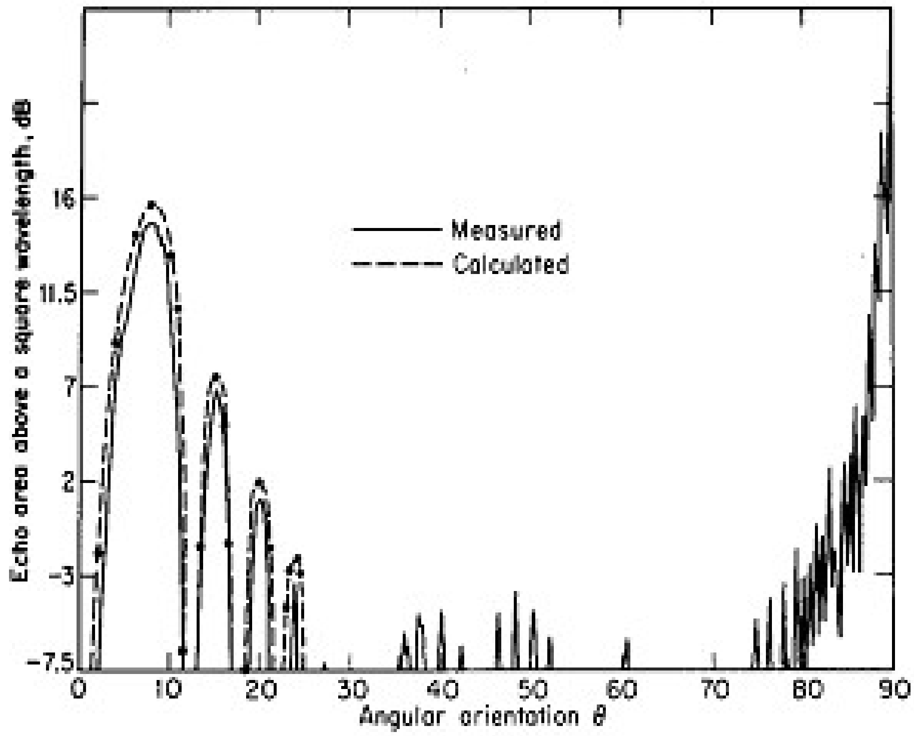


Figure 2.10 Backscatter cross section of a long thin rod. (*From Peters,*[26] *IRE Trans.*)

Here the first derivatives of the cone and sphere contours are the same at the point of joining. The nose-on σ is shown in Fig. 2.12

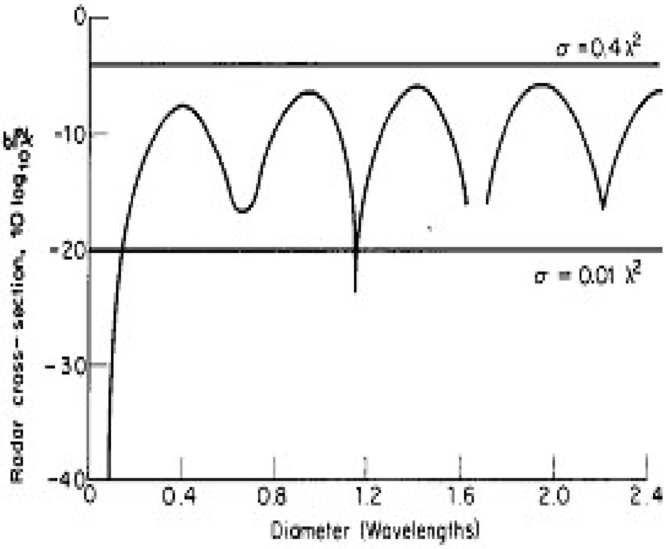


Figure 2.11 Radar cross section of a cone sphere with 15° half angle as a function of the diameter in wavelengths. (*After Blore,*[17] *IEEE Trans.*)

Note: Fig. 2.12. The σ for θ near 0˚ (-45˚ to +45˚) is quite low. This is because scattering occurs from discontinuities. Here the discontinuities are small: the tip, the join and the base of the sphere (which allows a creeping wave to travel around the sphere).

- When the cone is viewed at perpendicular incidence (θ = 90 - α, where α is the cone half angle) a large specular return is contained.
- From the rear, the σ is approximately that of a sphere.
- The nose on σ for f above the Rayleigh region and for a wide range of α, has a max of $0.4\lambda^2$ and a min of $0.01\lambda^2$. This gives a very low backscatter (e.g. at λ = 3 cm, σ = 10-4 m^2).

Example: σ at S band for 3 targets having the same projected area:

Corner reflector: 1000 m^2, Sphere 1 m^2, Cone sphere 10-3 m^2

In practice, to achieve a low σ with a cone sphere, the tip must be sharp, the surface smooth and no holes or protuberances allowed.

A comparison of nose-on σ for several cone shaped objects is given in figure 2.13

Note: the use of materials such as carbon fibre composites can further reduce σ.

Complex Targets.

The σ of complex targets (ships, aircraft, and terrain) is complicated functions of frequencyand viewing angle.

The σ can be computed using GTD (Geometric Theory of Diffraction), measured experimentally or found using scale models.

A complex target can be considered as being composed of a large number of independent objects which scatter energy in all directions.

The relative phases and amplitudes of the echo signals from the individual scatterers determine the total σ. If the separation between individual scatterers is large compared to λ the phases will vary with the viewing angle and cause a scintillating echo.

An example of the variation of σ with aspect angle is shown in Fig. 2.16. The σ can change by 15dB for an angular change of 0.33˚. Broadside gives the max σ since the projected area is bigger and is relatively flat (The B-26 fuselage had a rectangular cross-section). This data was obtained bymounting the actual aircraft on a turntable above ground and observing its σ with a radar.

A more economical method is to construct scale models. An example of a model measurement is given in Fig. 2.17 bythe dashed lines. The solid lines are the theoretical (computed using GTD) data. The computed data is obtained bybreaking up the target into simple geometrical shapes. And then computing the contributions of each (accounting for shadowing).

The most realistic method for obtaining σ data is to measure the actual target in flight. The US Naval Research Lab has such a facility with L, S, C, and X band radars. The radar track data establishes the aspect angle. Data is usually averaged over a 10˚ x 10˚ aspect angle interval.

A single value cross section is sometimes given for specific aircraft targets for use in the range equation. This is sometimes an average value or sometimes a value which is exceeded 99% of the time.

Note: even though single values are given there can be large variations in actual σ for any target e.g. the AD 4B, a propeller driven aircraft has a σ of 20 m2 at L band but its σ at VHF is about

100 m^2 This is because at VHF the dimensions of the scattering objects are comparable to λ and

produce a resonance effect.

For large ships, an average cross section taken from port, starboard and quarter aspects yields

$$\sigma_{median} = 52\sqrt{f}D^{3/2}$$ Here σ is in m^2

f is in MHz and D is ship displacement in kilotons

This equation applies only to grazing angles i.e. as seen from the same elevation.

Small boats 20 ft. to 30 ft. give σ(X band) approx 5 m^2

40 ft. to 50 ft.

" " " " " 10 m^2 Automobiles give σ(X band) of approx 10 m^2 to 200 m^2

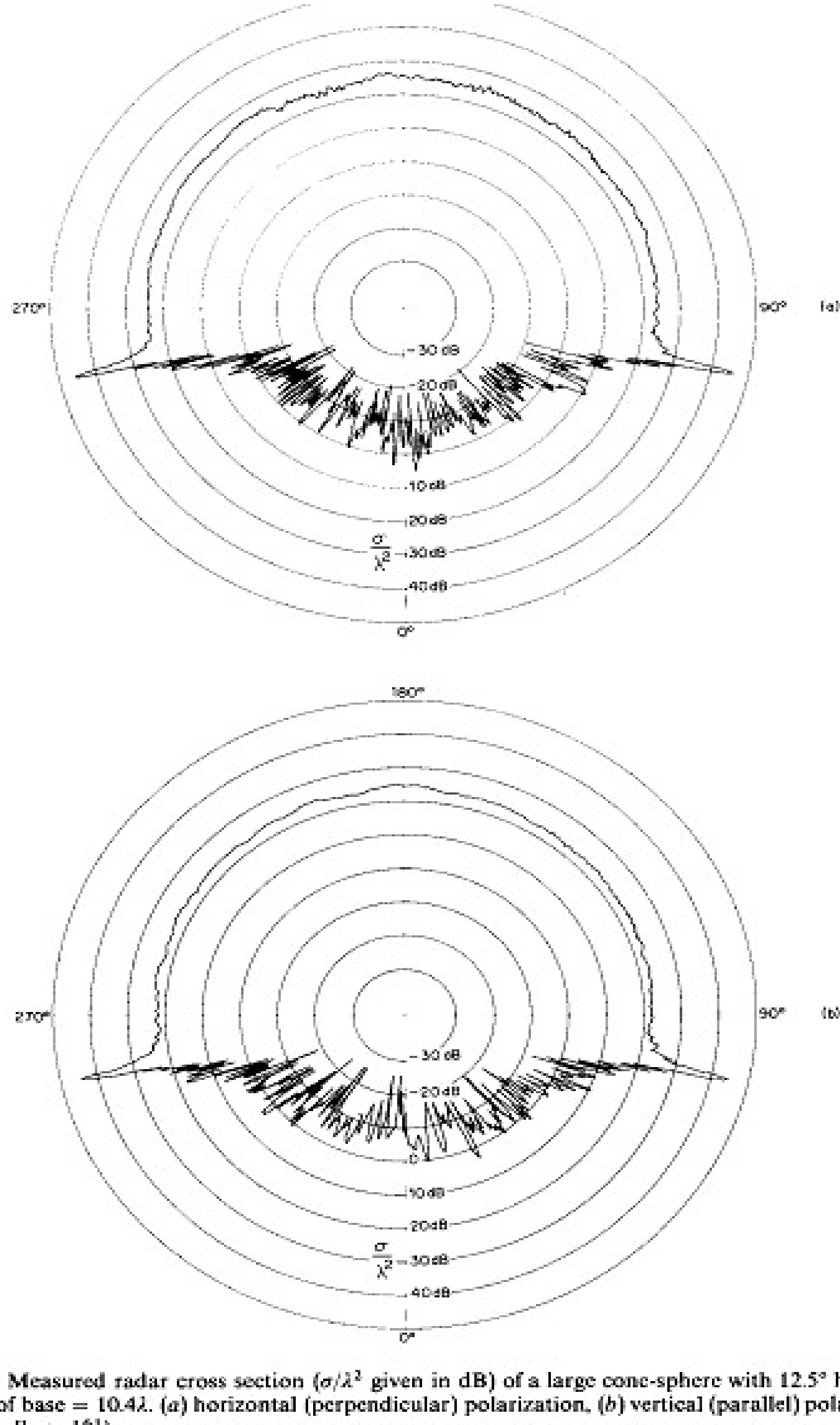


Figure 2.12 Measured radar cross section (σ/λ^2 given in dB) of a large cone-sphere with 12.5° half and radius of base = 10.4λ. (*a*) horizontal (perpendicular) polarization, (*b*) vertical (parallel) polariz (*From Pannell et al.*[61])

- a comparison of nose-on σ for several cone shaped objects is given in figure 2.13

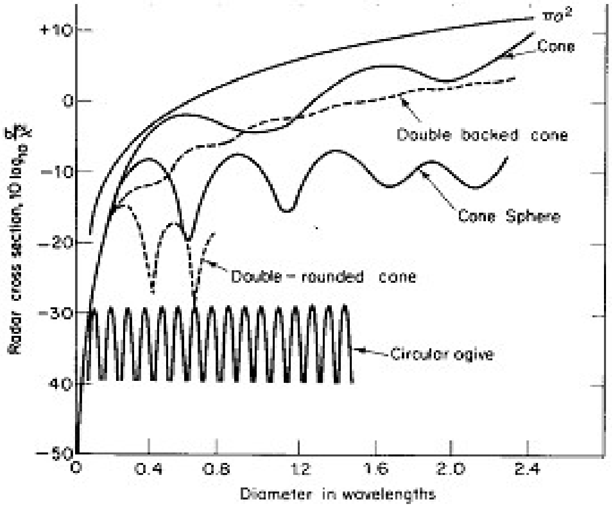


Figure 2.13 Radar cross section of a set of 40° cones, double-backed cones, cone-spheres, double-rounded cones, and circular ogives as a function of diameter in wavelengths. (*From Blore,*[27] *IEEE Trans.*)

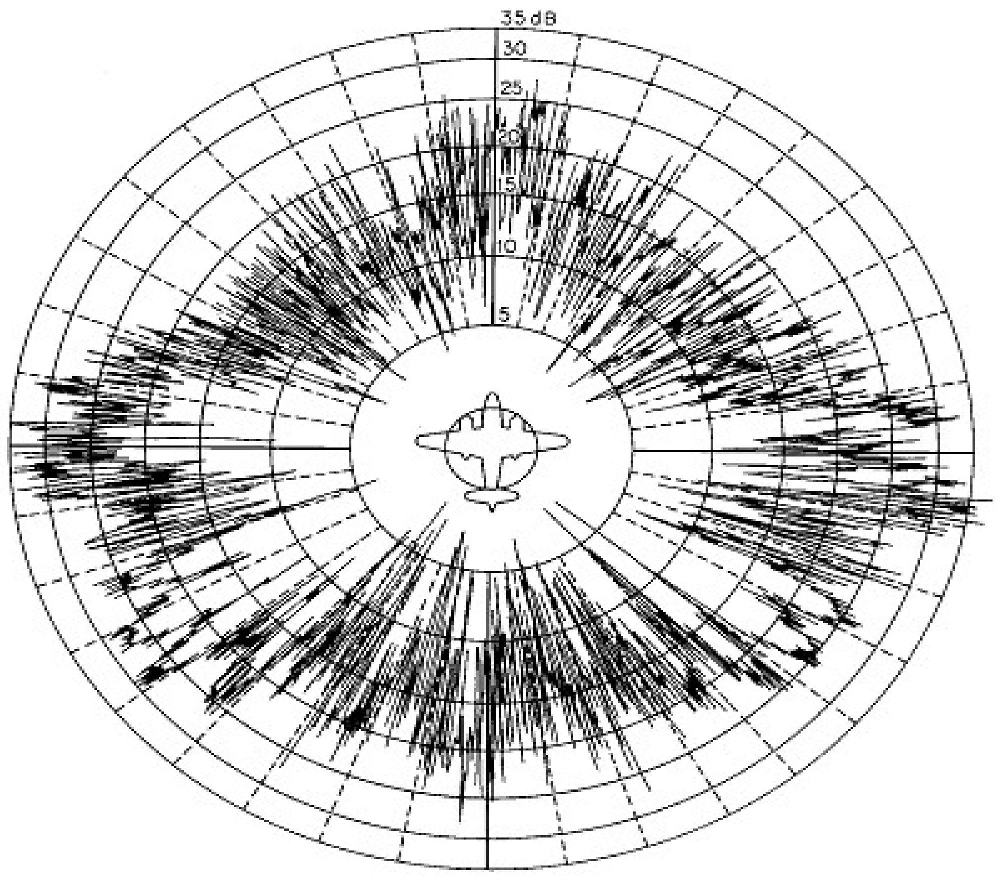


Figure 2.16 Experimental cross section of the B-26 two-engine bomber at 10-cm wavelength as a function of azimuth angle. (*From Ridenour,*[28] *courtesy McGraw-Hill Book Company, Inc.*)

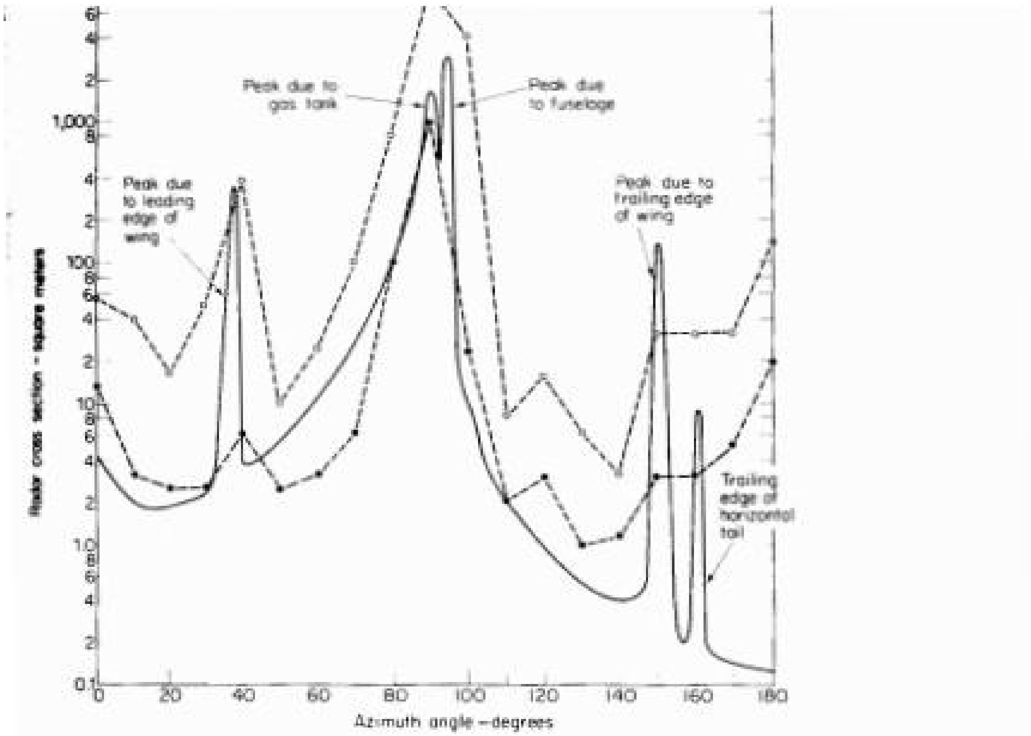


Figure 2.17 Comparison of the theoretical and model-measurement horizontal-polarization radar cross sections of the B-47 medium bomber jet aircraft with a wing span of 35 m and a length of 33 m. Solid curve is the average of the computed cross sections obtained by the University of Michigan Engineering Research Institute at a frequency of 980 MHz. Dashed curves are model measurements obtained by the Ohio State University Antenna Laboratory at a frequency of 600 MHz. Open circles are the maximum values averaged over 10° intervals; solid circles are median values. Radar is assumed to be in the same plane as the aircraft.[64]

Examples of radar cross sections for various targets (in m^2))

Conventional, unmanned winged missile	0.5
Small, single engine aircraft	1
Small fighter, or 4-passenger jet	2
Large fighter	6
Medium bomber or medium jet airliner	20
Large bomber or large jetliner	40
Jumbo jet	100
Small open boat	0.02
Small pleasure boat	2
Cabin cruiser	10
Pickup truck	200
Car	100
Bicycle	2
Man	1
Bird	.01
Insect	10^{-5}

Human being gives σ as shown:

f (MHz)	σ (m^2)
410	0.033 - 2.33
1120	0.098 - 0.997
2890	0.140 - 1.05
4800	0.368 - 1.88
9375	0.495 - 1.22

Cross-Section Fluctuations:

The echo from a target in motion is almost never constant. Variations are caused by meteorological conditions, lobe structure of the antenna, equipment instability and the variation in target cross section. Cross section of complex targets is sensitive to aspect.

One method of dealing with this is to select a lower bound of σ that is exceeded some specified fraction of the time (0.95 or 0.99). This procedure results in conservative prediction of range.

Alternatively, the PDF and the correlation properties with time may be used for a particular tar get and type of trajectory. The PDF gives the probability of finding any value of σ between the values of σ and σ + dσ. The correlation function gives the degree of correlation of σ with time (i.e. number of pulses).

The power spectral density of σ is also important in tracking radars. It is not usually practical to obtain experimental data for these functions. It is more economical to assess the effects of fluctuating σ is to postulate a reasonable model for the fluctuations and to analyze it mathematically.

Swerling has done this for the detection probabilities of 5 types of target.

Case 1

- Echo pulses received from the target on any one scan are of constant envelope throughout the entire scan, but are independent (uncorrelated) scan to scan.
- This case ignores the effect of antenna beam shape the assumed PDF is:

$$p(\sigma) = \frac{1}{\sigma_{ave}} \exp\left(\frac{-\sigma}{\sigma_{ave}}\right) \qquad \sigma \geq 0$$

Case 2

- Echo pulses are independent from pulse to pulse instead of from scan to scan

$$p(\sigma) = \frac{1}{\sigma_{ave}} \exp\left(\frac{-\sigma}{\sigma_{ave}}\right)$$

Case 3

- Same as case 1 except that the PDF is

$$p(\sigma) = \frac{4\sigma}{\sigma_{ave}^2} \exp\left(\frac{-2\sigma}{\sigma_{ave}}\right)$$

Case 4

- Same as case 2 except that the PDF is

$$p(\sigma) = \frac{4\sigma}{\sigma_{ave}^2} \exp\left(\frac{-2\sigma}{\sigma_{ave}}\right)$$

Case 5

Nonfluctuating cross section

The PDF assumed in cases 1 and 2 applies to complex targets consisting of many scatterers (in practice 4 or more). The PDF assumed in cases 3 and 4 applies to targets represented byone large reflector with other small reflectors.

For all cases the value of σ to be substituted in the radar equation is $\sigma_{avg.}$

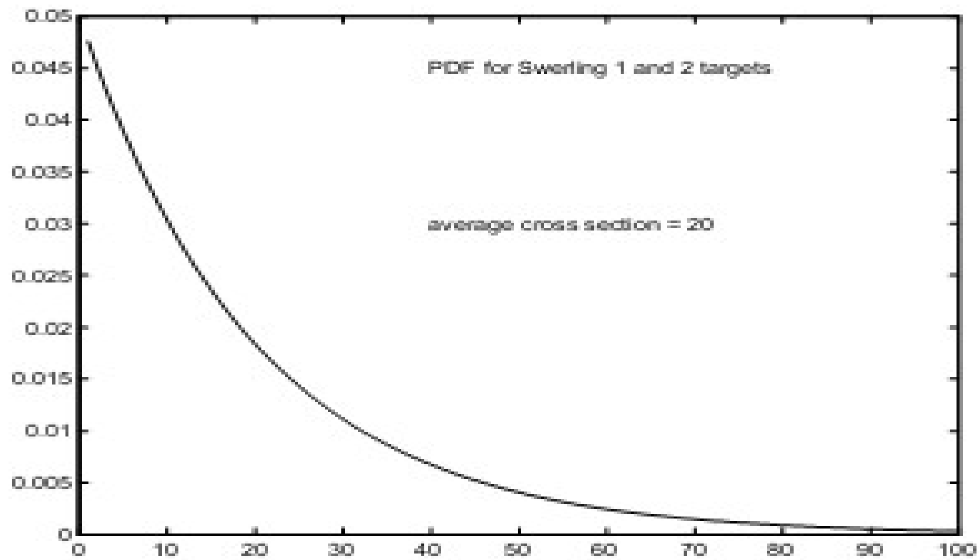
When detection probability is large, all 4 cases in which σ is not constant require greater

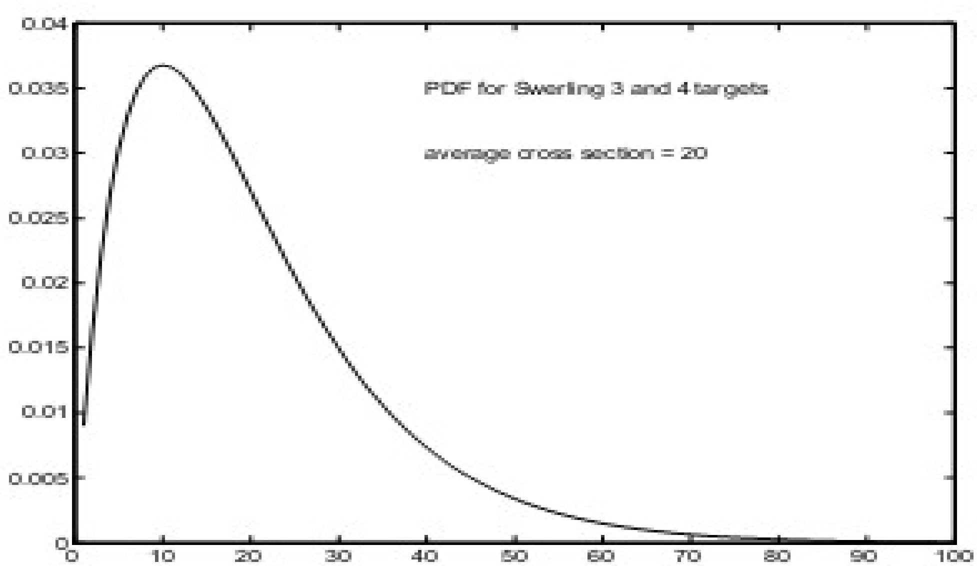
SNR than the constant σ case (case 5)

Note for P_d =0.95 we have

Case #	S/N
1	16.8 dB/pulse
2	6.2 dB/pulse

This increase in S/N corresponds to a reduction in range bya factor of 1.84. Hence if the characteristics of the target are not properly taken into account, the actual performance of the radar (for the same value of σave) will not measure up to the predicted performance.





Comparison of the five cases for a false alarm number n_f = 108 is shown in Fig. 2.22

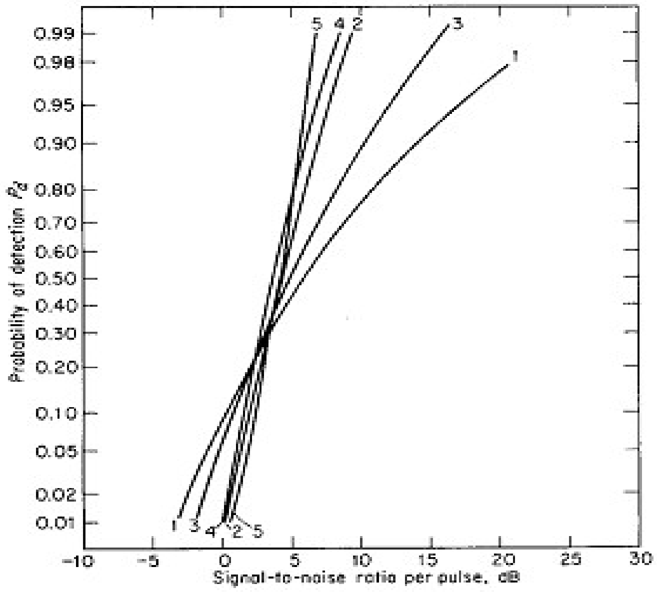


Figure 2.22 Comparison of detection probabilities for five different models of target fluctuation for $n = 10$ pulses integrated and false-alarm number $n_f = 10^8$. (*Adapted from Swerling.*[37])

Also when $P_d > 0.3$, larger S/N is required when fluctuations are uncorrelated scan to scan (cases 1 & 3) than when fluctuations are uncorrelated pulse to pulse.

This results since the larger the number of independent pulses integrated, the more likely

the fluctuations will average out ⇒ cases 2 & 4 will approach the nonfluctuating

case.

Figures 2.23 and 2.24 may be used as corrections for probability of detection (Fig. 2.7)

Procedure:

1) Find S/N from Fig. 2.7 corresponding to desired P_d and P_{fa}

2) From Fig. 2.23 find correction factor for either cases 1 and 2 or cases 3 and 4 to be applied to S/N found in Step 1. The resulting $(S/N)_1$ is that which would applyif detection were based on a single pulse

3) If n pulses are integrated, The integration improvement factor $I_i(n)$ is found from Fig.

2.24. The parameters $(S/N)_1$ and $nE_i(n)=I_i(n)$ are substituted into the radar equation 2.33 along with σ_{ave}.

Note: in Fig. 2.24 the integration improvement factor Ii(n) is sometimes greater than n. Here the S/N required fro n=1 is larger than for the nonfluctuating target. The S/N per pulse will always be less than that of the ideal predetection integrator.

Note: data in Fig. 2.23 and 2.24 are essentially independent of the false alarm number

(106<nf<1010).

Note: the PDF s for cases 1 &2 and # & 4 of the Swerling fluctuations are special cases of the

Chisquare distribution of degree 2m (also called the Gamma distribution)

$$p(\sigma) = \frac{m}{(m-1)!\sigma_{ave}}\left(\frac{m\sigma}{\sigma_{ave}}\right)^{m-1}\exp\left(\frac{-m\sigma}{\sigma_{ave}}\right) \qquad \sigma > 0$$

Note: For target cross section models, 2m is not required to be an integer. It maybe any positive real number.

For cases 1 and 2, m=1

For cases 3 and 4, m=2

Note: For the Chi-square PDF

$$\frac{\sqrt{\mu_2}}{m_1} = m^{-\frac{1}{2}}$$

here $\sqrt{\mu_2}$ is the standard deviation

and m_1 is the mean value

Note: as m increases, the fluctuations become more constrained. With $m = \infty$, we have the

nonfluctuating target.

The Chi-square distribution may not always fit observed data, but it is used for convenience. It is described by two parameters σave and the number of degrees of freedom 2m.

Aircraft flying straight and level fit Chi-square distribution with m between 0.9 and 2,

and with σ_{ave} varying 15 dB from min to max.

The parameters of the fitted distribution vary with aspect angle, type of aircraft and frequency. The value of m is near unity for all aspect angles except broadside which give a Rayleigh distribution with varying $\sigma_{ave.}$ It is found that σave has more effect on the calculation of probabilityof detection than the value of m.

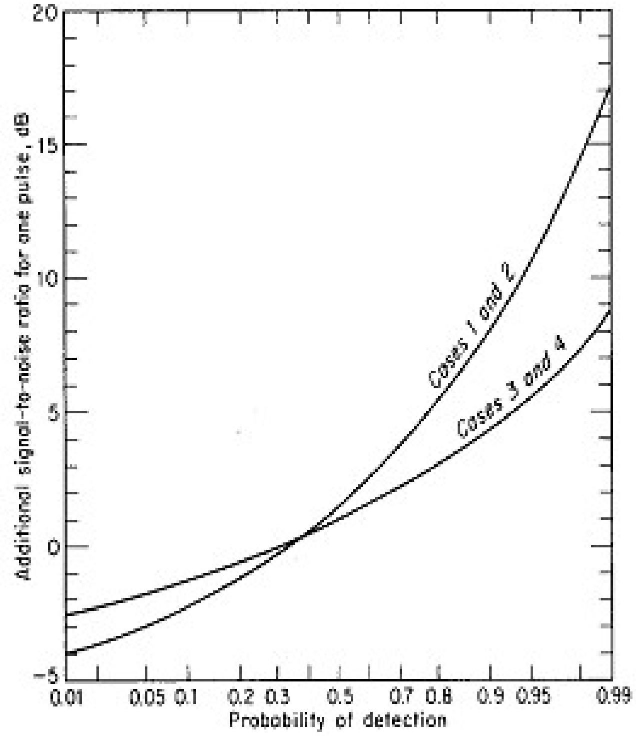


Figure 2.23 Additional signal-to-noise ratio required to achieve a particular probability of detection, when the target cross section fluctuates, as compared with a nonfluctuating target; single hit, $n = 1$. (To be used in conjunction with Fig. 2.7 to find $(S/N)_1$.)

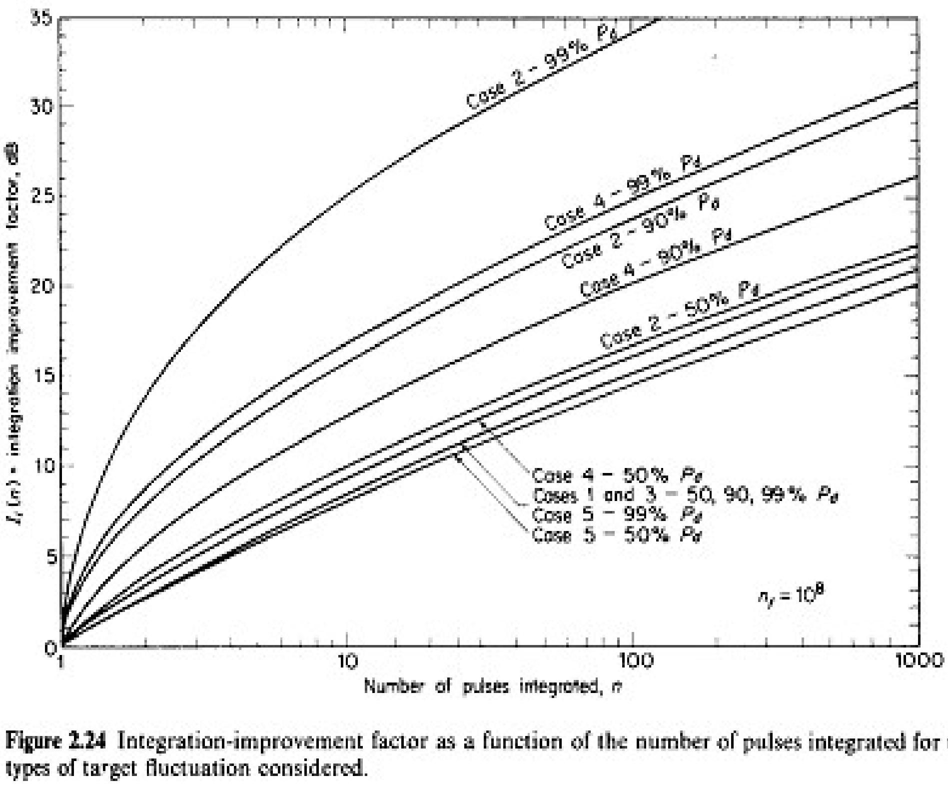


Figure 2.24 Integration-improvement factor as a function of the number of pulses integrated for the five types of target fluctuation considered.

The Chi-square distribution also describes the cross section of shapes such as cylinders, cylinders with fins (e.g. some satellites). Here m varies between 0.2 and 2 depending on the aspect angle. The Rice distribution is a better description of the cross section fluctuations of a target dominated by a single scatterer than the Chi-square distribution with m=2.

Here the Rice distribution is $p(\sigma) = \frac{1+s}{\sigma_{ave}} \exp\left[-s - \frac{\sigma}{\sigma_{ave}}(1+s)\right] I_0\left(2\sqrt{\frac{\sigma}{\sigma_{ave}} s(1+s)}\right)$

Where s is the ratio of the cross section of the single dominant scatterer to the total cross section of the smaller scatterers. I_0 is a modified Bessel function of zero order

Note: when s=1 the results using the Rice distribution approximate the Chi-square with m=2, for small probabilities of detection.

The Log Normal distribution has been suggested for describing the cross sections of some satellites, ships, cylinders, plates, arrays

$$p(\sigma) = \frac{1}{\sqrt{2\pi} s_d \sigma} \exp\left[-\frac{1}{2s_d^2}\left(\ln\left(\frac{\sigma}{\sigma_m}\right)\right)^2\right] \qquad \sigma > 0$$

Where s_d = standard deviation of and σ_m = median of

σ. Also the ratio of the mean to median value of σ is

Comparisons of several distributions models fro false alarm number nf = 106, with all pulses during a scan correlated and pulses in successive scans independent, are shown in Fig. 2.25.

Note: The two extreme cases treated are for pulses correlated in any particular scan but with scan-to-scan independence (slow fluctuations), and for complete independence (fast fluctuations).

There could be partial correlation of pulses within a scan. The results for this case would fall some where between the two cases.

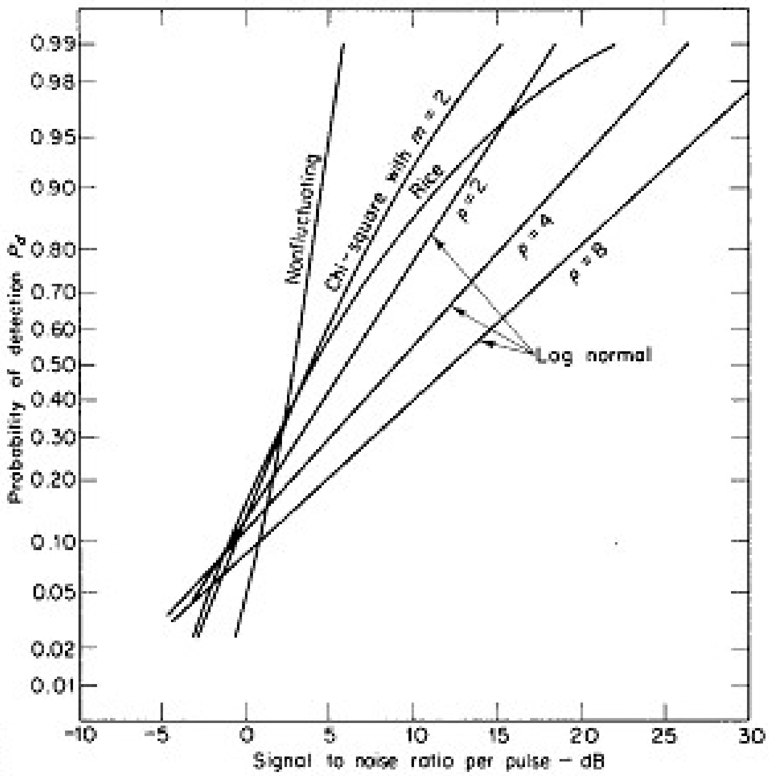


Figure 2.25 Comparison of detection probabilities for Rice, log normal, chi-square with $m = 2$ (Swerling case 3) and nonfluctuating target models with $n = 10$ hits and false-alarm number $n_f = 10^6$. Ratio of dominant-to-background equals unity ($s = 1$) for Rice distribution. Ratio of mean-to-median cross section for log-normal distribution $= \rho$.

Transmitter Power:

P_t in the radar equation is the peak power. This is not the instantaneous peak power of the carrier sine wave. It is the power averaged over a carrier cycle which occurs at the maximum of a pulse.

The average radar power, Pav is the average transmitter power over the PRI

$$P_{av} = P_t \frac{\tau}{T_p} = P_t \tau f_p$$

Here τ = pulse width, T_p= PRI and f_p = PRF

Now $\frac{P_{av}}{P_t} = \frac{\tau}{T_p}$ which defines the duty cycle

The typical duty cycle for surveillance radar is 0.001.

Thus the range equation in terms of average power is $R_{max}^4 = \frac{P_{av} G A_e \sigma n E_i(n)}{(4\pi)^2 k T_0 (B_n \tau) F_n (S/N)_1 f_p}$

Here $(B_n\tau)$ are grouped together since the product is usually of the order of unity for pulse radars.

If the transmitted waveform is not a rectangular pulse, we can express the range equation in terms of energy.

$$E_\tau = \frac{P_{av}}{f^p}$$

$$R_{max}^4 = \frac{E_\tau G A_e \sigma n E_i(n)}{(4\pi)^2 k T_0 (B_n \tau) F_n (S/N)_1}$$

Note: In this form R_{max} does not depend explicitly on λ or f_p

Pulse Repetition Frequency and Range Ambiguities:

PRF is determined primarily by the maximum range at which targets are expected. Echoes received after an interval exceeding the PRI are called "multiple-time-around" echoes. These can result in erroneous range measurements.

Consider three targets A, B and C. here A is within the maximum unambiguous range Runambig, B is between Runambig and $2R_{unambig}$ and C is between $2R_{unambig}$ and $3R_{unambig}$.

One wayof distinguishing multiple times around targets is to operate with a carrying PRF. The echo from an unambiguous target will appear at the same place on each sweep; however echoes from multiple time around targets will spread out.

The number of separate PRFs will depend on the degree of multiple time targets. Second time around targets need only 2 separate PRFs to be resolved.

Alternative methods to mark successive pulses to identify multiple times around targets include changing amplitude, pulse width, frequenc y, phase or polarization from pulse to pulse. These schemes are not very successful in practice.

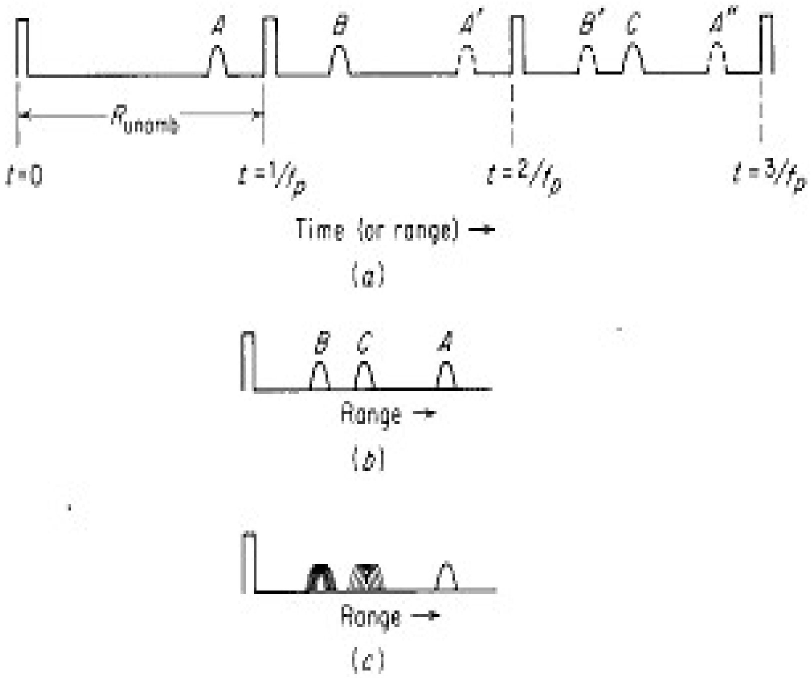


Figure 2.26 Multiple-time-around echoes that give rise to ambiguities in range. (*a*) Three targets A, B and C, where A is within R_{unamb}, and B and C are multiple-time-around targets; (*b*) the appearance of the three targets on the A-scope; (*c*) appearance of the three targets on the A-scope with a changing prf.

One limitation is the foldover of nearby targets (e.g. nearby strong ground targets, clutter) which can mask weak multiple time around targets. A second limitation is increased processing requirement to resolve ambiguities.

The range ambiguity in multiple PRF radar can be convenientlydecoded byuse of the Chinese remainder theorem.

Antenna Parameters:

The gain of an antenna is

$$G(\theta,\phi) = \frac{\text{power radiated per unit solid angle in direction of azimuth } \theta \text{ and elevation } \phi}{\text{power accepted by antenna from transmitter}/4\pi}$$

G is a function of direction. If it is greater than unityin some directions, it must be less than unityin others.

From reciprocity, if an antenna has a larger gain in transmission in a specific direction, then it also has a larger effective area in that direction.

The most common beam shapes fro radar are the pencil beam and the fan beam. Pencil beams are axially symmetric with a width of a few degrees. They are used where it is necessary to measure the angular position of a target continuously in azimuth and elevation (e.g.a tracking radar for weapons control or missile guidance). These are generated with parabolic reflectors.

To search a large sector of sky with a narrow beam is difficult. Operational requirements place restrictions on the maximum scan time (time for beam to return to the same point) so that the radar can not dwell too long at any particular cell.

To reduce the number of cells, the pencil beam is replaced with the fan beam which is narrow in one dimension and wide in the other.

Fan beams can be generated with parabolic reflectors with a shaped projected area. manylong range ground based radars use fan beams.

Even with fan beams, a trade-off exists between the rate at which the target position is updated (scan time) and the ability to detect weak signals (by use of pulse integration).

Scan rates are typically from 1 to 60 rpm. For long range surveillance, scan rates are typically 5 or 6 rpm.

Coverage of a simple fan beam is not adequate for targets at high altitudes close to the radar. The elevation pattern is usually shaped to radiate more energy at high angles as in the csc^2 pattern.

$$G(\phi) = G(\phi_0)\frac{\csc\phi^2}{\csc\phi_0^{\;2}} \qquad \text{for } \phi_0 < \phi < \phi_m$$

Here ϕ_0 and ϕ_m are the angular limits of the $\csc^2\phi$ fit

This pattern is used for airborne search radars observing ground targets as well as ground based radars observing aircraft. For the airborne case ϕ is the depression angle. Ideally ϕ_m should be 90˚ but it is always less. $Csc^2\phi$ patterns can be generated bya distorted section of a parabola or

with special multiple horn feed on a true parabola, or with an array such as a slotted waveguide.

The $\csc^2\phi$ pattern gives constant echo power Pr independent of range for a target of constant height, h and having a constant σ.

Substituting into the range equation (simple form)

$$P_r = \frac{P_t G^2(\phi_0)\csc(\phi)^2\lambda^2\sigma}{(4\pi)^3\csc(\phi_0)^2 R^4} = K_1\frac{\csc(\phi)}{R^4}$$

Now for a constant height, h of a target, we have $\csc(\phi) = R/h$

Therefore $P_r = \frac{K_1}{h^4} = K_2$ Hence the echo signal is independent of range.

In practice P_r varies due to σ varying with viewing angle, the earth not being flat and non perfect $\csc^2\phi$ patterns.

Note: the gain of $\csc^2\phi$ antennas for ground based radars is about 2 dB less than for a fan beam

having the same aperture.

The maximum gain of any antenna is related to its size by $G = \frac{4\pi A}{\lambda^2}\rho$

Where ρ is the antenna efficiency which depends on the aperture illumination

This is controlled by the complexity of the feed design. **Note:** $A_\rho = A_{eff}$

A typical reflector gives a beam width of $\theta(deg) = \frac{65\lambda}{l}$ where l is the dimension

System Losses:

Losses in the radar reduce the S/N at the receiver output. Losses which can be calculated include the antenna beam shape loss, the collapsing loss and the plumbing loss. Losses which cannot be calculated readily include those due to field degradation, operator fatigue and lack of operator motivation.

Note: loss has a value greater than unity - Loss = $[\text{Gain}]^{-1}$

1) ***Plumbing Loss***

- Loss in transmission lines between the transmitter and antenna and between antenna and receiver.
- Note from the Fig 2.28 that, at low frequencies, the transmission lines introduce little loss. At high frequencies the attenuation is significant
- Additional loss occurs at connectors (0.5 dB), bends (0.1dB) and at rotary joints

  (0.4 dB)

Note: If a line is used for both transmission and reception, its loss is added twice.

- The duplexer typically adds 1.5 dB insertion loss. In general, the greater the isolation required, the greater the insertion loss.

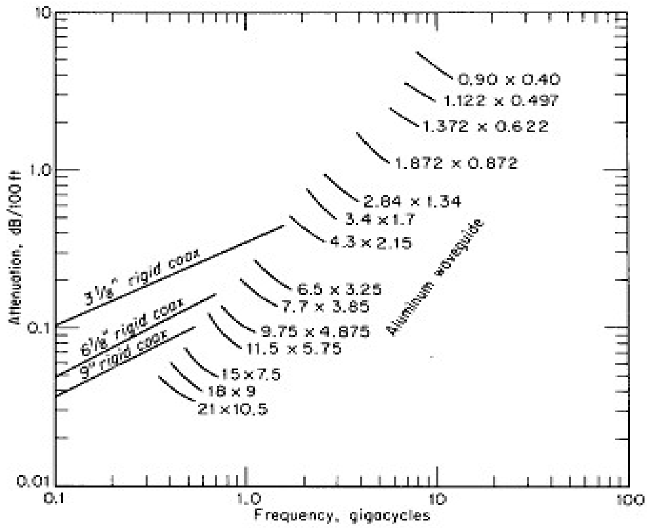


Figure 2.28 Theoretical (one-way) attenuation of RF transmission lines. Waveguide sizes are inches and are the inside dimensions. (*Data from Armed Services Index of R.F. Transmission Lines and Fittings, ASESA, 49-2B.*)

2) Beam Shape Loss

The train of pulses returned from the target to a scanning radar are modulated in amplitude by the shape of the antenna beam.

A beam shape loss accounts for the fact that the maximum gain is used in the radar equation rather than a gain which changes from pulse to pulse. (This approach is approximate since it does not address Pd for each pulse separately).

Let the one way power pattern be approximated by a Gaussian shape

$$S^2 = \exp\left[\frac{-2.78\theta^2}{\theta_B^2}\right]$$ Here θ_B is the half power beam width

n_B is the number of pulses received within θB and if n is the number of pulses integrated, then the beam shape loss (relative to a radar that integrated n pulses with equal gain) is

$$L(beamshape) = \frac{n}{1 + 2\sum_{k=1}^{(n-1)/2} \exp((-5.55k^2)/n_B^2)}$$

Example integrating 11 pulses gives L (beam shape) = 1.66 dB

Note: the beam shape loss above was for a beam shaped in one plane only (i.e. fan beam or pencil beam where the target passes through the centre of the beam).

If the target passe through any other part of the beam the maximum signal will not correspond to the signal from the beam centre.

When many pulses are integrated per beamwidth, the scanning loss is taken as 1.6 dB for a fan beam scanning in one coordinate, and as 3.2 dB when two coordinate scanning is used.

When the antenna scans so rapidly that the gain on transmission is not the same as the gain on reception, an additional “scanning loss” is added.

Additional loss for phased array search using a step scanning pencil beam since not all regions of space are illuminated by the same value of antenna gain.

3) Limiting Loss

Limiting in radar can lower the P_d. This is not a desirable effect and is due to a limited dynamic range. Limiting can be due to pulse compression processing and intensity modulation of CRT (such as PPI).

Limiting results in a loss of onlya fraction of a dB for large numbers of pulses integrated providing the limiting ratio (ratio of video limit level to RMS noise level) is greater than 2.

For small SNR in bandpass limiters, the reduction of SNR of a sine wave in narrowband

Gaussian noise is $\pi/4$ (approx 1 dB).

If the spectrum of the input noise is shaped correctly, this loss can be made ne gligible.

4) Collapsing Loss

If the radar integrates additional noise samples along with the wanted signal +noise pulses, the added noise causes degradation called the collapsing g loss. This occurs on displays which collapse range information (C scope which displays E_l vs A_z).

In some 3D radars (range, Az, El) that displayoutputs at all Elevations on one PPI (range, Az) display, the collapsing of the 3D information into 2 D display results in loss.

Can also occur when the output of a high resolution radar is displayed on a device which is of coarser resolution than the radar.

Marcum has shown that for a square law detector, the integration of m noise pulses, along with n signal + noise pulses with SNR per pulse $(S/N)_n$, is equivalent to the integration of m+n signal-to-noise pulses each with SNR of

$$\left(\frac{S}{N}\right)_{(m+n)equiv} = \frac{n}{m+n}\left(\frac{S}{N}\right)_n$$

The collapsing loss then is the ratio of the integration loss Li for m+n pulses to the integration loss foe n pulses

$$L_i(m,n) = \frac{L_i(m+n)}{L_i(n)}$$

- recall $L_i(n) = \frac{1}{E_i(n)}$

Example: 10 signal pulses are integrated with 30 noise pulses Required $P_d = 0.9$, $n^f = 10^8$

From Fig 2.8b. L_i (40) = 3.5 dB, L_i (10) = 1.7 dB

Therefore L_i (m,n) = 1.8 dB

Collapsing loss for a linear detector can be much greater than for a square law detector. Fig 2.29 shows the comparison of loss for each detector

5) Nonideal Equipment

Transmitter power - the power varies from tube to tube (for same type), and with age for a specific tube. Power is also not uniform over the operating band. Hence Pt may be other than the design value. To allow for this, a loss factor of about 2 dB can be used.

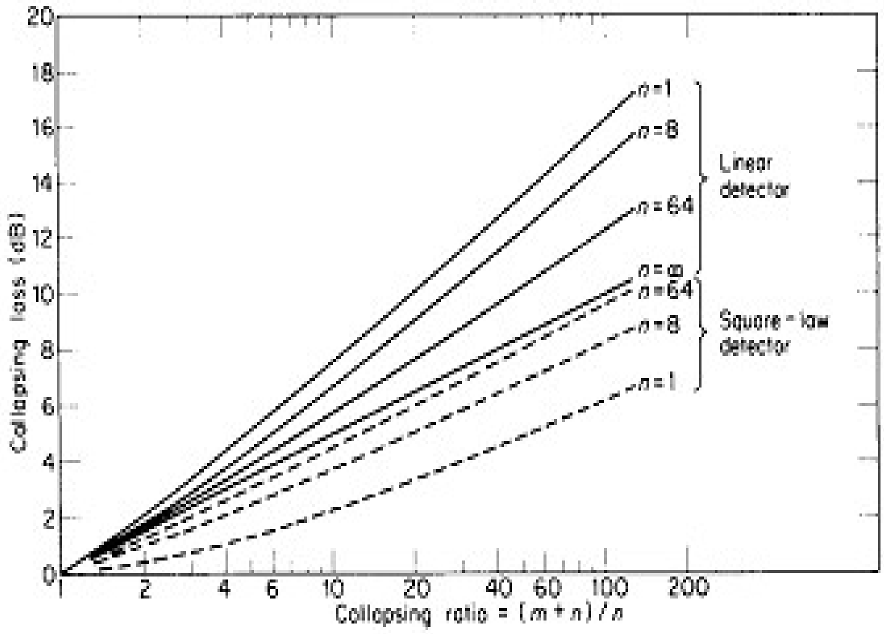


Figure 2.29 Collapsing loss versus collapsing ratio $(m + n)/n$, for a false alarm probability of 10^{-6} and a detection probability of 0.5. (*From Trunk,*[55] *courtesy Proc. IEEE.*)

Receiver noise figure: the NF will vary over the band, hence if the best NF is used in the radar equation, a loss factor must account for its poorer value elsewhere in the band.

Matched filter: if the receiver is not the exact matched filter fro the transmitted waveform, a loss of SNR will occur (typically 1 dB).

Threshold level: due to the exponential relationship between T_{fa} and V_T a slight change in V_T can cause significant change to Tfa hence, V_T is set slightly higher than calculated to give good T_{fa} in the event of circuit drifts. This is equivalent to a loss.

6) Operator Loss

A distracted, tired, overloaded, poorly trained operator will perform less efficiently. The operator efficiency factor (empirical) is where Pd is the single scan probability of detection.

Note: operator loss is not relevant to systems where automatic detection is done

7) Field Degradation

When a radar is operated under field conditions, the performance deteriorates even more than can be accounted for in the above losses.

Factors which cause field degradation are:

- poor training

- weak tubes
- water in the transmission lines
- incorrect mixer crystal current
- deterioration in the receiver NF
- poor TR tube recovery
- loose cable connections

Radars should be designed with BIST (built - in system test) and BITE (built - in test equipment) to aid in performance monitoring. A preventative maintenance plan should be used.

BITE parameters to be monitored are

- Transmitted power P_t

- NF of receiver
- Transmitter pulse shape
- Recovery time of TR tube

With no other information available, 3 dB is assumed for field degradation loss

8) Other Loss Factors

MTI radars introduce additional loss. The MTI discrimination technique results in complete loss of sensitivity for certain target values (blind speeds).

In a radar with overlapping range gates, the gates may be wider than optimum for practical reasons.

The additional noise introduced bynonoptimum gate width leads to degradation performance.

Straddling loss accounts for loss in SNR for targets not at the centre of a range gate, or at the centre of a filter in a multiple bank processor

9) Propagation Effects

The radar equation assumes free space propagation. The earth's surface and atmosphere have a significant effect on radar performance.

The effects fall into three categories:

- attenuation
- refraction by the earth's atmosphere
- lobe structure caused by interference between the direct wave and the ground reflected wave

For most microwave radars, attenuation through the normal atmosphere or through precipitation is not significant. However reflection from rain (clutter) is a limiting factor in radar performance in adverse weather.

The deceasing density of atmosphere with altitude results in bending (refraction) of the electromagnetic wave. This normally increases the line of sight. the refraction can also be accounted for by assuming the earth to have a larger radius than actual. A "typical" earth radius

is 4/3 actual radius.

At times atmospheric conditions create ducting (or super refraction) and increases the radar range considerably. It is not necessarily desirable since it can not be counted on. Also it de grades MTI performance by extending the range at which ground clutter is seen.

The presence of the earth also breaks the antenna elevation pattern into many lobes. this arises since the direct and reflected waves interfere at the target either destructively or constructively to produce nulls or lobes. This results in non uniform illumination.

Other Considerations:

The radar equation is now written.

$$R_{max}^4 = \frac{P_{av} G(A\rho_a)\sigma n E_i(n)}{(4\pi)^2 k T_0 (B_n \tau) F_n (S/N)_1 f_p L_s}$$

Note: The following substitutions can be made:

$E\tau = P_{av}/f_p = P_{t\tau}$

$N_0 = N/B$ (power spectral density of noise)

$B_\tau \approx 1$ and $T_0F_n = T_s$

Note: The above radar equation was derived for rectangular pulses but applies to other waveforms provided that matched filter detection is used. The equation can be modified to accommodate CW, FM-CW, pulse doppler MTI or tracking radar.

Radar Performance Figure - ratio of pulse power of Transmitter to S_{min} of receiver not often used

Blip-Scan ratio - same as single scan P_d.

- method used to check performance of ground-based radars

- here an aircraft is flown on a radial course and for each scan of the antenna it is recorded whether or not a target blip is detected. The ration of the number of scans the target was seen at a particular range to the total number of scans is the blip scan ratio
- head on and tail on aspects are easiest to provide.

Cumulative Probability of Detection:

If single scan probability of detection id Pd, the probability of detecting a target at least once during N scans is the cumulative probability of detection

$$P_c = 1-(1-P_d)N$$

Note: the variation of P_d with range might have to be taken into account in computing P_c.

The variation with range based on the cumulative probability of detection can be the 3rd power rather than the 4th power which is based on a single scan probability. In practice P_c is not easy to apply. Furthermore radar operators do not usually report a detection the first time it is observed (which is required by the definition of P_c). Instead theyreport a detection based on threshold crossing on two successive scans, or on two out of three scans.

For track while scan radars, the measure of performance might be the probabilityof initiating a target track rather than just probability of detection.

Surveillance Radar Equation:

The radar equation which describes the performance of a radar which dwells on the target for n pulses is sometimes called the searchlight range equation.

In a search or surveillance radar, the additional constraint that the radar must search a specified volume of space in a specified time modifies the range equation significantly.

If Ω represents the angular region to be searched in scan time ts, then we have $t_s = t_0 \frac{\Omega}{\Omega_0}$

Where t_0 is the time on target = n/f_p

Ω_0 = solid angle beamwidth and $\Omega_0 \approx \theta_A \theta_E$

Where θ_A is the Azimuth beamwidth And θ_E is the Elevation beamwidth

Also $G = \frac{4\pi}{\Omega_0}$

With these substitutions the range equation becomes $R^4_{max} = \frac{P_{av} A_e \sigma E_i(n) t_s}{4\pi k T_0 F_n (S/N)_1 L_s \Omega}$

This indicates that the important parameters for a search radar are the average power and the antenna effective aperture.

Frequency does not appear explicitly, however low frequency is preferred since high power and large aperture are easier to obtain at low frequencyand it is easier to build MTI (moving target indicator) and weather has little effect on performance.

Note: the radar equation will be considerably different if clutter or external noise (jamming) rather than receiver noise determine the background for the signal

Accuracy of the Radar Equation:

The predicted value of range from the range equation cannot be checked experimentally with any accuracy. The safest means to achieve a specified range performance is to include a safety factor. This is sometimes difficult to do in competitive bids but results in fine radars.

OBJECTIVE TYPE QUESTIONS

1. A high noise figure in a receiver means---------- [ ]

a. poor minimum detectable signal b. good detectable signal

c. receiver bandwidth is reduced d. high power loss.

2. Which of the following will be the best scanning system for tracking after a target has been acquired.

a. conical b. spiral c. Helical d. Nodding

3. Which of the following noise figure. [ ]

a. (Si Ni) / (So No) b. (So No) / (Si Ni)

c. (So / No) / sqrt. (Si / Ni d. (Si / Ni) / sqrt .(So / No

4. The average power of a pulsed radar transmitter is given by [ ]

a. The product of peak power of the pulse and the duty cycle

b. Peak power divided by the number of pulses repeated in one

second. c. Peak power divided by the duty cycle

d. Peak power divided by the duty cycle and pulse

5. Which of the following diode is used as detector in radar. [ ]

a. gunn diode b. schotky diode c. Impact diode d. varactor diode

6. In case the target cross section is changing the best system for accurate tracking is [ ]

a. monopulse b. lobe switching c. sequential lobing d. conical scanning.

7. In a radar in case the return echo arrives after the allocated pulse interval, then [ ]

a. it will not be received b. the receiver will get overloaded

c. it may interfere with the operation of the

transmitter d. the target will appear closer

than it really.

8. PPI in a radar system stands for [ ]

a. plan position indicator b. pulse position indicator

c. plan position image d. prior position identification

9. Which of the following is unlikely to be used as a pulsed device [ ]

a. TWT b. BWO c. CFA d. Multicavity klystron

10. Radar detection is limited to line of sight because [ ]

a. curvature of the earth b. the waves are not reflected by the

ionosphere c. long wavelengths are used d. short wavelengths are

used

11. Second time around echoes are caused by [ ]

a. Second time reflection from target

b. echoes returning from targets beyond the cathode ray tube range. c. echoes that arrive after transmission of the next pulse.

d. extreme ends of bandwidth.

12. The resolution of pulsed radar can be improved by [ ]

a. increasing pulse width b. decreasing pulse width

c. increasing the pulse amplitude d. decreasing the pulse repetition frequency.

13. The most important application of monopulse antenna is in [ ]

a. determining the range of target b. tracking a target

c. identifying a target d. Isolating the track of target.

14. In case the antenna diameter in a radar system is increased to four times. The maximum range will increase by [ ]

a. √2 times b. 2 times c. 4 times d. 8 times.

15. In case the ratio of the antenna diameter to the wavelength in a radar system is high, this is likely not to result in [ ]

a. increased capture area b. good target discrimination c. difficult target acquisition d. large maximum range

16. The term RADAR stands for [ ]

a. radio direction and reflection b. radio detection and ranging

c. radio waves dispatching and receiving d. random detection and re radiation.

17. The duty cycle in a pulsed radar transmitter cannot be increased beyond a point

because it [

]

a. affects the operating frequency

b. increase the average power of the

transmitter tube. c. does not detect weak

signals

d. increase minor lobes

18. In case of radar receiver the IF bandwidth is inversely proportional to [ ]

a. pulse interval

b. pulse repetition frequency

c. square root of the peak transmitted power

d. pulse width.

19. The Doppler effect is used in [ ]

a. MTI b. CW c. FM d. Radar Altimeter

20. The gain of a radar transmitting antenna is [ ]

a. loss than that of radar receiving antenna

b. almost equal to that of radar receiving antenna

c. slightly higher than that of radar receiving antenna d. much higher than that of radar receiving antenna.

Answers:

1.a	2.a	3.a	4.a	5.c	6.a	7.d	8.b	9.b	10.a

11.c	12.b	13.b	14.c	15.a	16.b	17.b	18.d	19.d	20.d

ESSAY TYPE QUESTIONS

1. Describe how threshold level for detection is decided in the presence of receiver noise for a specified probability of occurrence of false alarms.

2. Describe the effect of pulse repetition frequency on the estimated unambiguous range of radar.

3. Obtain the SNR at the output of IF amplifier of radar receiver for a specified probability of detection without exceeding a specified probability of false alarm.

4. Explain system losses will effect on the radar range?

5. Discuss about the factors that influence the prediction of radar range.

6. Define noise bandwidth of a radar receiver. How does it differ from 3 db band width? Obtain the expression for minimum detectable signal in terms of noise bandwidth, noise figure and other relevant parameters.

7. If the noise figure of a receiver is 2.5 db, what reduction (measured in dB) occurs in the signal noise ratio at the output compared to the signal noise ratio at the input?

8. Describe the effect of (in terms of wavelength of operation) size of a simple spherical target on determination of radar cross section of the sphere.

9. What are multiple-time-around echoes? Explain the relation between unambiguous range estimation and multiple-time-around echoes.

10. Establish a relation between the probability of false alarm and detection threshold level of
a radar receiver in the presence of noise.

11. Estimate the radar cross-section of a spherical target if the wavelength of transmitting signal with reference to the target size is in Rayleigh region.

12. Justify the requirement of integration of radar pulses to improve target detection process.

13. List all the possible losses in a radar system and discuss the possible causes of each of them.

14. Discuss about the factors that influence the prediction of Radar range.

15. Define noise bandwidth of a radar receiver. How does it differ from 3-dB bandwidth?Obtain the expression for minimum detectable signal in terms of noise bandwidth, noise figure and other relevant parameters.

16. Explain the principle and process of binary moving window detector.

17. Obtain the SNR at the output of IF amplifier of Radar Receiver for a specified probability of detection without exceeding a specified probability of false alarm.

18. Explain how system losses will a effect on the Radar Range?

19. Determine the probability of detection of the Radar for a process of threshold detection with a graphic illustration.

UNIT-III

CW AND

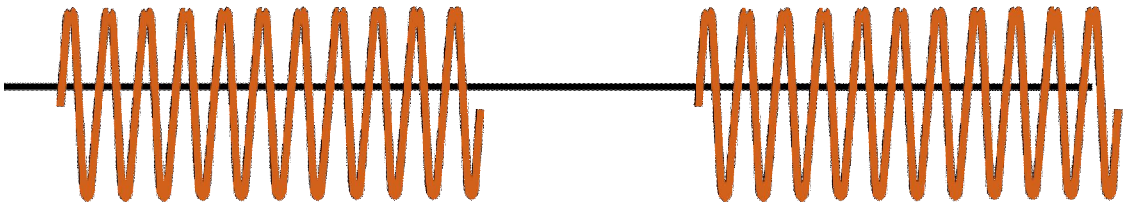
FREQUENCY

MODULATED

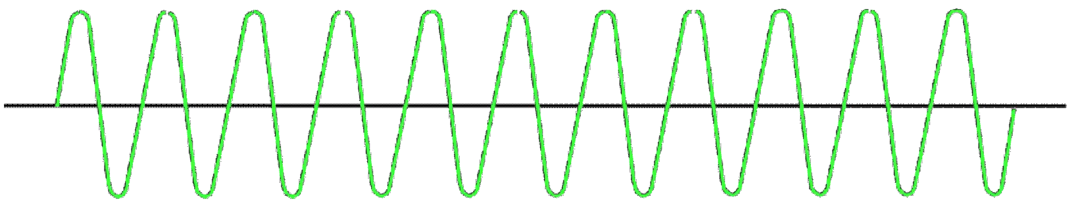
RADAR

- CW radar detects objects and measures velocity from Doppler shift.
- CW radar sets transmit a high-frequency signal continuously.
- It can’t measure range.
- It can be mono-static or bi-static.
- Employs continual RADAR transmission
- Separate transmit and receive antennas
- Relies on the “DOPPLER SHIFT”

- **Pulse Transmission**



- **Continuous Wave**



PRINCIPLE OF DOPPLER EFFECT:

- The radars radiate electromagnetic waves towards the targets for detection and also to obtain details of the target.
- When the target is stationary, the frequency of the received echoes is constant.
- However, when the target is moving, the frequency of the received echoes is found to be different from transmitted frequency.
- If the target approaches the radar, the frequency is increased and if the target moves away from the radar, the frequency is decreased.
- That is, in the moving targets, there exists a frequency shift in the received echo signals.
- The presence of frequency shift in the received echo signals in the radar due to moving targets is known as Doppler Effect.
- The frequency shift is known as Doppler frequency shift

Doppler Effect:

When there is a relative motion between Radar and target is based on recognizing the change in the echo-signal frequency caused by the Doppler Effect.

If either observer or the source is in motion, then it results in a frequency shift, the resultant frequency shift is known as Doppler Effect.

When an observer moves relative to a source, there is an apparent shift in frequency is known as Doppler frequency.

If distance between observer and the source is increasing, the frequency apparently decreases, where as the frequency apparently increases if the distance between the observer and the source is decreasing is known as Doppler frequency.

$$2\pi f_d = 4\pi v_r/\lambda$$

$$\mathbf{f_d = 2v_r/\lambda} \qquad\qquad \mathbf{f_d = 2v_r/\lambda}$$

Where f_d = Doppler frequency shift in hertz

v_r = relative velocity or radial velocity of target with respect Radar in knots.

$= dR/dt$

λ = wavelength in meters = c/f_0

The distance R and wavelength λ are measured in same units.

The Doppler frequency shift is $f_d = 2v_r/\lambda = 2v_r/(c/f_{0)}$

$$f_d = 2v_rf_0/c \ = \ —— \qquad\qquad f_d = 2v_rf_0/c$$

Where f_0 = transmitted frequency

c = velocity of propagation = 3×10^8 m/sec.

If f_d in Hz, v_r in knots and λ in meters: $f_d = 1.03v_r/\lambda$ $\qquad f_d = 1.03v_r/\lambda$

The relative velocity may be written $v_r = v\cos\theta$

Where v is the target speed and θ is the angle made by the target trajectory and the line joining

Radar and target.

$$\text{Therefore, } f_d = 2v\cos\theta/\lambda$$

When $\theta = 0^0$, the Doppler frequency is maximum. The Doppler frequency is zero, when

the trajectory is perpendicular to the Radar line of sight ($\theta = 90^0$).

CW RADAR:

A Simple CW Radar is shown in fig1 (a). The CW Radar consists of a transmitter, detector, beat-frequency amplifier and indicator.

The transmitter transmits a continuous wave of frequency oscillation or oscillation of frequency f_0, which is radiated by the antenna. An amount of radiated energy is intercepted by the target and some of this energy is scattered back in the direction of the Radar. This energy is

Collected by the receiving antenna.

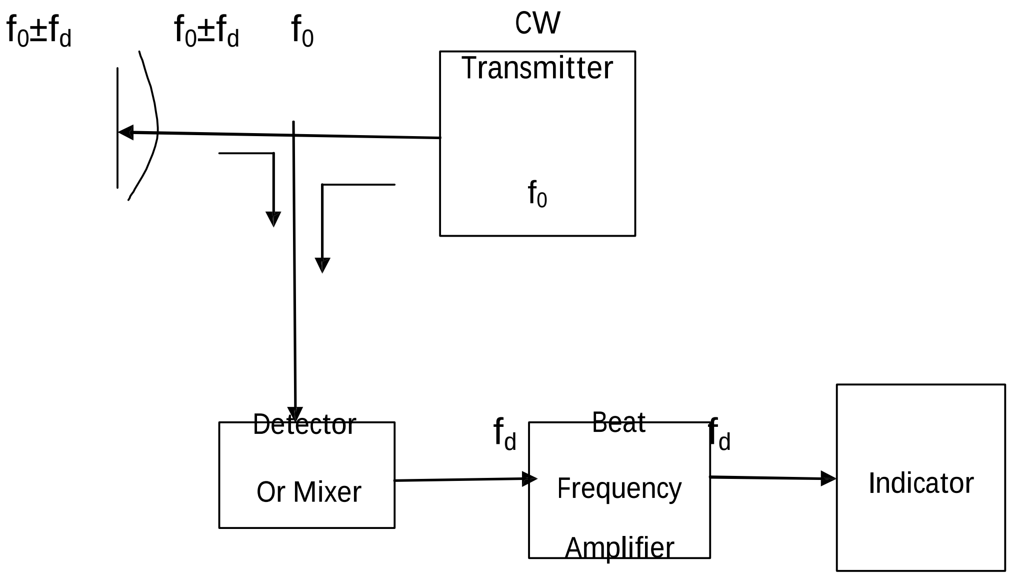
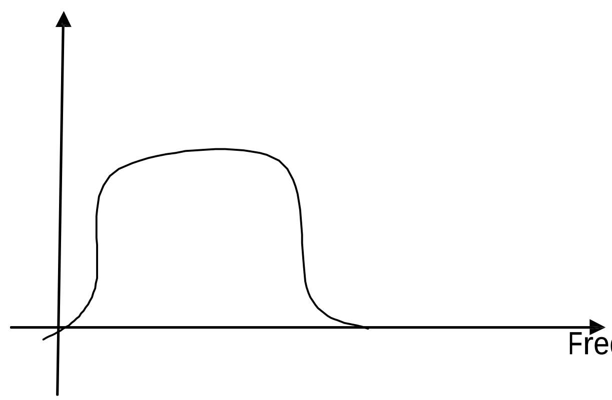

f0±fd
f0±fd
f0
CW
Transmitter
f0
Detector
Or Mixer
fd
Beat
Frequency
Amplifier
fd
Indicator

Fig 1(a): Simple CW Radar Block Diagram

Response



Fib 1(b): Response characteristic of beat-frequency amplifier

If the target is in motion with a relative velocity v_r to the Radar, then the received signal will be shifted by an amount of '$\pm f_d$'.

The plus sign associated with the Doppler frequency applies if the distance between target and Radar is decreasing (closing target), i.e., when the received signal frequency is greater than the transmitted signal frequency. The minus sign associated with the Doppler frequency applies if the distance between target and Radar is increasing (receding target), i.e., when the received signal frequency is less than the transmitted signal frequency.

The received echo signal at a frequency $f_0 \pm f_d$ enters the Radar via the antenna and is heterodyne in the detector with a part of the transmitted signal f_0 to produce a Doppler beat frequency $f_{d.}$ The sign is lost in this process.

The purpose of Doppler amplifier is to eliminate the echoes from stationary targets and to amplify the Doppler echo signal to a level where it can be used to operate an indicating device. It must have a frequency characteristic similar to that of fig1 (b).

The low frequency cut-off must be high enough to reject the dc component caused by stationary targets. The upper cut-off frequency is selected to pass the higest Doppler frequency expected. Finally, an indicator devices used, generally it consists of a pair of earphones or a frequency meter.

Isolation between transmitter and receiver:

The main impotence of providing isolation between transmitter and receiver is to eliminate the transmitter leakage signal.

Generally separate antennas are used for transmission and reception, so that there is no chance of leakage entering the receiver. Even the isolation between transmitter and receiver is possible using a single antenna as in CW Radar.

The noise that accompanies the transmitter leakage signal will determine the amount of isolation needed in a long range CW Radar. For example, a 10 mwatt of leakage signal is appeared at the receiver for a proper isolation between transmitter and receiver. The transmitter noise must be at least 110 dB below the transmitted carrier for a minimum detectable signal of 10^{-13} watt.

The isolation between transmitter and receiver can be obtained with a single antenna (like CW Radars) by using a hybrid junction, circulator, turnstile junction or with separate polarizations. Even though the isolation achieved by hybrid junctions such as magic Tee, rat race or directional coupler is 60 dB in extreme cases, the isolation in practical cases is limited 20 or

30 dB.

The limitation of 6dB on overall performance using junctions is going to waste half the transmitted power and half the received power. Thus, hybrid junctions are applicable to short range Radars. Similarly, ferrite isolation devices and turnstile junctions are also limited to short range Radar due to the difficulty in obtaining large isolations.

The large isolations are obtained in CW tracker-illuminator using separate antennas for transmission and reception. By the proper insertion of a controlled sample of the transmitted signal directly into the receiver, additional isolation can be obtained. The part of transmitted signal that leaks into the receiver can be cancelled by adjusting the phase and amplitude of the "buck-off" signal. The arrangement introduces additional 10dB isolation. But the phase and amplitude of the leakage signal may vary as the antenna scans. Thus a dynamic canceller can be used that senses the proper phase and amplitude of leakage signal for obtaining the additional isolation. Thus, the above dynamic cancellation of leakage signal can exceed isolation to 30dB.

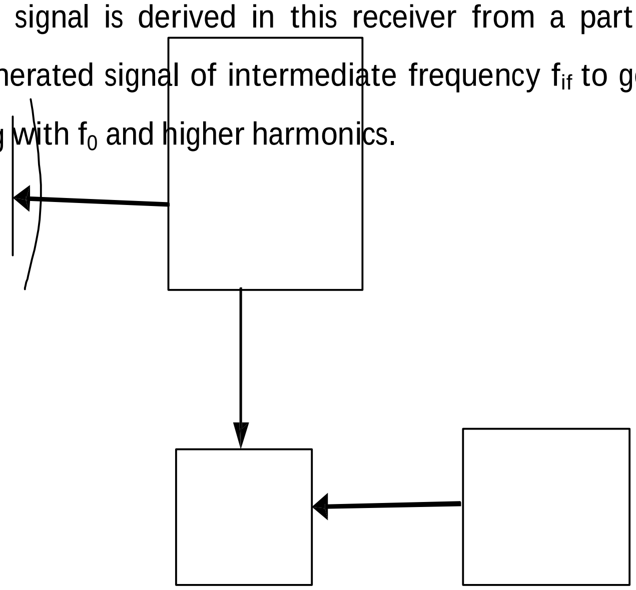
Intermediate-Frequency Receiver:

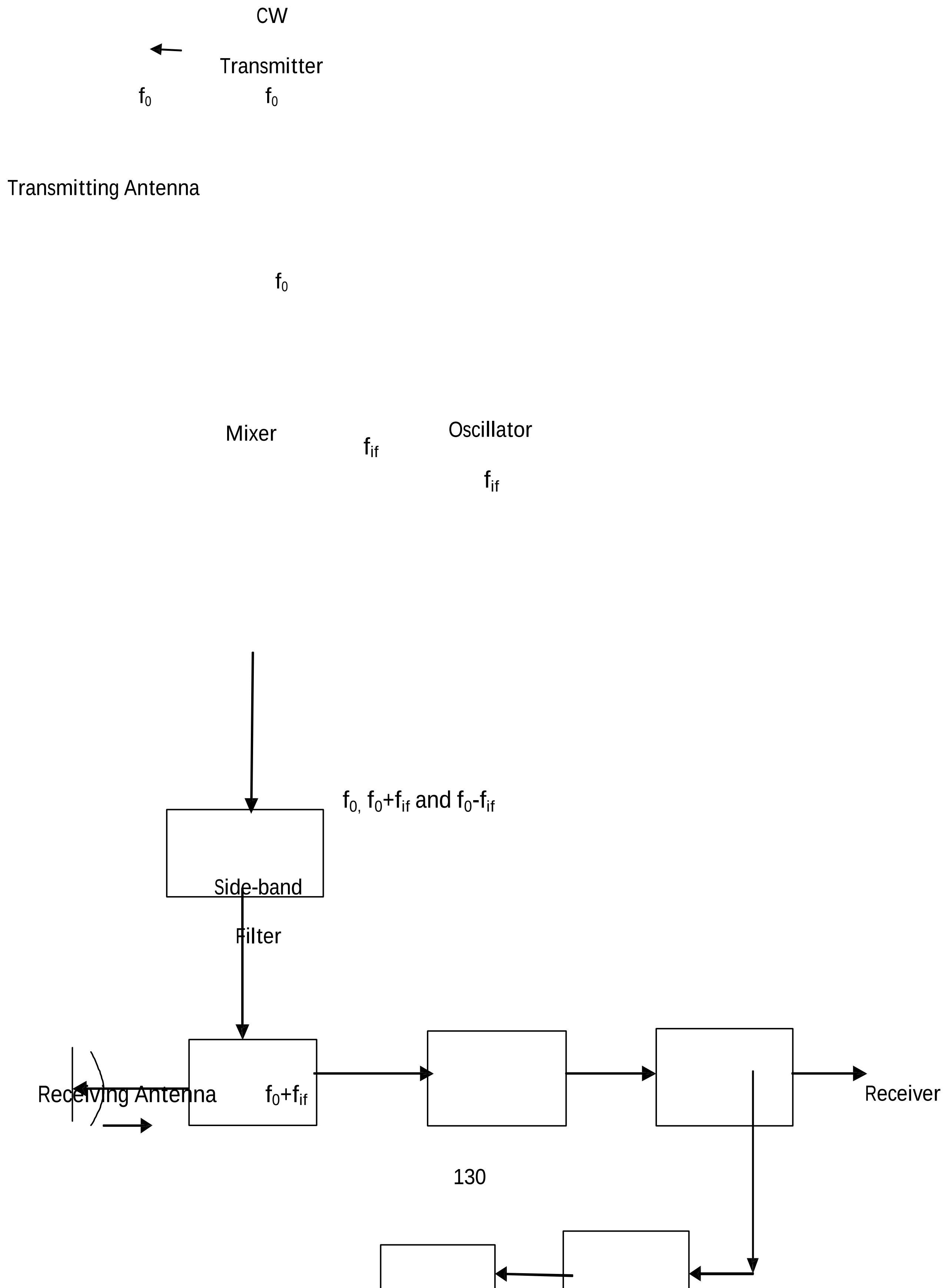
The receiver of the simple block diagram of CW Radar is in some respect analogous to a super heterodyne receiver. Receivers of this type are called homodyne receivers or super heterodyne receivers with zero.

The function of the local oscillator is replaced by the leakage signal from the transmitter. Such a receiver is simpler than one with a more conventional intermediate frequency since no IF amplifier or local oscillator is required. However, the simple receiver or zero IF receiver is not as sensitive because of increased noise at lower intermediate frequency caused by flicker effect.

Flicker effect noise occurs in semiconductor devices such as diode detector and cathodes of vacuum tubes. The noise power produced by flicker effect varies as $1/f^{\alpha}$, where α is approximately unity.

The fig (2) shows a block diagram of CW Radar whose reciver operates with a non-zero IF. In this case, separate antennas are shown for transmission and reception. Instead of the local oscillator determined in the conventional super heterodyne receiver, the local oscillator or reference signal is derived in this receiver from a part of transmitted signal mixed with a locallyGenerated signal of intermediate frequency f_{if} to generate the two side bands f_0+f_{if} and f_0-f_{if} along with f_0 and higher harmonics.




CW
Transmitter
f_0
f_0
Transmitting Antenna
f_0
Mixer
f_{if}
Oscillator
f_{if}
$f_{0,}$ f_0+f_{if} and f_0-f_{if}
Side-band
Filter
Receiving Antenna
f_0+f_{if}
Receiver

Mixer

$f_{if} \pm f_d$

IF Amplifier

Second Detector

$f_0 \pm f_d$

Indicator

Doppler

F_d

Amplifier

Fig (2): Block diagram of CW Doppler radar with non-zero IF receiver, sometimes called side-band super heterodyne.

From this signal one of the side band is selected by passing it through narrowband filter as the reference signal.

The purpose of Doppler amplifier is to eliminate the echoes from stationary targets and to amplify the Doppler echo signal to a level where it can be used to operate an indicating device, such as a pair of earphones or a frequency meter.

Comparison between zero and non-zero IF receivers:

1. Zero IF receiver is not as sensitive because of increased noise at lower intermediate frequency caused by flicker effect.

2. The reduction in sensitivity has greater effect on the maximum efficiency with CW Radar.

3. The improvement in receiver sensitivity with an non-zero IF receiver might be around 30 dB over the zero IF receiver.

Receiver Bandwidth:

The factors which tend to spread to the CW signal energy over a finite band of frequencies are discussed in the following.

If the received waveform were a sine infinite duration, its frequency spectrum would be a delta function and the receiver bandwidth would be infinitesimal. But a sine wave of infinite duration and an infinitesimal bandwidth does not exist in nature. So, considering an echo signal which a sine wave of finite duration rather than infinite duration. Then the spectrum of finite duration sine wave is given by $\sin[\pi(f-f_0)\delta]/\pi(f-f_0)$

Where f_0 = frequency of sine wave

δ = duration of sine wave

f = frequency variable over which is spectrum is plotted as shown in fig (3b).

The above characteristic is approximated by the practical receivers. In practical aspects, the echo may not be a pure sine wave, so we need to broaden the bandwidth still further. Assuming that for a CW Radar, the duration of the received signal is given by $\delta = \theta_B/\dot{\theta}_S$

i.e., time taken on target = $\delta = \theta_B/\dot{\theta}_S$

Where θ_B = antenna beam width in degrees and $\theta^{\cdot}_S$ = antenna scanning rate in degrees/sec

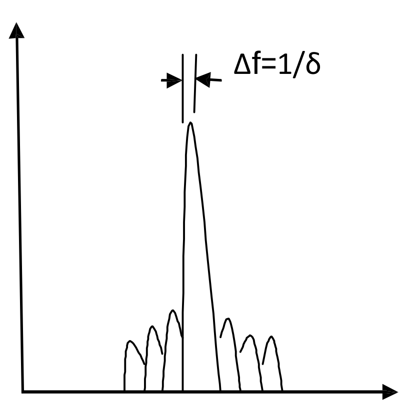
Thus, the signal is of finite duration and bandwidth of the receiver must be of the order of the reciprocal of the time on target $\theta^{\cdot}_S/\theta_B$. For example, if the antenna beam width and antenna scanning rate are 2^0 and 36^0/sec (6 rpm) respectively. Then the spread in the spectrum of the received signal due to the finite time on the target is 18 Hz, which is independent of the

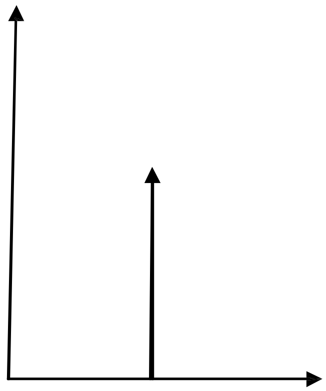
transmitted frequency.

In addition to the spread of the received signal spectrum caused by the finite time on the target, the spectrum may be further widened, if the target cross section fluctuates.

Energy/Hz of bandwidth

Energy/Hz of bandwidth





f_0 Frequency f_0 Frequency

(a) Infinite duration (b) Finite duration

Fig(3): Frequency spectrum of CW oscillation

The echo signal from a propeller driven aircraft can also contain modulation component at a frequency proportional to the propeller rotation.

The frequency range propeller modulation depends upon the shaft rotation speed and the number of propeller blades. Propeller may cause an error in measurement of Doppler frequency. This modulation may also be advantageous in the detection of propeller driven aircraft by passing tangential trajectory even in the absence of Doppler frequency shift.

The modulation of echo signal is also resulted by rotating blades of a helicopter and the compressor stages of a jet engine and which will degrade the performance of the CW Radar by widening the sacrum.

A further widening of the received signal spectrum can occur, if the relative velocity of the target is not constant. If a_r is the acceleration of the target with respect to the Radar, the signal will occupy a bandwidth: $\Delta f_d = (2a_r/\lambda)^{1/2}$

If for example, a_r is twice the acceleration of gravity, the receiver bandwidth must be approximately 20 Hz when Radars wavelength is 10cm.

Filter Bank in CW Radar receiver:

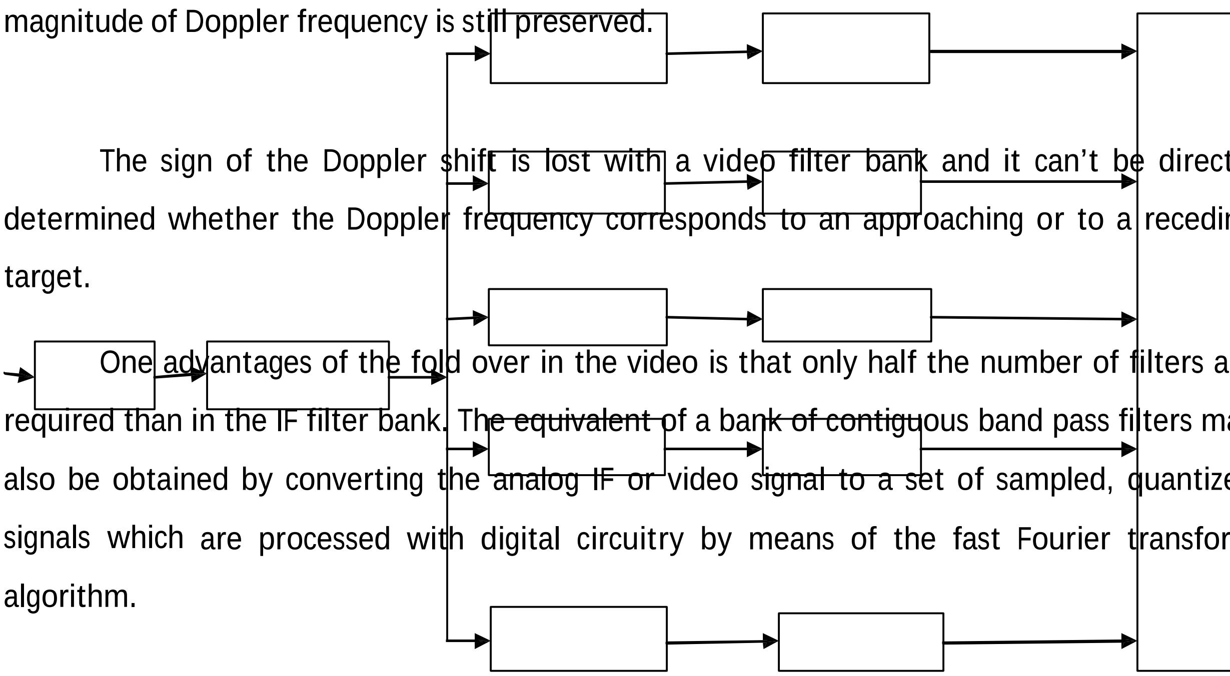
A relative wide band of frequencies called as bank of narrow band filters are used to measure the frequency of echo signal. These are also used to improve the signal to noise ratio of the receiver. The bandwidth of each individual filter is such that, it accepts the signal energy but should be taken that it does not introduce more noise because of wide bandwidth.

The center frequencies of the filters are staggered to cover the entire range of Doppler frequencies. If the filters are spaced with their half power points overlapped, the maximum reduction in signal to noise ratio of a signal which lies midway between adjacent channels compared with the signal to noise ratio at midband is 3 dB. By using the large number of filters, the maximum loss will be reduced but it increases the probability of false alarm. The fig 4(a) shows the block diagram of IF Doppler filter bank.

A bank of narrow band filters may be used after the detector in the video of the simple CW Radar, instead of in the IF. The improvement in signal to noise ratio with a video filter bank is not as good as can be obtained with an IF filter bank, but the ability to measure the magnitude of Doppler frequency is still preserved.

The sign of the Doppler shift is lost with a video filter bank and it can't be directly determined whether the Doppler frequency corresponds to an approaching or to a receding target.

One advantages of the fold over in the video is that only half the number of filters are required than in the IF filter bank. The equivalent of a bank of contiguous band pass filters may also be obtained by converting the analog IF or video signal to a set of sampled, quantized signals which are processed with digital circuitry by means of the fast Fourier transform algorithm.



Filter No.1 Detector I

N

Filter No.2 Detector D

I

Filter No.3 Detector C

A

Mixer IF amplifier

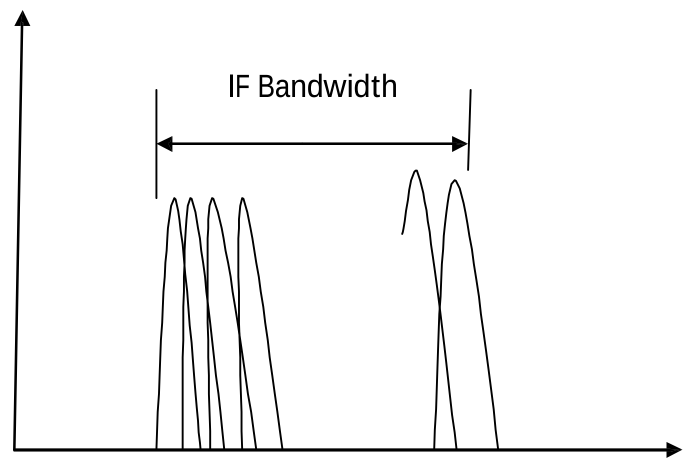
Filter No.4 Detector T O

R

Filter No.n Detector

Fig 4(a): Block diagram of IF Doppler filter bank

Response



f_1 f_2 f_3 f_4 f_n Frequency

Fig 4(b): Frequency-Response characteristic of Doppler filter bamk.

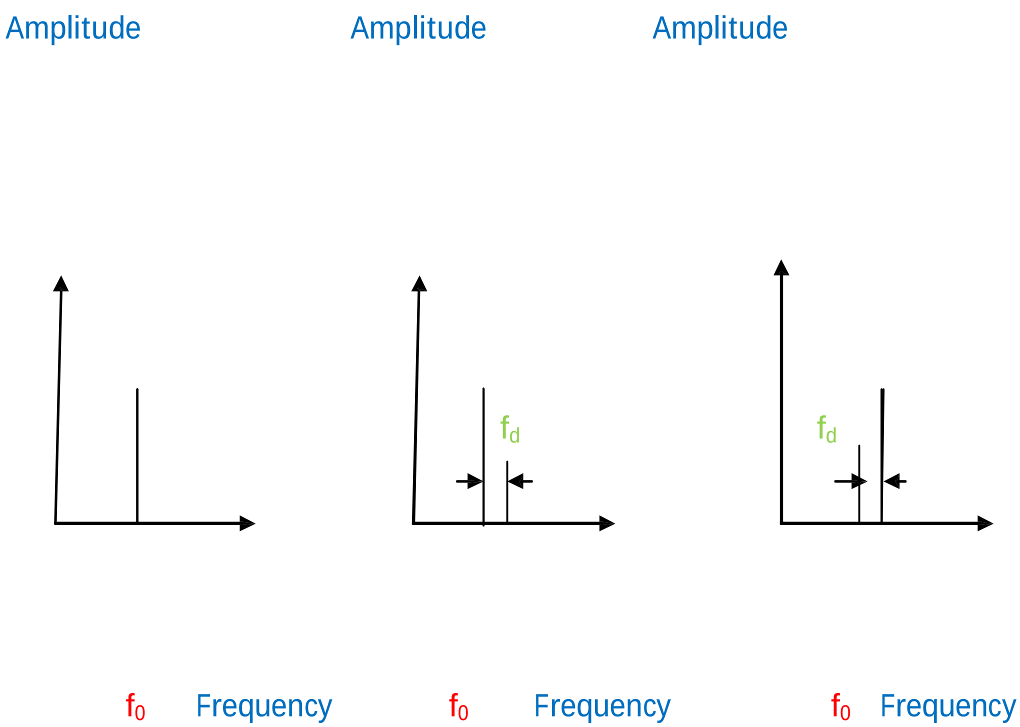
The complexity of the receiver is increased by the bank of overlapping Doppler filters whether in IF or video. The bank of Doppler filters may be replaced by a narrowband tunable filter, when the system requirements permit a time sharing of the Doppler frequency range.

Measurement of Doppler direction with CW Radar OR Sign of Radial

Velocity:

In some applications of CW Radar it is of interest to know whether the target is approaching or receding. This might be determined with separate filters located on either side of the intermediate frequency.

If the echo signal frequency lies below the carrier, the target is receding. If the echo signal frequency is greater than the carrier, the target is approaching, which is shown in fig (5).



Fig(5): Spectra of received signal. (a) No Doppler shift, no relative target motion, (b) Approaching target and (c) Receding target

Although the Doppler frequency spectrum "folds over" in the video because of the action of the detector, it is possible to determine its sign from a technique borrowed from single sideband communications.

If the transmitter signal is given by $E_t = E_0 \cos(\omega_0 t)$

The echo signal from a moving target will be $E_r = k_1 E_0 \cos[(\omega_0 \pm \omega_d)t + \phi]$

Where E_0 = amplitude of transmitted signal

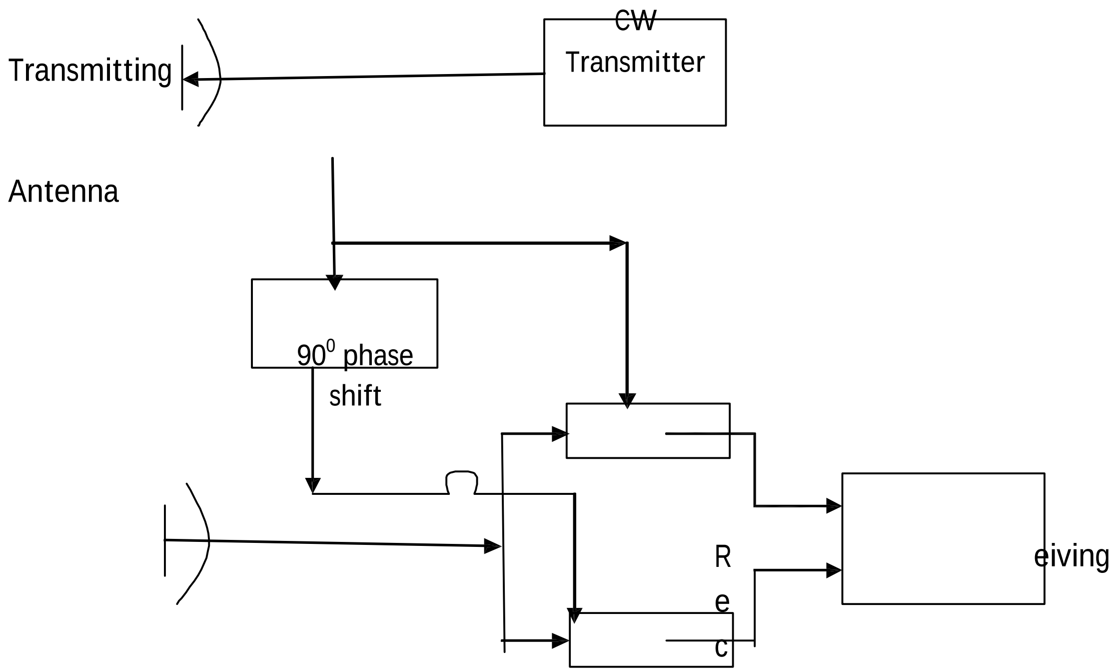
k_1 = a constant determined from the Radar equation

ω_0 = angular frequency of transmitter in rad/sec

ω_d = Doppler angular frequency shift

ϕ = a constant phase shift, which depends upon range of initial detection

The sign of the Doppler frequency and the direction of target motion may be determined by splitting the received signal into two channels as shown in fig (6).



Antenna

Mixer

Mixer B

synchronous motor

Fig (6): Measurement of Doppler direction using synchronous two-phase motor

In channel A, the signal is processed as in the simple CW Radar. A part of the transmitted signal and the received signal are heterodyne in the detector (mixer) to produce a difference signal as

$$E_A = k_2 E_0 \cos[\pm\omega_d t + \phi]$$

In channel B, except for a 90^0 phase delay introduced in the reference signal, it will work similar to that of channel A. Then the output of the channel B is given by

$$E_B = k_2 E_0 \cos[\pm\omega_d t + \phi + \pi/2]$$

If the target is approaching (positive Doppler), the outputs from the two channels are given by

$$E_A(+) = k_2 E_0 \cos[\omega_d t + \phi]$$

$$E_B(+) = k_2 E_0 \cos[\omega_d t + \phi + \pi/2]$$

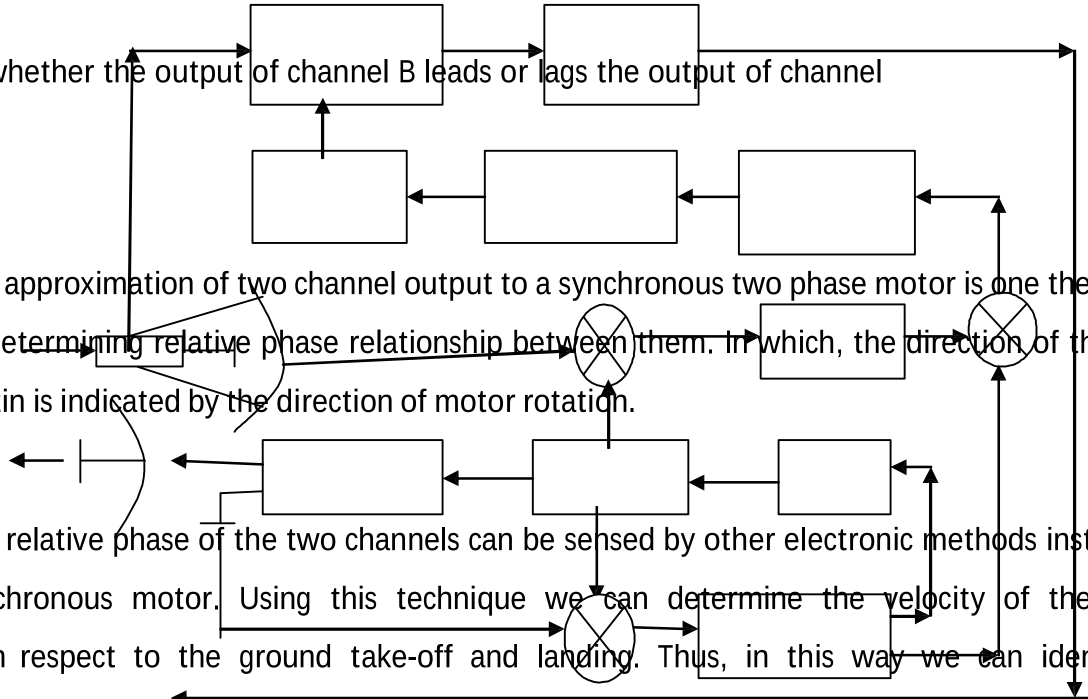
If the target is receding (negative Doppler), the outputs from the two channels are given by

$$E_A(-) = k_2 E_0 \cos[-\omega_d t + \phi] \quad \text{OR} \quad E_A(-) = k_2 E_0 \cos[\omega_d t - \phi]$$

$$E_B(-) = k_2 E_0 \cos[-\omega_d t + \phi + \pi/2] \quad \text{OR} \quad E_B(-) = k_2 E_0 \cos[\omega_d t - \phi - \pi/2]$$

The sign of Doppler frequency and the direction of target's motion may be determined according

to whether the output of channel B leads or lags the output of channel A.



The approximation of two channel output to a synchronous two phase motor is one the method of determining relative phase relationship between them. In which, the direction of the target motin is indicated by the direction of motor rotation.

The relative phase of the two channels can be sensed by other electronic methods instead of a synchronous motor. Using this technique we can determine the velocity of the aircraft with respect to the ground take-off and landing. Thus, in this way we can identify the Doppler direction with respect to CW Radar.

CWTracking Illuminator:

Phase Compensator

Servo System

Scan Detector

Speed Gate

Doppler Amplifier

XTAL

Signal IF

X+D

X

Magnetron

Klystron

AFC

X

XTAL

Reference IF

Pedestal Control

Fig (7): Block diagram of a CW tracking illuminator.

The figure shows the basic block diagram of CW tracking illuminator. It is a tracking Radar as well as illuminator. Since it must be able to follow the target as it travels through space. The operation in presence of clutter is possible, due to the Doppler discrimination of continuous wave Radar. In this type of Radar, the receiver in the missile receives the energy from the target and this energy has been transmitted to the missile by an illuminator. The illuminator may be at the launch platform. This CW illuminator has been used in many successful systems.

Speed gate is a wide band Doppler amplifier shown in fig (7), which is also referred as narrowband tracking filter. The main function of this gate is to acquire the target and track, by changing Doppler frequency shift.

Advantages of CW Doppler Radar:

1. CW Doppler Radars are not pulsed and simple to manufacture.
2. These Radars have no minimum or maximum range and maximize power on a target because they are always broad casting.
3. These are having the ability to measure velocity with extreme accuracy by means of the

   Doppler shift in the frequency echo.
4. The detected, reflected wave is shifted in frequency by an amount which is a function of the relative velocity between the target and the transmitter receiver.
5. Range data are extracted from the change in Doppler frequency.

Disadvantages of CW Doppler Radar:

1. When a single antenna is used for both transmission and reception, it is difficult to protect the receiver against the transmitter because in constant to pulse Radar, both are ON all the time.
2. These are able to detect only moving targets, as stationary targets will not cause a

   Doppler shift and the reflected signals will be filtered out.
3. CW Radars are not able to measure range, where range is normally measured by timing the delay between a pulse being sent and received but as CW Radars are always broadcasting; there is no delay to measure.

Applications of CW Doppler Radar:

1. Simple unmodulated CW Radar can be used to find the relative velocity of a moving target without any physical constant with the target. For example: in police speed monitors, in rate of climb-meter for vertical-take-off aircraft, measurement of turbine- blade vibration, the peripheral speed of grinding wheels and the monitoring of vibrations in the cables of suspension bridges.
2. CW radars are also used for the control of traffic lights, regulation of toll booths, vehicle counting.
3. In railways CW Radars can use as a speed meter to replace the conventional axle-driven tachometer.
4. In measurement of rail roadfrieght car velocity during humping operations in marshalling yards.
5. It can also be used as detection device to give track maintenance personnel advance warning of approaching trains.
6. It is also employed for monitoring the docking speed of large ships.

Pulse Vs. Continuous Wave:

Pulse Echo	Continuous Wave

1. Single Antenna 2. Gives Range, usually Alt. as well 3. Susceptible To Jamming 4. Physical Range Determined By PW and	1. Requires 2 Antennae 2. Range or Alt. Info 3. High SNR 4. More Difficult to Jam But Easily

Unambiguous Range in CW Radar:

Consider a CW Radar with the folling waveform $S(t) = A \sin(2\pi f_0 t)$

The received signal from moving target at range is $S_r(t) = A_r \sin(2\pi f_0 t - \phi)$

Where the phase $\phi = 2\pi f_0 T$

But, $R=cT/2$

$T=2R/c$ and $\lambda=c/f_0$

Therefore, $\phi = 2\pi f_0(2R/c) = 4\pi f_0 R/c= 4\pi R/\lambda$

$$R = \lambda\phi/4\pi \text{ ------- } (1)$$

Where, c is the velocity of propagation = 3×10^8

m/sec.

From the above equation we observe that, the maximum unambiguous range occurs when ϕ is maximum i.e., $\phi=2\pi$. Therefore, even for relatively large Radar wavelength R is limited to impractical small values.

Now consider a Radar with two CW signals denoted by $S_1(t)$ and $S_2(t)$ respectively.

$$S_1(t) = A_1 \sin(2\pi f_1 t) \text{ and } S_2(t) = A_2 \sin(2\pi f_2 t)$$

The received signal from moving target at range is $S_{1r}(t) = A_{r1} \sin(2\pi f_1 t-\phi_1)$

$$S_{2r}(t) = A_{r2} \sin(2\pi f_2 t-\phi_2)$$

Where $\phi_1 = 2\pi f_1(2R/c) = 4\pi f_1 R/c$ and $\phi_2 = 2\pi f_2(2R/c) = 4\pi f_2 R/c$

After mixing the carrier frequency, the phase difference between the two received signals is

$$\Delta\phi = \phi_2-\phi_1$$

$$\Delta\phi = (4\pi f_2 R/c)-(4\pi f_1 R/c) = (4\pi R/c)(f_2-f_1)$$

$$\Delta\phi = (4\pi R/c)(\Delta f) = 4\pi R\Delta f/c$$

After R is maximum when $\Delta\phi = 2\pi$

Therefore $\Delta\phi = 4\pi R\Delta f/c \Rightarrow 2\pi = 4\pi R\Delta f/c$

$$R = c/2\Delta f \text{ ----------- } (2)$$

Since, $\Delta f << c$, the range computed by eq(2) is such greater than that computed by eq(1).

Prob (1): For an ambiguous range of 81 nautical miles (1nmi=1852 meters) in a two frequency CW Radar. Determine f_2 and Δf when f_1=4.2 kHz.

Sol: Given that, unambiguous range R = 81 nmi

$$R_{unamb} = 81\times1852=150.012 \text{ km}$$

$$\text{The unambiguous range } R = c/2\Delta f = 999.92 \text{ Hz}$$

$$\Delta f = f_2 - f_1$$

$$f_2=\Delta f+f_1=5.199 \text{ kHz}$$

Prob (2): Determine the acceleration of a target if the received signal bandwidth is 40 Hz and the operating wavelength is 9 cm.

Sol: Given that for a moving target

Received signal bandwidth Δf_d=40Hz

Operating wavelength = 9cm = 9×10^{-2}m

Acceleration of a target a_r=?

We know that, $\Delta f_d = (2a_r/\lambda)^{1/2}$

a_r=72 m/sec^2

Prob (3): Determine the operating wavelength if the target is moving with acceleration as same as acceleration of gravity and the received signal bandwidth is 50 Hz.

Sol: Given that, acceleration of a moving target a_r=9.8 m/sec^2 = acceleration of gravity Received signal bandwidth Δf_d=50Hz

Operating frequency = ? and $\lambda = c/f_0$

We know that, $\Delta f_d = (2a_r/\lambda)^{1/2}$

f_0=38.27 Hz.

Prob (4): with a transmit (CW) frequency of 5GHz, calculate the Doppler frequency seen by stationary Radar when the target radial velocity is 100km/hr.

Sol: f_0=5GHz, v_r=100km/hr = 100×5/18 m/sec = 27.778 m/sec.

The Doppler frequency, $f_d = 2v_rf_0/c = 925.93$ Hz.

OBJECTIVE TYPE QUESTIONS

1. Stagger PRF is used to [ ]

a. shift the target velocities to which the MTI system is blind

b. improves the detection of a moving target against cluster

background c. increase the average power transmitted

d. increase the peak power transmitted.

2. COHO stands for [ ]

a. coherent output b. counter housed oscillator

c. coherent local oscillator d. carrier oscillator and Hartley oscillator

3. If the peak transmitted power in a radar system in increased by a factor of 16, the maximum range will be increased [ ]

a. 2 times b. 4 times c. 4 times d. 16 times

4. Which of the following statement is incorrect, The radar cross section of a target? [ ]

a. depends on the aspect of a target, if this is non

spherical. b. depends on the frequency used.

c. is equal to the actual cross sectional area for small

targets d. may be reduced by special coating of the target.

5. which of the following statement is incorrect High PRF will [ ]

a. increase the maximum range

b. make target tracking easier to distinguish from noise

c. make the returned echoes easier to distinguish from

noise d. have no effect on the range resolution.

6. Side lobe of an antenna causes [ ] a.

reduction in gain of antenna b. reduction in beam width of antenna c.

ambiguity in direction finding d. increases directivity

7. A radar which is used for determining the velocity of the moving aircraft along with its position and range is [ ]

a. moving target indicator b. radar speedometer

c. pulse radar d. radar range finder

8. Blind speed in MTI radar results in [ ]

a. restriction in speed of detectable targets b. blanking to PPI.

c. no change in phase detector output d. absorption of electromagnetic waves.

9. The quartz delay line in MTI radar is used to [ ]

a. match the signal with echo b. subtract a complete scan from previous scan

c. match the phase of COHO and STALO

d. Match the phase of COHO and output of oscillator

10. Which one of the following applications or advantages of radar beacons is false [ ]

a. navigation b. target identification c.

more accurate tracking of enemy target

d. very significant extension of the maximum range.

11. STALO stands for [ ]

a. standard local oscillator b. stable L-band output

c. stabilized local oscillator d. saturated and linear oscillator.

12. Large antenna is used in radar because it [ ]

a. gives higher gain b. gives lesser side lobes. c. increases the bean width d. increases band width

13. The range of radar is [ ]

a. directly proportional to the gain of the radar antenna

b. directly proportional to the minimum detectable signal by the receiver c. inversely proportional to the gain of the radar antenna

d. inversely proportional the transmitted power.

14. A bistatic radar has [ ]

a. One antenna for transmitting as well as for receiving

b. Two antennas for receiving the signal. c. Two antennas for transmitting signal d. transmitting and receiving antennas

15. Blind speed causes target to appear [ ]

a. moving uniformly b. moving irregularly c. stationary d. intermittently

16. For precise target location and tracking retards operate in ` [ ]

a. s- band b. D- Band c. L- Band d. X - Band

17. The sensitivity of a radar receiver is ultimately set by [ ]

a. high S/N ratio b. lower limit of useful signal input c. overall all noise temperature d. low S/N ratio

18. A radar system cannot be used [ ]

a. to detect moving objects b. to detect trajectory of moving objects

c. to detect aircraft d. to detect storms

19. Which of the following is essential for fast communication [ ]

a. Hgh S/N ratio b. High channel capacity c. large bandwidth d. Higher directivity

20. The major advantage of pulsed radar CW radar is that [ ]

a. pulsed radar readily gives the range of target while CW radar cannot give range information

b. pulsed radar can identify a target more easily than CW radar.

c. Pulses get reflected from the target more efficiently as compared to CW waves d. Pulses have variation of magnitude and frequency both

Answers:

1.a	2.c	3.a	4.c	5.a	6.a	7.c	8.a	9.b	10.c
11.c	12.a	13.b	14.d	15.c	16.a	17.c	18.d	19.a	20.a

ESSAY TYPE QUESTIONS

1. With the help of a suitable block diagram, explain the operation of CW Doppler radar in a sideband super heterodyne receiver.
2. Calculate the Doppler frequency of stationary CW radar transmitting at 6 MHz frequency when a moving target approaches the radar with a radial velocity of 100 Km/Hour.
3. List the limitations of CW radar.
4. What is Doppler frequency shift? Establish a relation between Doppler frequency shift and radial velocity of a moving target.
5. Explain how isolation between transmitter and receiver of a radar system can be achieved

if single antenna is used for transmission and
reception.

6. What is Doppler frequency shift? Discuss the effect of receiver bandwidth on the efficiency of detection and performance of a CW Doppler radar.

7. With the help of a suitable block diagram, explain the operation of a CW tracking illuminator application of a CW radar.

8. With the help of a suitable block diagram, explain the operation of a CW radar with non- zero IF in the receiver.

9. Describe methods to achieve isolation between transmitter and receiver of CW Doppler radar if same antenna is to be used for transmission and reception

10. What is the beat frequency? How it is used in FMCW radar?

11. Explain how the multipath signals produce error in FM altimeter?

12. Explain how earphones are used as an indicator in CW Radar?

13. The transmitter power is 1 KW and safe value of power which might be applied to a receiver is 10mW. Find the isolation between transmitter and receiver in dB. Suggest the appropriate isolator.

14. Why the step error and quantization errors which occur in cycle counter are used for frequency measurement in FMCW Radar?

15. What is the Doppler Effect? What are some of the ways in which it manifests itself? What are its radar applications?

16. Find the relation between bandwidth and the acceleration of the target with respect to radar?

17. How to find the target speed from Doppler frequency?

18. Write the applications of CW Radar.

19. What are the factors that limit the amount of isolation between Transmitter and Receiver of CW Radar?

20. Explain how earphones are used as an indicator in CW Radar?

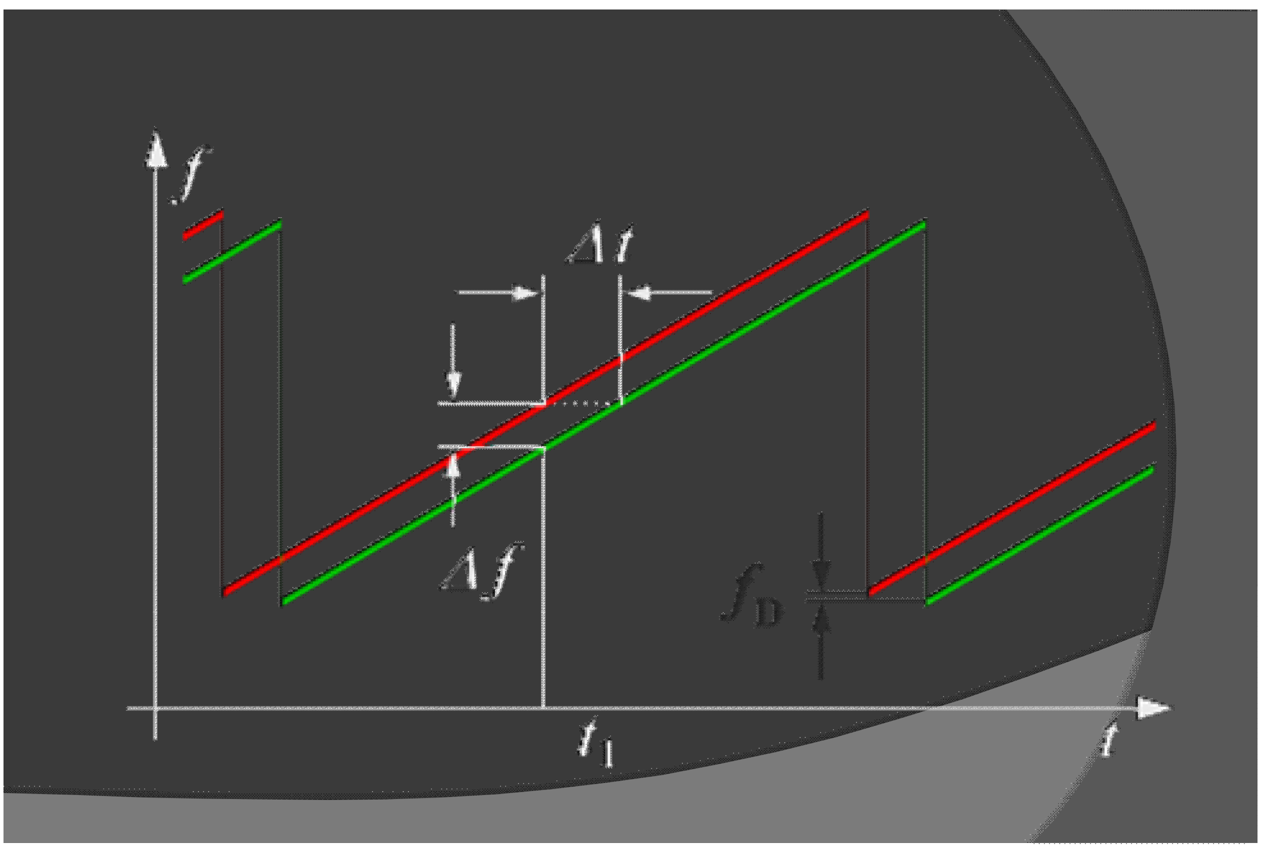
UNIT-IV

FM-CW

RADARR

Frequency-Modulated Continuous-Wave Radar:

CW radars have the disadvantage that they cannot measure distance, because it lacks the timing mark necessary to allow the system to time accurately the transmit and receive cycle and convert the measured round-trip-time into range. In order to correct for this problem, phase or frequency shifting methods can be used. In the frequency shifting method, a signal that constantly changes in frequency around a fixed reference is used to detect stationary objects and to measure the rage. In such Frequency-**M**odulated **C**ontinuous **W**ave radars (**FMCW**), the frequency is generally changed in a linear fashion, so that there is an up-and-down or a sawtooth-like alternation in frequency. If the frequency is continually changed with time, the frequency of the echo signal will differ from that transmitted and the difference Δf will be proportional to round trip time Δt and so the range R of the target too. When a reflection is received, the frequencies can be examined, and by comparing the received echo with the actual step of transmitted frequency, you can do a range calculation similar to using pulses:



$R = c_0|\Delta t|/2 = c_0|\Delta f|/(2df/dt)$

Where: c_0 = speed of light = $3 \cdot 10^8$ $^m/_s$

Δt = measured time-difference [s]

R = distance altimeter to terrain [m]

df/dt = transmitters frequency shift per unit time

Characteristic Feature of FMCW radar:

1. The distance measurement is done by comparing the actual frequency of the received signal to a given reference (usually direct the transmitted signal)
2. The duration of the transmitted signal is much larger than the time required for measuring the installed maximum range of the radar

Doppler direction in FMCW radar:

A block diagram illustrating the principle of the FM-CW radar is shown in Fig. A portion of the transmitter signal acts as the reference signal required to produce the beat frequency. It is introduced directly into the receiver via a cable or other direct connection. Ideally the isolation between transmitting and receiving antennas is made sufficiently large so as to reduce to a negligible level the transmitter leakage signal which arrives at the receiver via the coupling between antennas. The beat frequency is amplified and limited to remove any amplitude fluctuations. The frequency of the amplitude-limited beat note is measured with a cycle-counting frequency meter calibrated in distance.

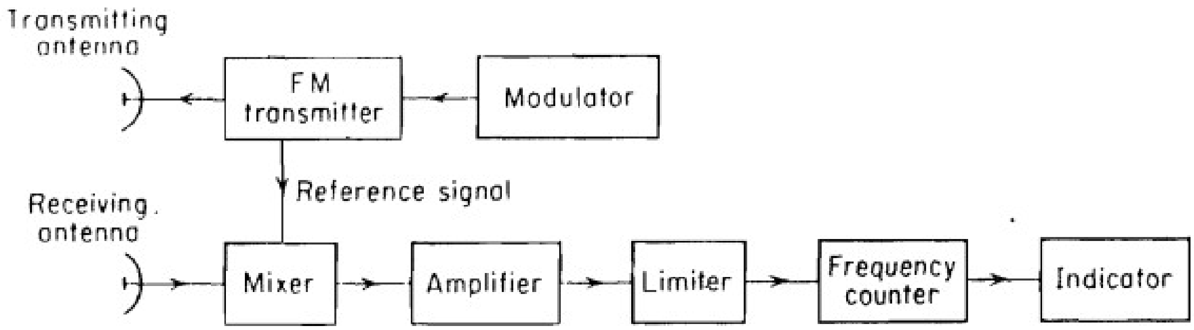


Fig: Block diagram of FM-CW radar

In the above, the target was assumed to be stationary. If this assumption is not applicable, a Doppler frequency shift will be superimposed on the FM range beat note and an erroneous range measurement results.

The Doppler frequency shift causes the frequency-time plot of the echo signal to be shifted up or down (Fig. 4.1.2 (a)). On one portion of the frequency-modulation cycle the heat frequency (Fig, 4.1.2 (b)) is increased by the Doppler shift, while on the other portion it is decreased. If for example, the target is approaching the radar, the beat frequency fb(up) produced during the increasing, or up, portion of the FM cycle will be the difference between the beat frequency due to the range from and the doppler frequency shift fd. Similarly, on the decreasing portion, the beat frequency, f_b(down) is the sum of the two.

$f_b(up) = f_r - f_d$ and $f_b(down) = f_r + f_d$

The range frequency f_r, may be extracted by measuring the average beat

frequency; That is, $f_r = 1/2[f_b(up) + f_b(down)]$.

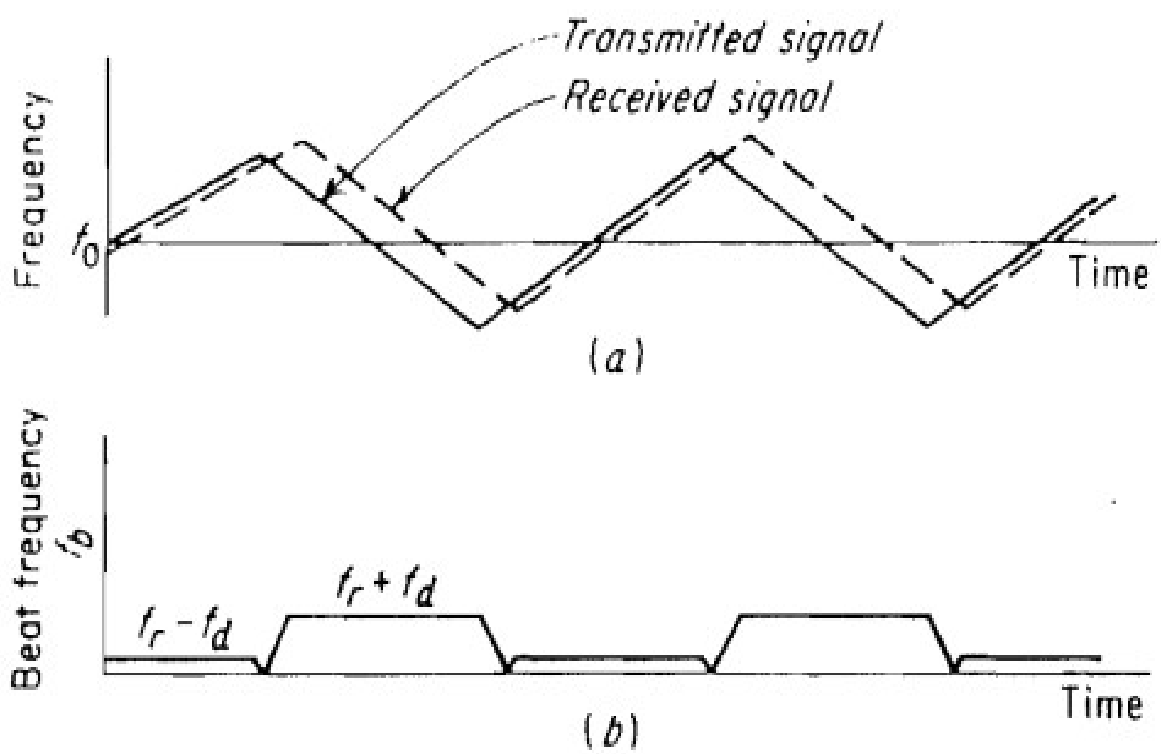


Fig: Frequency-time relation-ships in FM-CW radar when the $f_r + f_d$ received signal is shifted in frequency by the doppler effect (a) Transmitted (solid curve) and echo (dashed curve); (b) beat frequency.

If f_b(up) and f_b(down) are measured separately, for example, by switching a frequency counter every half modulation cycle, one-half the difference between the frequencies will yield the doppler frequency. This assumes $f_r > f_d$.

If, on the other hand, $f_r < f_d$ such as might occur with a high-speed target at short range,

the roles of the averaging and the difference-frequency measurements are reversed; the averaging meter will measure Doppler velocity, and the difference meter, range. If it is not known that the roles of the meters are reversed because of a change in the inequality sign between f_r and f_d an incorrect interpretation of the measurements may result.

Derive an expression for range and Doppler measurement for FMCW radar:

In the frequency-modulated CW radar (abbreviated as FM-CW), the transmitter frequency is changed as a function of time in a known manner. Assume that the transmitter frequency increases linearly with time, as shown by the solid line in Fig (a).

If there is a reflecting object at a distance R, an echo signal will return after a time T =

2R/c. The dashed line in the figure represents the echo signal. If the echo signal is heterodyned with a portion of the transmitter signal in a nonlinear element such as a diode, a beat note f_b will be produced.

If there is no Doppler frequency shift, the beat note (difference frequency) is a measure of the target's range and $f_b = f_r$ where f_r is the beat frequency due only to the target's range. If the rate of change of the carrier frequency is f_0, the beat frequency is

$$f_r = f_0 T = 2Rf_0/c$$

In any practical CW radar, the frequency cannot be continually changed in one direction only. Periodicity in the modulation is necessary, as in the triangular frequency-modulation waveform shown in Fig(b).

The modulation need not necessarily be triangular; it can be sawtooth, sinusoidal, or some other shape. The resulting beat frequency as a function of time is shown in Fig(c) for triangular modulation. The beat note is of constant frequency except at the turn around region. If the frequency is modulated at a rate f_m over a range Δf, the beat frequency is

$$f_r = 2*2Rf_m/c = 4Rf_m\Delta f /c$$

Thus the measurement of the beat frequency determines the range R.

$$R = cf_r/4f_m\Delta f$$

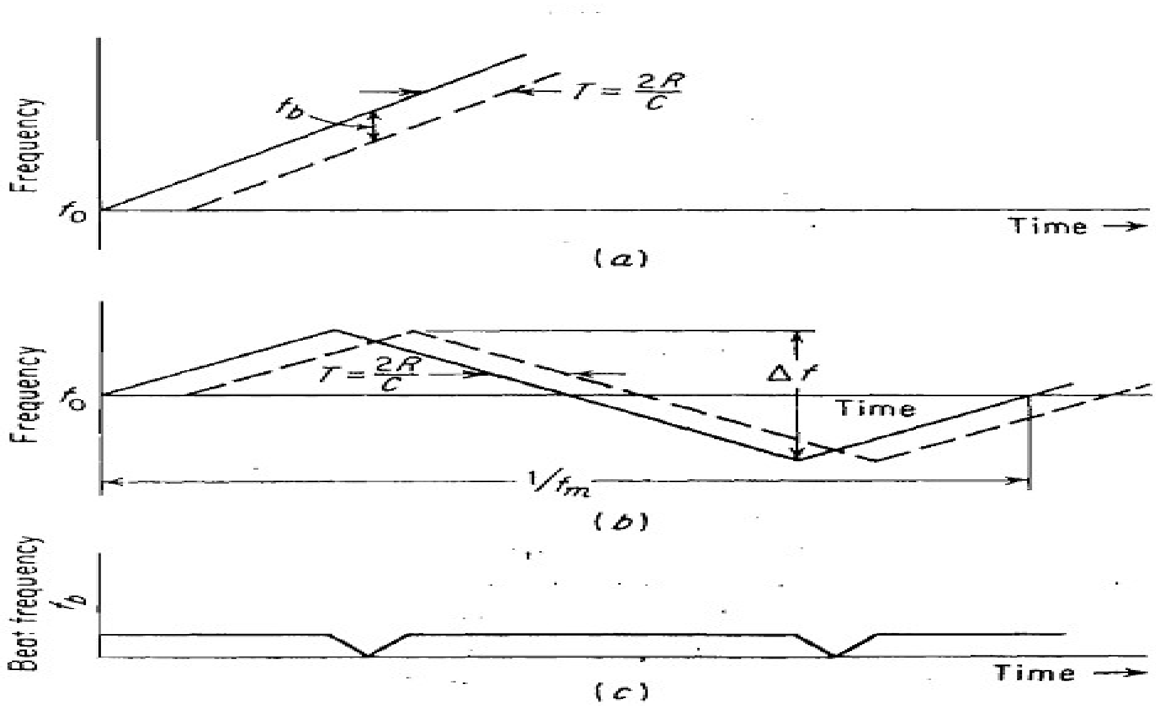
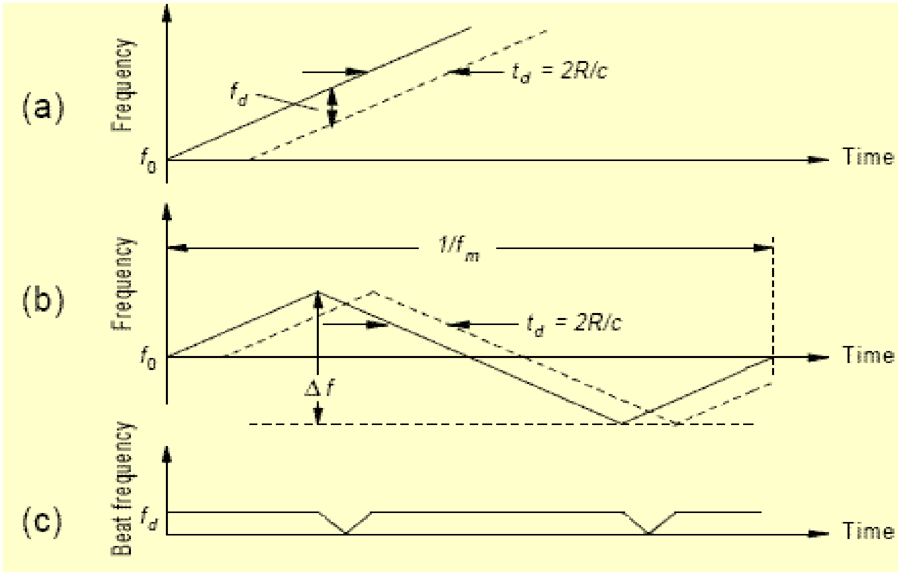


Fig: Frequency-time relationships in FM-CW radar. Solid curve represents transmitted signal, dashed curve represents echo. (a) Linear frequency modulation; (b) triangular frequency modulation; (c) beat note of (b).



Principle of operation of FMCW Altimeter:

The FM-CW radar principle is used in the aircraft radio altimeter to measure height above the surface of the earth. The large backscatter cross section and the relatively short ranges required of altimeters permit low transmitter power and low antenna gain. Since the relative motion between the aircraft and ground is small, the effect of the Doppler frequency shift may usually be neglected.

The band from 4.2 to 4.4 G Hz is reserved for radio altimeters, although they have in the past operated at UHF. The transmitter power is relatively low and can be obtained from a CW magnetron, a backward-wave oscillator, or a reflex klystron, but these have been replaced by the solid state transmitter.

The altimeter can employ a simple homodyne receiver, but for better sensitivity and stability the superheterodyne is to be preferred whenever its more complex construction can be tolerated.

A block diagram of the FM-CW radar with a sideband superheterodyne receiver shown in Fig. A portion of the frequency-modulated transmitted signal is applied to a mixer along with the oscillator signal. The selection of the local-oscillator freq uency is a bit different from that in the usual superheterodyne receiver.

The local-oscillator frequency f_{IF} should be the same as the intermediate frequency used in the receiver, whereas in the conventional superheterodyne the LO frequency is of the same order of magnitude as the RF signal.

The output of the mixer consists of the varying transmitter frequency $f_o(t)$ plus two sideband frequencies, one on either side of $f_o(t)$ and separated from $f_o(t)$ by the local-oscillator frequency f_{IF}. The filter selects the lower sideband $f_o(t) - f_{IF}$ and rejects the carrier and the upper sideband.

The sideband that is passed by the filter is modulated in the same fashion as the transmitted signal. The sideband filter must have sufficient bandwidth to pass the modulation, but not the carrier or other sideband. The filtered sideband serves the function of the local oscillator.

When an echo signal is present, the output of the receiver mixer is an IF signal of frequency $f_{IF} + f_b$ where f_b is composed of the range frequency f_r and the doppler velocity frequency f_d. The IF signals is amplified and applied to the balanced detector along with the localoscillator signal f_{IF}. The output of the detector contains the beat frequency (range frequency and the Doppler velocity frequency), which is amplified to a level where it can actuate the frequencymeasuring circuits.

In Fig. the output of the low-frequency amplifier is divided into two channels: one feeds an average-frequency counter to determine range, the other feeds a switched frequency counter to determine the doppler velocity (assuming fr > fd) Only the averaging frequency counter need be used in an altimeter application, since the rate of change of altitude is usually small.

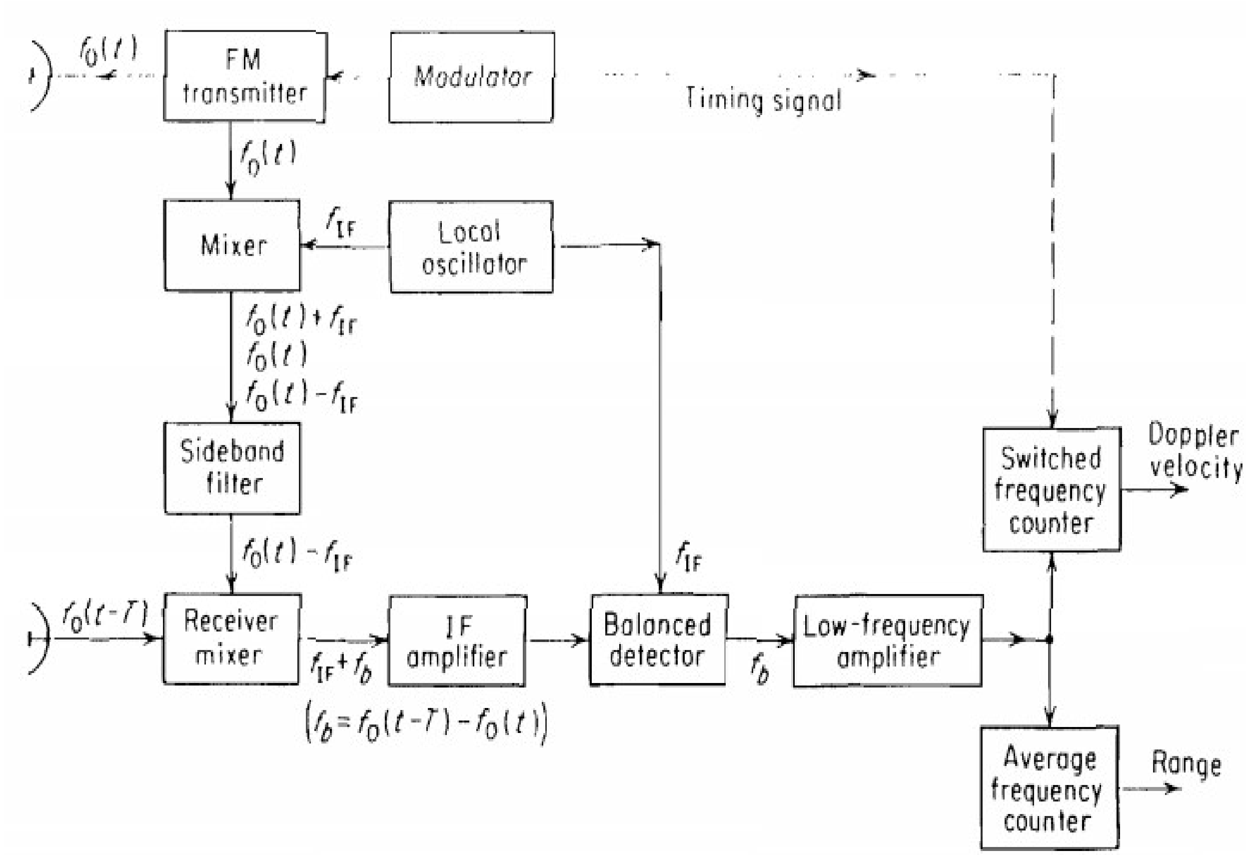


Fig: Block diagram of FM-CW radar using sideband superheterodyne receiver

EFFECT OF NOISE SIGNALS ON FM ALTIMETER:

The different noise signals occurring in a typical FM altimeter are:

- Due to the mismatch in impedance a part transmitted signal gets reflected from the space causing error in the altimeter.
- The mismatch between the sideband filter and receiving gives rise to standing wave pattern.

- The leakage signal due to the transmitting and receiving antennas reach the receiver and cause error..
- The double bounce signal.

Hence the different noise signals accompanying the transmitted signal may reach the receiver and effect its.

Advantages of FMCW altimeter over pulse based altimeter and compare both?

- What is the difference between altimeter and cabin altimeter: The main difference between altimeter and cabin altimeter is the place where they take their pressure: Altimeter takes the pressure from static ports, while cabin altimeter takes it's pressure from the cabin.
- Difference of radio altimeter to radar altimeter**:** They're both the same thing
- What is an altimeter: **T**he altimeter is basically a specialized pressure gauge. It measures the pressure of the column of air above it. As the altitude varies, the air column height varies, which registers on the altimeter. Since the air pressure also varies with changes in the barometric pressure, altimeters must have an adjustment to compensate for changes in local barometric pressure.

Sinusoidally modulated FM-CW radar:

The block diagram for sinusoidally modulated FM-CW radar extracting the third harmonic is shown in fig.

The ability of the FM-CW radar to measure range provides an additional basis for obtaining isolation. Echoes from short-range targets-including the leakage signal may be attenuated relative to the desired target echo from longer ranges by properly processing the difference frequency signal obtained by heterodyning the transmitted and received signals.

If the CW carrier is frequency-modulated by a sine wave, the difference frequency obtained by heterodyning the returned signal with a portion of the transmitter signal may be expanded in a trigonometric series whose terms are the harmonics of the modulating frequency f_m.

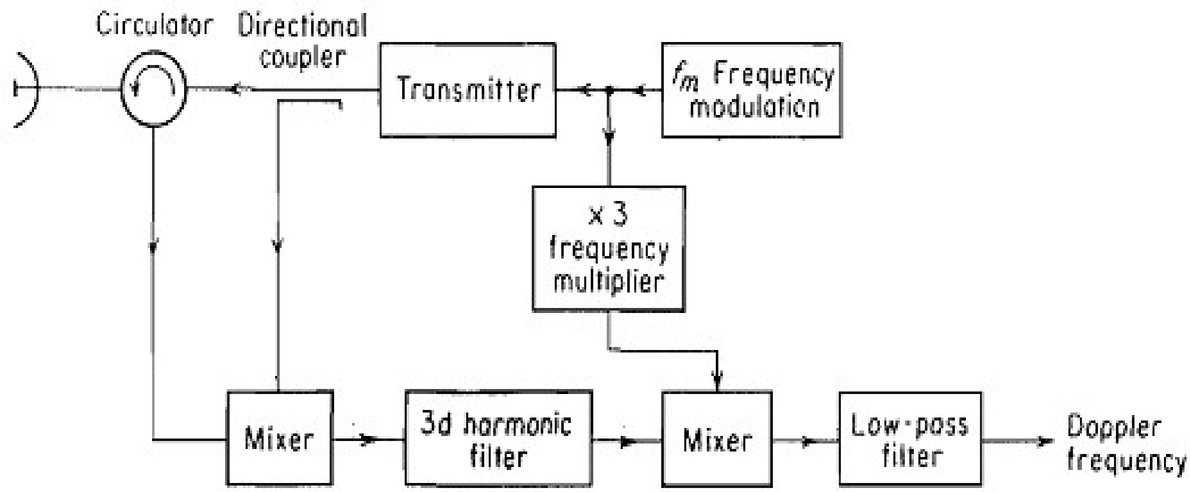


Fig: sinusoidally modulated FM-CW radar extracting the third harmonic.

Assume the form of the transmitted signal to be

Where,

f_o = carrier frequency, f_m = modulation frequency

Δf = frequency excursion (equal to twice the frequency derivation)

The difference frequency signal may be written as

$$
\begin{aligned}
v_D = {} & J_0(D)\cos(2\pi f_d t - \phi_0) + 2J_1(D)\sin(2\pi f_d t - \phi_0)\cos(2\pi f_m t - \phi_m) \\
& - 2J_2(D)\cos(2\pi f_d t - \phi_0)\cos 2(2\pi f_m t - \phi_m) \\
& - 2J_3(D)\sin(2\pi f_d t - \phi_0)\cos 3(2\pi f_m t - \phi_m) \\
& + 2J_4(D)\cos(2\pi f_d t - \phi_0)\cos 4(2\pi f_m t - \phi_m) + 2J_5(D)\cdots
\end{aligned}
$$

Where J_0, J_1, J_2, etc = Bessel functions of first kind and order 0, 1, 2, etc., respectively

$$D = (\Delta f/f_m) \sin (2\pi f_m R_0/c)$$

R_0 = distance to target at time t = 0 (distance that would have been measured if target were stationary)

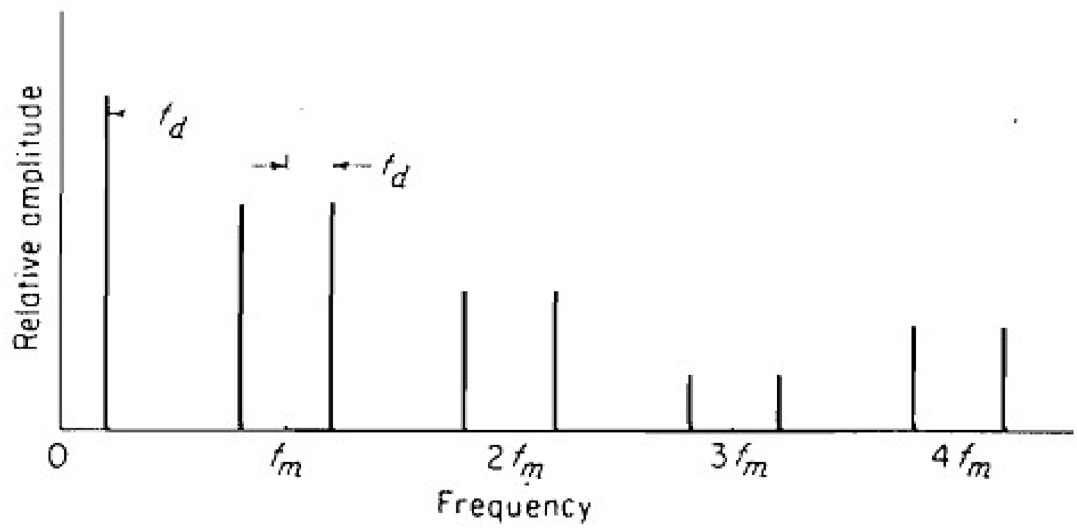
c = velocity of propagation, f_d = doppler frequency shift

v_r = relative velocity of target with respect to radar

ϕ_0 = phase shift approxirnately equal to angular distance $2\pi f_0 R_0/c$

ϕ_m = phase shift approxirnately equal to $2\pi f_m R_0/c$

The difference-frequency signal consists of a doppler-frequency component of amplitude $J_o(D)$ and a series of cosine waves of frequency f_m, $2f_m$, etc. Each of these harmonics of f_m is modulated by a doppler-frequency component with amplitude proportional to $J_n(D)$. The product of the doppler-frequency factor times the nth harmonic factor is equivalent to a suppressed- carrier double-sideband modulation.



The above figure shows a plot of several of the Bessel functions. The argument D of the Bessel function is proportional to range. The Jo amplitude applies maximu response to signals at zero range in radar that extracts the d-c doppler-frequency component. This is the range at which the leakage signal and its noise conlponents (including microphony and vibration) are found. At greater ranges, where the target is expected, the effect of the Jo Bessel function is to reduce the echo-signal amplitude in comparison with the echo at zero range (in addition to the normal range attenuation). Therefore, if the Jo term were used, it would enhance the leakage signal and reduce the target signal.

BANK OF NARROW BAND FILTERS IN FMCW RADAR:

The mixer output of FMCW Radar will contain more than one difference frequency, when more than one target is present within the view of Radar. In a linear system, each target is associated with a frequency component.

By measuring the individual frequency component, we can determine the range to each target and the following equation is applied to each frequency component

$$\text{i.e., } f_r = 4Rf_m\Delta f / c$$

The above frequency component must be separated from one another to measure the individual frequencies. Thus, the requirement of a single frequency corresponding to a single target can be accomplished and continuously observed by a bank of narrow band filters.

But for the following cases, the problem of resolving targets and measuring the range of each becomes more complicated.

- ❖ If the motion of the targets were to produce a Doppler frequency shift.
- ❖ If the frequency modulation waveforms were non linear.
- ❖ If the mixers were not operated in its linear region.

Multiple frequency CW radar:

The multiple frequency CW radar is used to measure the accurate range.

The transmitted waveform is assumed to consist of two continuous sine waves of frequency f_1 and f_2 separated by an amount Δf. Let the amplitudes of all signals are equal to unity. The voltage waveforms of the two components of the transmitted signal v_{1r} and v_{2r}, may be written as

$$v_{1r} = \sin(2\pi f_1 t + \phi_1)$$

$$v_{2r} = \sin(2\pi f_2 t + \phi_2)$$

Where ϕ_1 and ϕ_2 are arbitrary (constant) phase angles.

The echo signal is shifted in frequency by the Doppler Effect. The form of the dopplershifted signals at each of the two frequencies f_1 and f_2 may be written as

$$v_{1R} = \sin\left[2\pi(f_1 \pm f_{d1})t - \frac{4\pi f_1 R_0}{c} + \phi_1\right]$$

$$v_{2R} = \sin\left[2\pi(f_2 \pm f_{d2})t - \frac{4\pi f_2 R_0}{c} + \phi_2\right]$$

Where, R_o = range to target at a particular time $t = t_0$ (range that would be measured if target were not moving)

f_{d1} = Doppler frequency shift associated with frequency f_1

f_{d2} = Doppler frequency shift associated with frequency f_2

Since the two RF frequencies f_1, and f_2 are approximately the same the doppler frequency shifts f_{d1} and f_{d2} are approximately equal to one another. Therefore, $f_{d1} = f_{d2} = f_d$

The receiver separates the two components of the echo signal and heterodynes each received signal component with the corresponding transmitted waveform and extracts the two doppler-frequency components given below:

$$v_{1D} = \sin\left(\pm 2\pi f_d t - \frac{4\pi f_1 R_0}{c}\right)$$

$$v_{2D} = \sin\left(\pm 2\pi f_d t - \frac{4\pi f_2 R_0}{c}\right)$$

The phase difference between these two components is

$$\Delta\phi = \frac{4\pi(f_2 - f_1)R_0}{c} = \frac{4\pi \ \Delta f R_0}{c}$$

Hence

$$R_0 = \frac{c \ \Delta\phi}{4\pi \ \Delta f}$$

A large difference in frequency between the two transmitted signals improves the accuracy of the range measurement since large Δf means a proportionately large change in $\Delta\phi$ for a given range. However, there is a limit to the value of Δf, since $\Delta\phi$ cannot be greater than 2π radians if the range is to remain unambiguous. The maximum unambiguous range R_{unamb} is $\mathbf{R_{unamb} = c/2\Delta f}$

The two-frequency CW radar is essentially single target radar since only one phase difference can be measured at a time. If more than one target is present, the echo signal becomes complicated and the meaning of the phase measurement is doubtful.

Measurement Errors:

The absolute accuracy of radar altimeters is usually of more importance at low altitudes than at high altitudes. Errors of a few meters might not be of significance when cruising at altitudes of 10 km, but are important if the altimeter is part of a blind landing system.

The theoretical accuracy with which distance can be measured depends upon the bandwidth of the transmitted signal and the ratio of signal energy to noise energy. In addition, measurement accuracy might be limited by such practical restrictions as the accuracy of the frequency-measuring device, the residual path-length error caused by the circuits and transmission lines, errors caused by multiple reflections and transmitter leakage, and the frequency error due to the turn-around of the frequency modulation.

A common form of frequency-measuring device is the cycle counter, which measures the number of cycles or half cycles of the beat during the modulation period. The total cycle count is a discrete- number since the counter is unable to measure fractions of a cycle. The discreteness of the frequency measurement gives rise to an error called the fixed error, or step error. It has also been called the quantization error, a more descriptive name. The average number of cycles N of the beat frequency fb in one period of the modulation cycle *fm* is fb /*fm* , where the bar over, denotes time average.

$R = cN/4\Delta f$

Where, R = range (altitude). m

c = velocity of propagation. m/s

Δf = frequency excursion. Hz

Since the output of the frequency counter N is an integer, the range will be an integral multiple of $c/4\Delta f$ and will give rise to a quantization error equal to

$\delta R = c/4\Delta f$

δR (m) = $75/\Delta f$ (MHz)

Since the fixed error is due to the discrete nature of the frequency counter, its effects can be reduced by wobbling the modulation frequency or the phase of the transmitter output. Wobbling the transmitter phase results in a wobbling of the phase of the beat signal so that an average reading of the cycle counter somewhere between N and N + 1 will be obtained on a normal meter movement. In one altimeter, the modulation frequency was varied at a 10-Hz rate, causing the phase shift of the beat signal to vary cyclically with time. The indicating system was designed so that it did not respond to the l0-Hz modulation directly, but it caused the fixed error to be averaged. Normal fluctuations in aircraft altitude due to uneven terrain, waves on the water, or turbulent air can also average out the fixed error provided the time constant of the indicating device is large compared with the time between fluctuations. Over smooth terrain, such as an airport runway, the fixed error might not be averaged out.

Target motion can cause an error in range equal to vr T0, where vr, is the relative velocity and To is the observation time. The residual path error is the error caused by delays in the circuitry and transmission lines.

Multipath signals also produce error. Reflections from the landing gear can also cause errors.

The unwanted signals in FM altimeter:

The fig. shows some of the unwanted signals that might occur in the FM altimeter. The wanted signals are shown by the solid line while unwanted signals are shown by the broken arrows.

The unwanted signals include:

1. The reflection of the transmitted signals at the antenna caused by impedance mismatch.

2. The standing-wave pattern on the cable feeding the reference signal to the receiver, due to poor mixer match.

3. The leakage signal entering the receiver via coupling between transmitter and receiver antennas. This can limit the ultimate receiver sensitivity, especially at high altitudes.

4. The interference due to power being reflected back to the transmitter, causing a change in the impedance seen by the transmitter. This is usually important only at low altitudes. It can be reduced by an attenuator introduced in the transmission line at low altitude or by a directional coupler or an isolator.

5. The double-bounce signal.

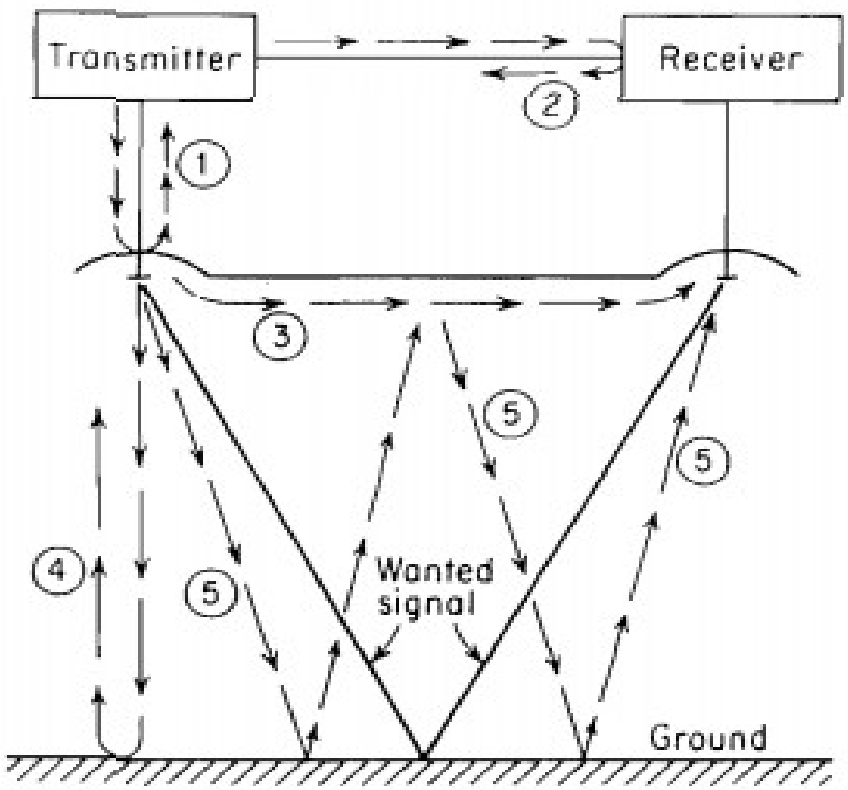


Fig: unwanted signals in FM altimeter

OBJECTIVE TYPE QUESTIONS

1. Which of the following statement is incorrect? The Doppler Effect is used in [ ]

a. MTI system b. CW radar c. FM Radar d. moving target plotting on the PPI

2. A transponder comprises of [ ]

a. transmitter only b. receiver only

c. transmitter and receiver d. transmitter and receiver and antenna

3. If A be the capture area of receiving antenna of radar then the maximum range will be

Proportional to [ ]

a. A b. A 2 c. 1 / sqrt A d. sqrt A

4. For high power radars the peak power is limited by [

] a. breakdown of RF components b. increased complexity of

transmitter circuitry c. cost aspects d. max. band width

5. The maximum range of radar depends on [ ]

a. pulse frequency b. pulse duration c. pulse energy d. pulse interrals

6. The term rat race in radar is associated with [ ]

a. duplexer. b. receiver bandwidth c. modulator d. hybrid ring

7. Most of the aircraft surveillance radars operate in [ ]

a. L band b. c band c. s band d. x band.

8. The minimum range of detection by a pulse radar depends on [ ]

a. pulse width b. Average transmitter power c. beam width of the antenna d. bandwidth of antenna

9. An MTI system eliminates permanent echoes while preserving echoes from a moving target by
[ ]

a. decreasing pulse width b. utilizing the Doppler

effect c. increasing peak transmitted power d. wide beam

width

10. A CW radar cannot give information about [
]

a. a range b. direction c. both range and

direction d. range, direction and past track

11. A target is moving with a velocity of 360km/hour radially towards the transmitting frequency generator of 3 GHz will be [
]

a. 300 KHz b. 1 KHz c. 1.5 KHz d. 2 KHz

12. A duplex is a device which [

] a. switches an antenna between transmitter and receiver by means of

selective filters b. switches an antenna between transmitter and receiver

by means of gas

Switching tubes.

c. neither of the above two d. connect at a time to transmitter or receiver only

13. The minimum receivable signal in a radar receiver who's IF bandwidth is 1.5 MHz and

which has a noise figure 9 dB will be [ ]

a. $4.16 X 10^{-10}$ Watt b. $4.16 X 10^{-12}$ Watt c. 4.16 X 10-13 Watt d. $4.16 X 10^{-14-}$ Watt

14. When p is the peak transmitted pulse power. The maximum range of the radar is proportional to [ ]

a. P b. $P^{1/2}$ c. $P^{1/4}$ d. $1/ P^{1/4}$

15. A RADAR SYSTEM OPERATES at 3 cm with a peak pulse power of 500KW. Its minimum receivable power is 10 -3 W, The capture area of th antenna is 5 m2 and the radar crosssectional area of the target is 20m 2 . the maximum range of the radar will be [ ]

a. 343 Km b. 44km c. 686 km d. 888 km

16. Which of the following is not a display method [ ]

a. PPI b. Computer feeding

c. Mono pulse conical scanning d. A scope

17. Which of the following is used for IFF [ ]

a. CW radar b. MTI c. Ordinary radar` d. Beacon

18. Which of the following statement about radar is valid [ ]

a. echoes from the target are random and noise impulses

are repetitive b. echoes and noise impulse both are

random

c. echoes and noise pulses both are repetitive

d. echoes from target are repetitive and noise impulses are random

19. A radar consists of [ ]

a. display unit b. switching modulator c. varactor diode d. gunn diode

20. In radar as soon as the transmitted pulse terminates, the transmitter is disconnected from the antenna by [ ]

a. Duplexer b. Mixer c. ART switches d. Detector

Answers:

1.d	2.d	3.d	4.c	5.c	6.a	7.a	8.a	9.a	10.a
11.d	12.b	13.d	14.c	15.c	16.c	17.d	18.d	19.a	20.a

ESSAY TYPE QUESTIONS

1. With the help of suitable block diagram, explain the operation of a FM-CW altimeter.
2. Discuss all the possible errors in the measurement accuracy of altitudes using a FM-CW

    a. radar.
3. List out the possible errors for measurement of altitudes accurately using a FM-CW altimeter.
4. Discuss the results of multiple frequency usage for operating FM-CW radar while mentioning the limitations of multiple frequency usage in CW radars.
5. Range and Doppler measurement of a target using a FM-CW radar.
6. Unwanted signals and the measurement errors in FM altimeter.
7. With necessary mathematical expressions, describe range and Doppler measurement if the transmitted signal of a CW radar is frequency modulated.
8. Describe the e ect of sinusoidal modulating signal in the place of rectangular pulses on the performance of a radar.
9. Draw the block diagram of IF Doppler bank and explain the operation of it with the help of frequency response of it.
10. What are the e ects which limit the amount of transmitter leakage power which can be tolerated at the receiver?
11. Why is amplitude comparison mono pulse more likely to be preferred over the phase comparison mono pulse and conical scan tracker over sequential lobbing, or lobe switching, tracker? Explain.

12. Discuss in detail about the Amplitude fluctuations and how its effects are minimized

13. Explain Mono pulse tracking in two angle coordinates.

14. Draw the block diagram of sinusoidally modulated FMCW radar and explain the function of each block.

15. What are the various unwanted signals which cause errors in FM altimeter?

16. Explain the two frequency CW technique for measuring the Radar range?

17. Explain the operation of the two frequency CW Radar.

18. How to select the difference between two transmitted signals of CW radar?

19. Why the step error and quantization errors which occur in cycle counter are used for frequency measurement in FMCW Radar?

20. Draw the block diagram of sinu soidally modulated FMCW radar and explain the function of each block.

UNIT-V

MTI AND PULSE DOPPLER RADAR

MTI Radar means moving target indication radar. This is one form of pulsed radar. MTI radar is characterized by its very low PRF and nence no range ambiguity in MTI Rdar. The unambigues range is given by $R_{unamb} = v_o/f_p$

Where v_o = velocity of electromagnetic wave in free space

f_p = pulse repetition frequency in Hz

Moving target indication is the process of rejecting fixed or slowly moving clutter while passing echoes from targets moving at significant velocities.

Moving Target Indication (MTI) radar: A delay line canceller filter to isolate moving targets from nonmoving background

i. Ambiguous velocity ii. Unambiguous range

Pulsed Doppler radar: Doppler data are extracted by the use of range gates and Doppler filters. i. Unambiguous velocity ii. Unambiguous or ambiguous range

Types of MTI Radar:

1. Area MTI Radar
2. Coherent MTI Radar
3. Non-coherent MTI Radar
4. Airborne Moving Target Indicator (AMTI) Rdar
5. Digital MTI (DMTI) Radar.

INTRODUCTION:

The Doppler shift produced by a moving target may be used in pulse radar: (1) To determine the relative velocity of the target or (2) To separate desired moving targets from undesired stationary clutter.

The second application has been of greater importance.

The MTI radar usually operates with ambiguous Doppler measurements (blind speeds) but with unambiguous range (no second time around echoes).

The pulse Doppler radar has a high enough PRF to operate with unambiguous Doppler, but at the expense of range ambiguities.

MTI is a necessity in high quality air surveillance radars that operate in the presence of clutter. MTI adds cost and complexity and digital signal processing.

Practical, economical MTI has been available only since the mid 1970s.

Operation: Simple CW radar is shown. In principle the CW radar can be converted into pulse radar by providing a power amplifier and a modulator to turn the amplifier on and off.

Note: The main difference between this pulse radar and the one described previouslyis that some portion of the CW power that generates the transmitted pulse is applied to the receiver as the local oscillator.

This LO provides the coherent reference needed for Doppler frequency detection. Coherent means that the phase of the transmitted signal is preserved in the reference signal.

Let the CW oscillator voltage be $V = A_1 \sin 2\pi f_t t$

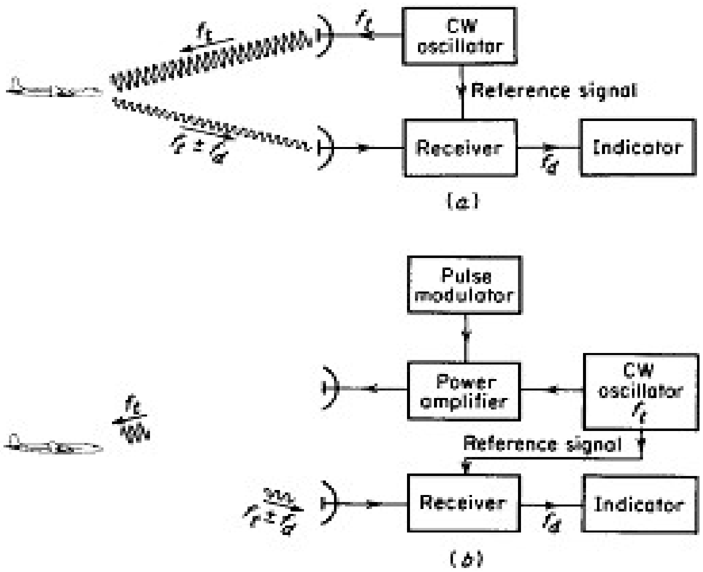


Figure 4.1 (*a*) Simple CW radar; (*b*) pulse radar using doppler information.

Therefore the reference voltage is $V_{ref} = A_2 \sin 2\pi f_t t$ ....(1)

The Doppler shifted echo voltage is $V_{echo} = A_3 \sin\left[2\pi(f_t \pm f_d)t - \frac{4\pi f_t R_0}{c}\right]$ ....(2)

Here c = velocity of propagation and R_0 = range of target

The reference and echo signal are mixed in the receiver and the difference extracted

$$V_{diff} = A_4 \sin\left[2\pi f_d t - \frac{4\pi f_t R_0}{c}\right]$$ ....(3)

Note: equations 2 and 3 represent carriers upon which the pulse modulation is imposed

Note: For stationary targets, $f_d = 0$ therefore V_{diff} will have a constant value some where between $-A_4$ and A_4 including zero.

Examples of V_{diff} are shown in figure 4.2

When $f_d > 1/\tau$ the doppler can be discerned from the information in a single pulse (4.2 b)

When $f_d < 1/\tau$ the pulses will be modulated an amplitude given by equation 3 (Fig 4.2.c)

In this second case, many pulses will be needed to extract the Doppler information.

The situation shown in Fig. 4.2 c is typical of aircraft detection radar. The situation shown in Fig. 4.2 b is typical of detection of missiles or satellites (i.e. high velocity). Doppler ambiguities can occur in Fig. 4.2 c but not in 4.2 b.

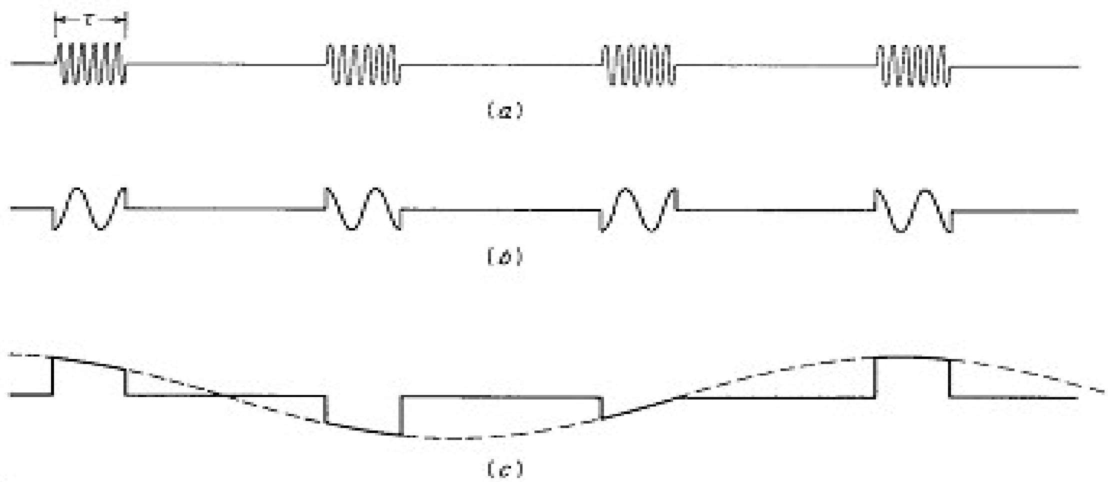


Figure 4.2 (*a*) RF echo pulse train; (*b*) video pulse train for doppler frequency $f_d > 1/\tau$; (*c*) video pulse train for doppler freuqncy $f_d < 1/\tau$.

Moving targets can be distinguished from stationary targets by observing the video on an A scope (amplitude vs range). A single sweep might appear as in Fig. 4.3 a.

This also shows several fixed targets and 2 moving targets indicated by the two arrows. From the single sweep it is impossible to distinguish moving targets from fixed targets. Successive A scope sweeps (PRIs) are shown. Echoes from fixed targets are constant while echoes from moving targets vary in amplitude at a rate corresponding to the Doppler shift. Superposition of the sweeps shows the moving targets producing a butterfly effect.

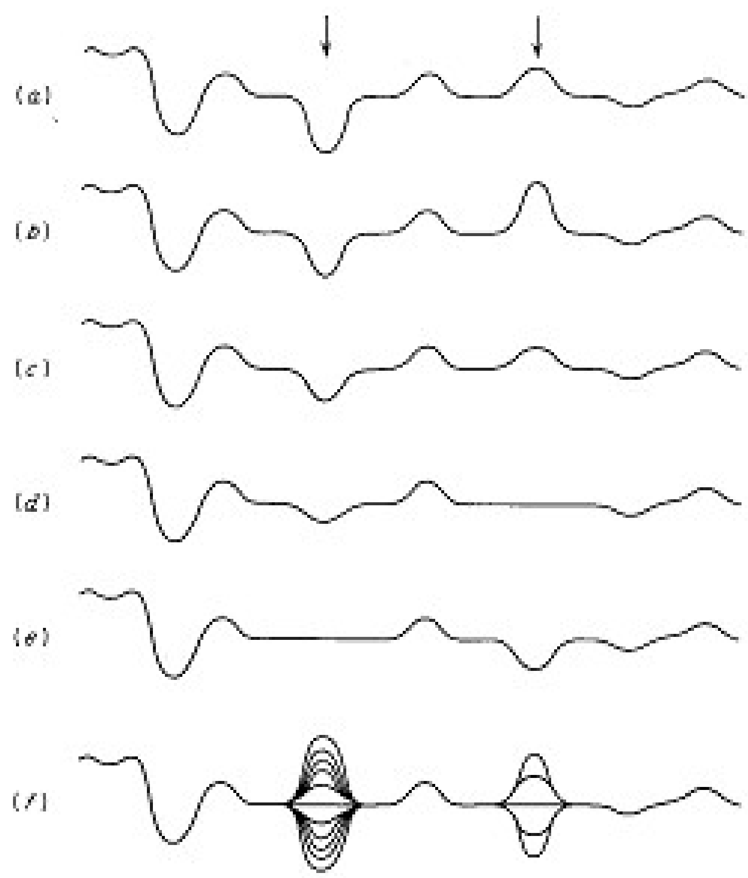
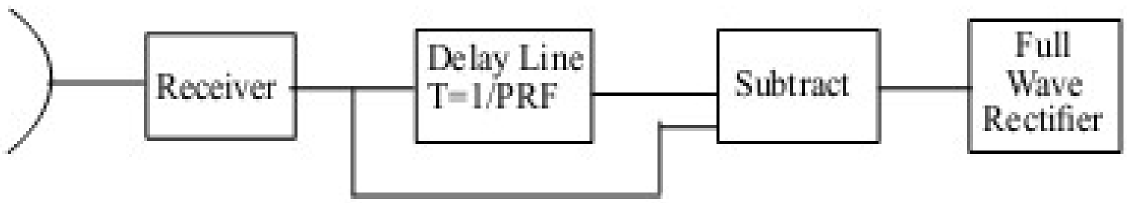


Figure 4.3 (*a–e*) Successive sweeps of an MTI radar A-scope display (echo amplitude as a function of time); (*f*) superposition of many sweeps; arrows indicate position of moving targets.

Note: the video is bipolar. Fixed targets can give either a positive or a negative pulse (corresponding to the previous comment on the value ranging from $-A_4$ and A_4.

The butterfly effect is not suitable for display on PPI. To extract Doppler information in a form suitable for a PPI, one can use a delay line canceller.



The delay line canceller acts as a filter to eliminate DC. Fixed targets with unchanging amplitudes pulse to pulse are cancelled. The amplitudes of the moving targets are not constant pulse to pulse and subtraction results in an uncancelled residue. The rectifier provides the video for the PPI.

A Diagram of an MTI employing an HPA (high power amplifier) is shown below. This arrangement is called a MOPA (master oscillator, power amplifier). Here the coherent reference is supplied by a coho (coherent oscillator). The coho is a stable oscillator whose frequency is the same as the IF frequency in the receiver. The coho f_c also mixes with the stalo (stable oscillator) $f_{l.}$ The RF echo is heterodyned with fl to produce the IF.

After amplification at IF the received signal is phase detected with fc. (Note: the heterodyne process is used to avoid 1/f noise) to give video proportional to the phase difference between the two signals. The electronics excluding the high power amplifier are collectively called the exciter-receiver.

Note: although the phase of the STALO influences the phase of the transmitted signal, any STALO phase is cancelled on reception.

Again, the main feature of MTI is that the transmitted signal must be coherent (phase referenced) with the downconverting oscillators in the receiver. The HPA maybe a Klystron, a TWT (travelling wave tube), crossed field amplifier, triode, tetrode. Each has its advantages and disadvantages.

Before the development of the Klystron amplifier, the only high power transmitter was the magnetron oscillator. Here, the phase of the oscillator bears no relationship from pulse to pulse. Hence the reference cannot be generated by a continuously operating oscillator. However a coherent reference can be obtained by readjusting the phase of the coho at the beginning of each sweep according to the phase of the transmitted power (sweep is the term for the period between pulses).

Here a portion of the transmitted signal is mixed with the STALO to produce a phase which is directly related to the phase of the transmitter. This IF pulse is applied to the coho and causes the coho to lock in step with the phase of the IF reference pulse.

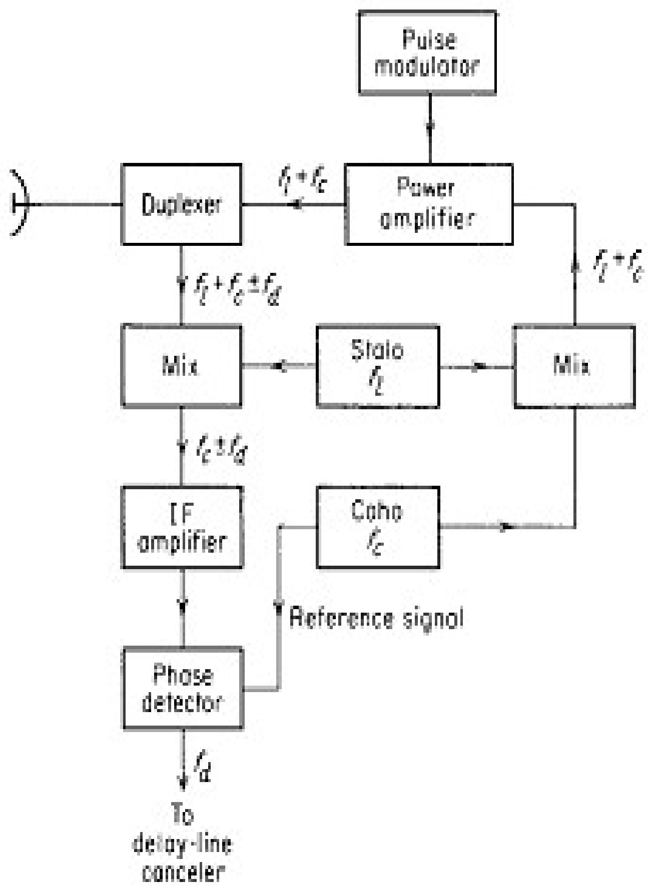


Figure 4.5 Block diagram of MTI radar with power-amplifier transmitter.

The phase of the coho is then related to the phase of the transmitted pulse and can be used as the reference for echoes from that **particular** transmitted pulse. Upon the next transmitted pulse, another IF locking pulse relocks the phase of the CW coho.

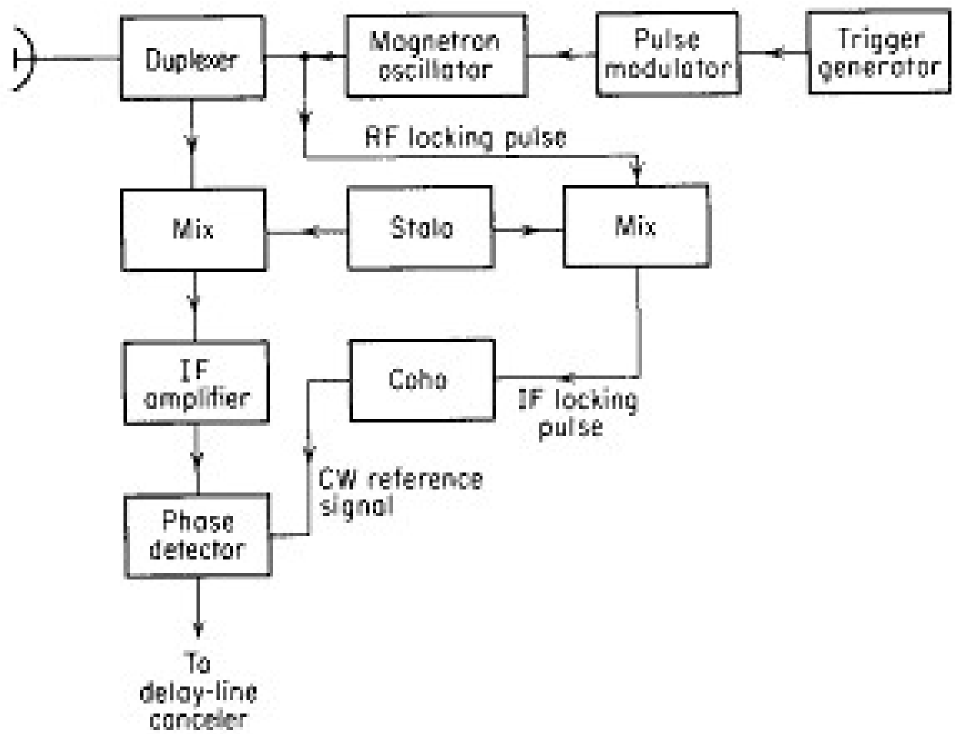


Figure 4.6 Block diagram of MTI radar with power-oscillator transmitter.

Advantages of MOPA (Master Oscillator Power Amplifier):

1. It can be easier to reach the desired performance i.e., in terms of line width, wavelength tuning range, beam quality or pulse duration. This is because various performance aspects are decoupled from the generation of high powers which gives the extra flexibility.

2. Low power seed laser can be modulated i.e., an optical modulator can be used between seed laser and power amplifier.

3. The combination of an existing laser with an existing amplifier may be simpler than developing a new laser with high output power.

4. Optical intensities are lower compared to intracavity intensities in an amplifier.

DISADVANTAGES OF MOPA:

1. The complexity of the setup is higher.
2. The wall-plug efficiency is often lower.
3. The resulting laser noise tends to be higher.
4. A MOPA can be sensitive to back reflections which are amplified again before entering the master laser.

APPLICATIONS OF MOPA:

1. In pulsed laser sources it can be used as a reservoir of energy.
2. This can be used in the deformation of the temporal pulse shape.

DIFFERENCE BETWEEM PULSE DOPLER RADAR AND MTI RADAR:

1. Pulse doppler radar, in this the radar send the pulse train to detect the position of target and MTI (moving target indicator) in which it detect the target which is moving

but pulse radar can detect the moving target but there is a disadvantage that the problem of blind speed arises and pulse radar dosn't continuosly transmit the pulse after transmitting wait for receiving in ths time it doesn't transmit any pulse.

2. Pulse doppler radar and MTI radar both are used to find the target range by using doppler effect(doppler shift) .but MTI radar uses low PRF whereas pulse doppler uses high PRF.

3. The different between MTI radar and PD radar is a unique even though they all relay on Doppler principle , but the MTI radar determine moving targets by detecting the phase and amplitude of the received wave and compare it with saved replica of the original transmitted wave but at opposite phase , so if the target are not moving then the phase and amplitude of the 2 signals will match but at different value will result of canceling each other, but if the 2 signals are not matched they will give positive or negative value and that is indication of moving target.

4. Pulse Doppler radar has anther interest, it is interested in the changes happen to the transmitted wave (DOPPLER SHIFT) either it will be compressed if the target moving toward the radar

   Example: received frequency may change from transmitted 6000MHz to 6010MHZ), or may stretch if the target are going away from the radar (the 6000MHz will be 5990MHz

5. The PD radar are not interested in the transmitted frequency any more after it has been transmitted but it does set filters around it at the expected reflected frequency Example:5970,5980,5990,6010,6020,6030. If there are moving targets then the filters will receive power and that is an indication of presence of target.

Delay Line Cancellers:

The time domain delay-line canceller capability depends on the qualityof the medium used as the delayline. The delay required is equal to the PRI and might be several milliseconds (long!!!). This can not be achieved with electromagnetic transmission lines.

Converting the electromagnetic signal to an acoustic signal allows the design of delaylines with reasonable physical length since the velocity of acoustic waves is approximately

10-5 that of electromagnetic
waves.

The signal at the output of the acoustic delay device is then converted back to an electromagnetic wave. Solid fused quartz using multiple internal reflections to obtain a compact device was developed in the 1950's. These analog delay lines have been replaced by digital delay lines using A/D and digital processing.

The use of digital processing allows implementation of complex delay line cancellers which were not practical with the analog methods.

Note: An advantage of the time domain delay line canceler as compared with the frequency domain filter is that a single network operates at all ranges and does not need a separate filter for each range resolution cell.

Filter Characteristics of a Delay Line Canceler:

The canceler acts as a filter to reject DC clutter but because of its periodic nature it also rejects energy near the PRF and its harmonics. The video received from a target at range R_0 is

$$v_1 = k\sin(2\pi f_d t - \phi_0)$$

The signal from the previous transmission which is delayed by a time T = PRI is

$$v_2 = k\sin(2\pi(f_d(t - T) - \phi_0))$$

The output from the pulse canceler is $$v = v_1 - v_2 = 2k\sin\pi f_d T\cos\left[2\pi f_d\left(t - \frac{T}{2}\right) - \phi_0\right]$$

The frequency response of the single delay-line canceller is $H(f) = 2k\sin(\pi f_d T)$.

The response of the single delay line canceller will be zero when the argument $\pi f_d T$ is 0, π, 2π… etc.

Definitions:

1. Blind speed is defined as the radial velocity of the target at which the MTI response is zero.
2. It is also defined as the radial velocity of the target which results in a phase difference of

   exactly 2π radians between successive pulses.

3. Blind speed is defined as the radial velocity of the target at which no shift appears making the target appearing stationary and echows from the targets are cancelled.

The speeds at which these occur are called the blind speed of the radar. These are

$$v_n = \frac{n\lambda}{2T};\ n=0,1,2 \qquad \text{or } v_n = \frac{n\lambda f_p}{2}$$

For λ in metres, f_p in Hz and vn in knots we have the following: $v_n = n\lambda f_p/(1.02) \approx n\lambda f_p$

$$v_n = n\lambda f_p$$

The first blind speeds in knots is given by $v_{b1} = 0.97\lambda f_p = \lambda f_p$

If the first blind speed is to be greater than the expected maximum radial speed, then λf_p must be large. Hence the MTI must operate at a long wavelength or high PRF or both. However there are other constraints on λ and f_p and blind speeds are not easily avoided.

There are four methods to reduce the effect of blind speeds by operating the radar at:

a. Long wavelength
b. High PRF
c. More than one PRF and
d. More than one wavelength.

The blind speed is dependent on the transmitted frequency and on the pulse repetition frequency of the radar unit.

Example given:

A radar unit works with the tx-frequency of 2.8 GHz and a puls repetition time of 1.5 ms. Under these conditions the first blind speed has got the value:

$$v_{blind} = \frac{\lambda}{2 \cdot T_s} = \frac{c_0}{2 \cdot f \cdot T_s} = \frac{3 \cdot 10^8}{2 \cdot 2{,}8 \cdot 10^9 \cdot 1{,}5 \cdot 10^{-3}} = 35{,}72\ {}^m/_s$$

c_0 = speed of light

This speed of converted about 130 km/h and all integral multiples of this also well cause that the target isn't visible in the range of the effectiveness of the MTI system.

Large λ has the disadvantage that antenna beamwidth is wider and is not satisfactory

where angular accuracy is important.

High f_p reduces the unambiguous

range.

Figure 4.8 shows the first blind speed v1 as a function of maximum unambiguous range

($R_{unambig} = cT/2$)

In practice MTI radars in L and S band are designed for the detection of aircraft and must operate with blind speeds if they are to operate with unambiguous range. Systems which require good MTI performance must operate with ambiguous range (pulse Doppler radar).

Example:

Table 2:

$R_{unambig}$	f
130 NM	200 MHz
13 NM	3 GHz
4 NM	10 GHz

The effect of blind speeds can be reduced by operating with more than one PRF (staggered PRF MTI). Operating at more than one RF frequency can also reduce effect of blind speeds.

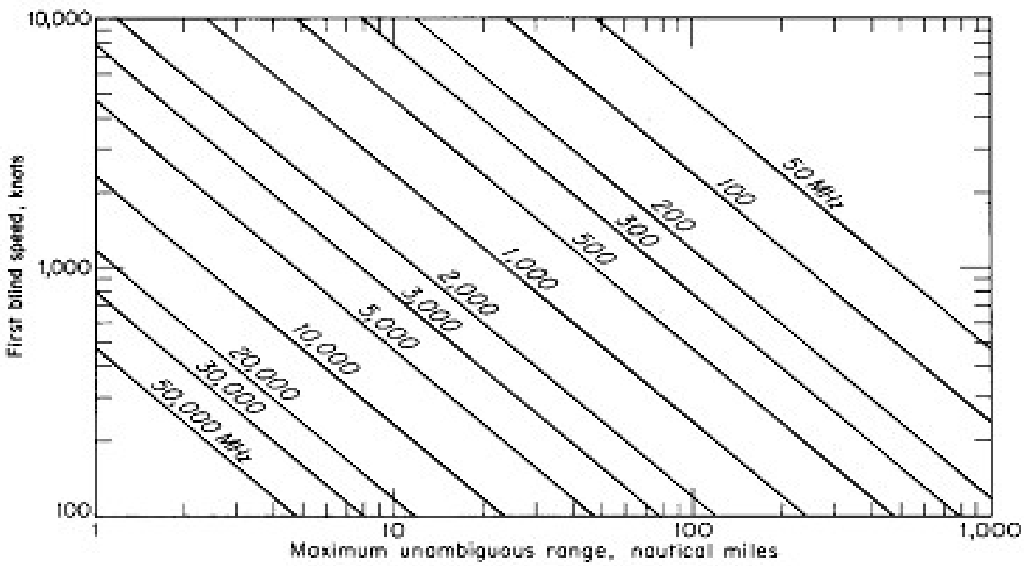


Figure 4.8 Plot of MTI radar first blind speed as a function of maximum unambiguous range.

Double Cancellation:

Single delay line cancelers do not always have as broad a clutter reject null at DC as might be desired. The null can be widened by the use of cancelers as shown In Fig 4.9a, the output of the two single delay line cancellers in cascade is the square of that from a single delay canceler. $|v| = 4\sin(\pi f_d T)^2$

The response of this double delay line canceler is shown in Fig. 4.10.The finite width of the typical clutter spectrum is also shown to illustrate the additional cancellation offered.

The two delayline canceler of Fig. 4.9b has the same frequencyresponse characteristic as the double delay line canceler.

Here the output of the adder is $f(t) = -2f(t+T) + f(t+2T)$ Where f(t) is the received video signal

This arrangement is called a three pulse canceler.

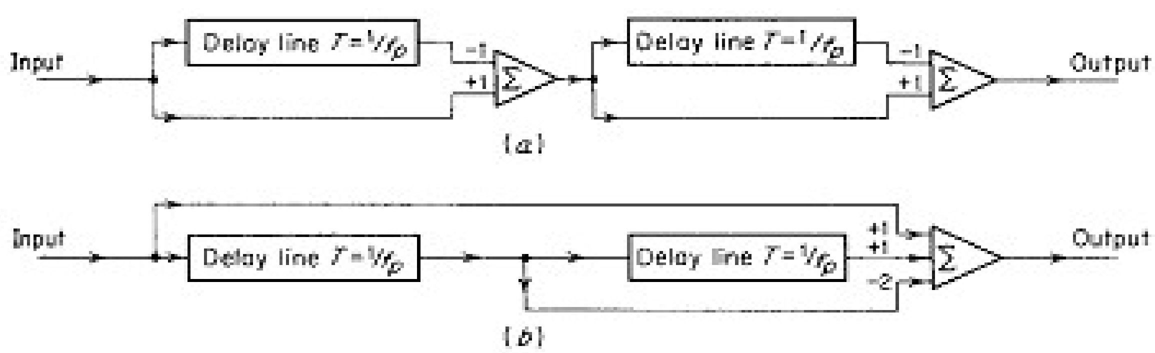


Figure 4.9 (a) Double-delay-line canceler; (b) three-pulse canceler.

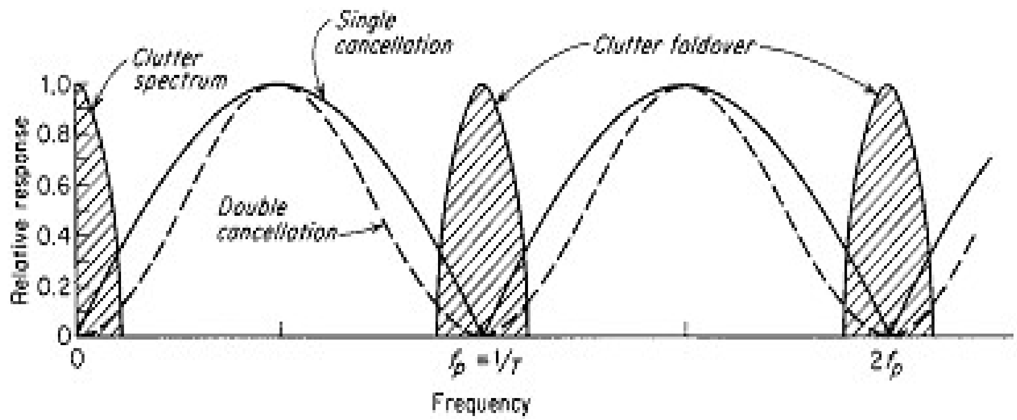


Figure 4.10 Relative frequency response of the single-delay-line canceler (solid curve) and the double-delay-line canceler (dashed curve). Shaded area represents clutter spectrum.

Transversal Filters:

The 3 pulse filter of Fig. 4.9b is a transversal filter. The general form is shown below for N pulses and N-1 delays. It is also called a feed-forward filter, a nonrecursive filter, a finite memory filter and a tapped delay line filter. The weights for a transversal filter with n delay lines that gives a $\sin^n(\pi f_d T)$ are the binomial coefficient with alternating sign

$$w_i = (-i)^{i-1}\frac{n!}{(n-i+1)!(i-1)!} \quad ; \quad i=1,2,3....n+1$$

Note: a cascade configuration of three delay lines each connected as a single canceller is called a triple canceler: but when connected as a transversal filter it is called a four pulse canceler.

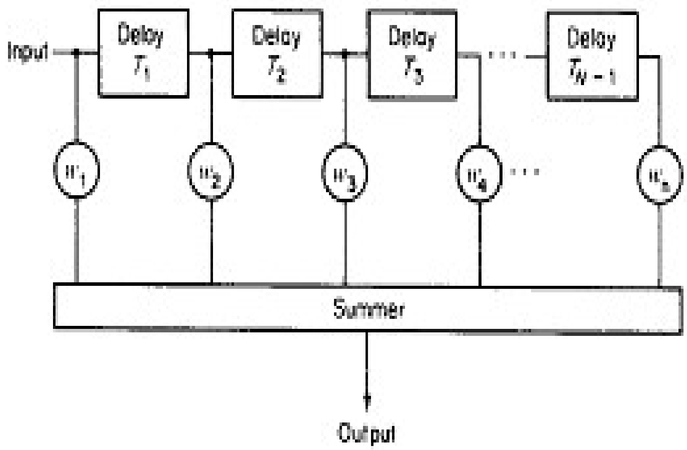


Figure 4.11 General form of a transversal (or nonrecursive) filter for MTI signal processing.

The transversal filter with alternating binomial weights is closely related to the filter which maximizes the average of the ratio

$$I_c = \frac{(S/C)_{out}}{(S/C)_{in}}$$

Where S/C is the signal to clutter ratio And I_c is the improvement factor for clutter

The average is over the range of Doppler frequencies. Ic is independent of target velocity and depends onlyon the weights w_i, the autocorrelation function (power spectral density) of the clutter, and the number of pulses.

For the two pulse canceler (single delay line), the optimum weights based on the above criterion are the same as the binomial weights when the clutter is Gaussian.

For the three pulse canceler, the difference between filter with optimal weights and one with binomial weights is less than 2 dB.

The difference is small for higher order cancelers also.

For anyorder canceler, the difference between optimum weights and binomial weights is small over a wide range of clutter spectral widths.

Similarlythe use of a criterion which maximizes the clutter attenuation (Cin/Cout) is also well approximated by a transversal filter with binomial weights with alternating signs, when the clutter spectrum can be represented by a Gaussian function whose spectral width is small compared to the PRF. The transversal filter with binomial weights also approximates the filter which maximizes the P_d for a target at midband Doppler frequency or its harmonics.

The disadvantage is due to notches at DC, the PRF and harmonics of PRF increasing in width as the number of delay lines increases. The added delay lines reduce the clutter but also reduce the number of moving targets because of the reduced pass band. For targets uniformly distributed across the Doppler frequency:

We have for the criterion that -10dB response of the filter is the threshold for detection, the following table:

Although these filters are optimum for I_c theyare not necessarilybest. They are onlybest under a given set of assumptions. It would seem that the MTI filter should be shaped to reject the clutter at DC and around the PRF, but have a flat response over the region where no clutter is expected.

% of targets rejected	n pulse canceler
20	2
35	3
48!	4

The transversal filter can be designed to achieve this desirable filter response but needs a large number of delay lines. This sets a restriction on the radar's PRF, beamwidth, antenna rotation rate and dwell time.

Fig 4.12 shows the amplitude response of a 3 pulse canceler with sin2 $\pi f_d T$(1), a 5 pulse "optimum" canceler which maximizes Ic (2) and a 15 pulse Chebyshev filter (3). Even a 5 pulse Chebyshev design gives a significantly wider bandwidth than the 5 pulse optimum design.

Note: the Chebyshev design has a lower Ic but is worthwhile if the clutter spectrum is narrow.

When there are onlya few pulses available for processing, there is little that can be done to control the filter shape. Hence for 3 or 4 pulse cancelers it is best to use the classical sin2 or sin4 response of the binomial cancelers.

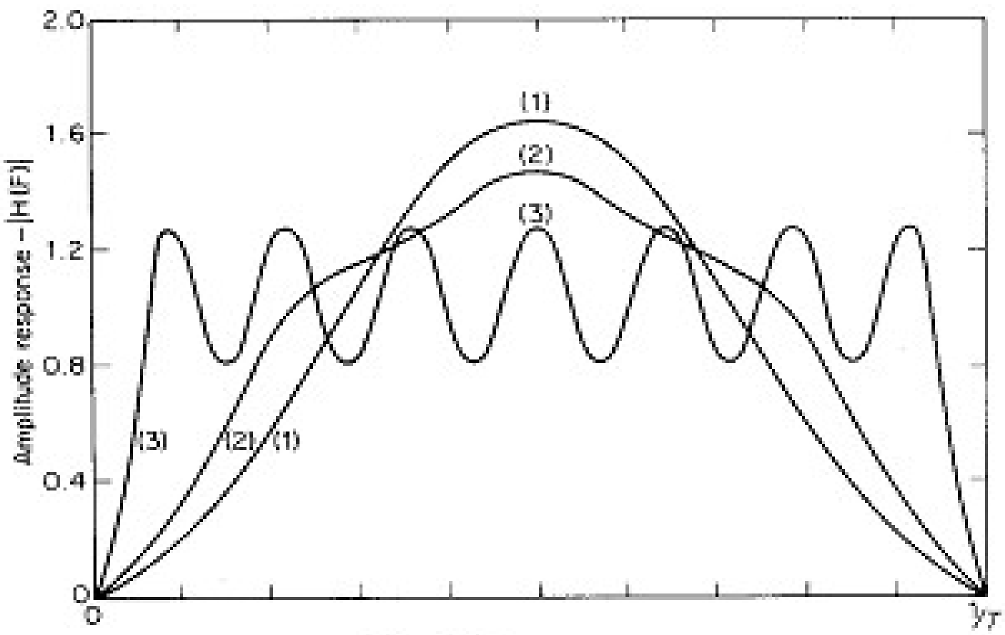


Fig.4.12

Shaping the Frequency Response:

Non recursive filters employ only feedforward loops. These yield zeroes for synthesizing the frequency response. If feedback loops are used, each delay line can provide one pole as well as one zero. The canonical form is shown below:

When feedback loops are used the filter is called recursive. Can synthesize in principle almost any frequency response using z transform techniques. Canonical configuration is not always desirable due to delay tolerances.

It has been shown that the canonical configuration can be broken into cascaded sections with no section having more than two delay elements. This synthesis technique applies to Chebyshev, Butterworth, Bessel and elliptical filter design.

The recursive delay line filters offer steady state response that is superior to that of nonrecursive filters. However they have poor transient response due to the feedback loops. The presence of large clutter returns can effectively appear as a large step input which leads to severe ringing in the filter output which can mask the target signal until the transient response dies out. In surveillance radar, the number of pulses from anytar get is limited. Hence the MTI filter is almost always in a transient state for most recursive filters.

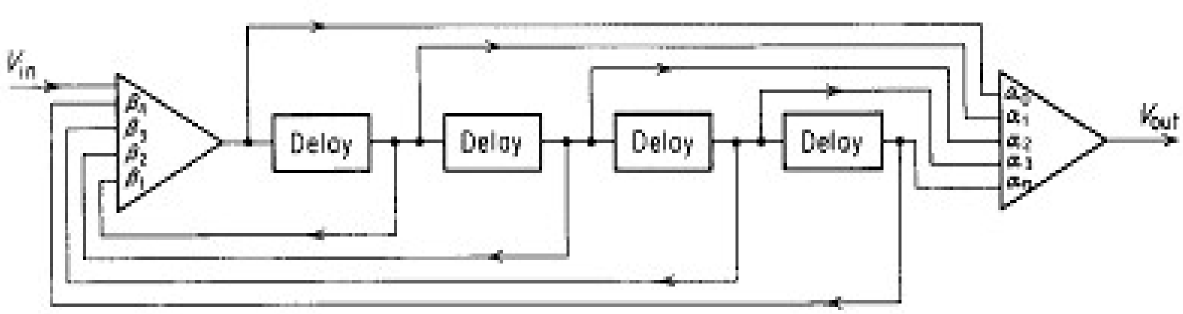


Figure 4.13 Canonical-configuration comb filter. (*After White and Ruvin,*[2] *IRE Natl. Conv. Record.*)

A filter with steep frequency response skirts might allow 15 - 30 pulses to be generated at the filter output due to feedback. This could make the system unusable in situations with large discrete clutter, or with interference or jamming from other radars. Poor transient response is also undesirable in radars with step scan phased array, since the e xtra pulses might have to be gated out after the beam has been moved.

For step scan radars the undesirable transient effects can be overcome by using the initial return from each new beam position to apply initial conditions to the MTI filter. These initial conditions are the steady state values that would appear in the filter after an infinitely long sequence. This effectively suppresses the transient response. An alternative to the recursive filter is to use multiple PRFs.

Multiple (Staggered) PRFs:

Multiple PRFs reduce the effect of blind speeds and also allow a sharper low frequency cutoff. The blind speeds of two independent radars will be different if their PRFs are different. This same result can be achieved with one radar which shares its PRFs between 2 or more values.

PRF can be switched every other scan, every time the antenna is scanned half a beam width or pulse to pulse (staggered PRF). Fig. 4.16 shows the composite response of MTI with two separate PRFs with a ratio of 5:4

Note: The first blind speed of the composite is greatly increased (i.e. at $f_d = 4/T1 = 5/T2$). But regions of low sensitivity appear.

The closer the ratio T_1: T_2 approaches unity, the greater the frequency y of the first blind speed, and the deeper he first null in the vicinity of f_d - $1/T_1$. The null depth can be reduced and the first blind speed increased by operating with more than 2 PRFs.

Fig 4.17 shows a five pulse stagger (4 periods) response with periods in the ratio 25:30:27:31

Here the first blind speed is 28.25 times that of a constant PRF with the same average

period. If the periods of the staggered pulses have the relationships

$$\frac{n_1}{T_1} = \frac{n_2}{T_2} = \cdots \frac{n_n}{T_N}$$ Where $n_1, n_2, \ldots n_N$ are integers

And if v_B is the first blind speed of a non staggered pulse with period equal to the average period

$$T_{ave} = \frac{1}{N}[T_1 + T_2 + \dots T_N]$$ Then the first blind speed v_1 is $$v_1 = v_B\left[\frac{n_1 + n_2 + \dots n_N}{N}\right]$$

Weighting can also be applied to received pulses of a staggered PRF. Fig. 4.18 dashed curve shows the response of a 5 pulse canceller with fixed PRF and using weights of 7/8;1;-3 3/4;

1 7/8. The solid curve is for a staggered PRF with the same weights but with 4 interpulse periods of -15%, -5%, 5%, 15%.

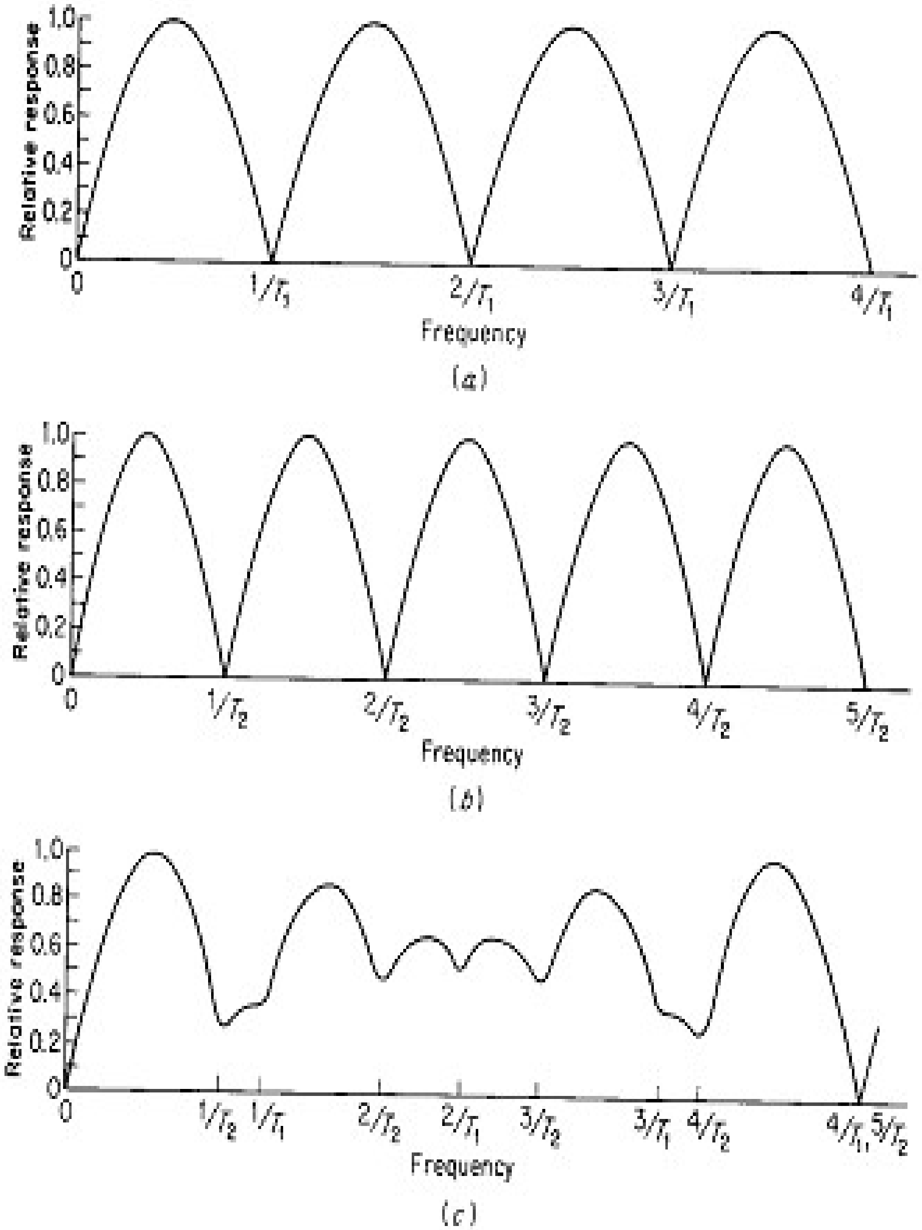


Figure 4.16 (a) Frequency-response of a single-delay-line canceler for $f_p = 1/T_1$; (b) same for $f_p = 1/T_2$; (c) composite response with $T_1/T_2 = \frac{4}{5}$.

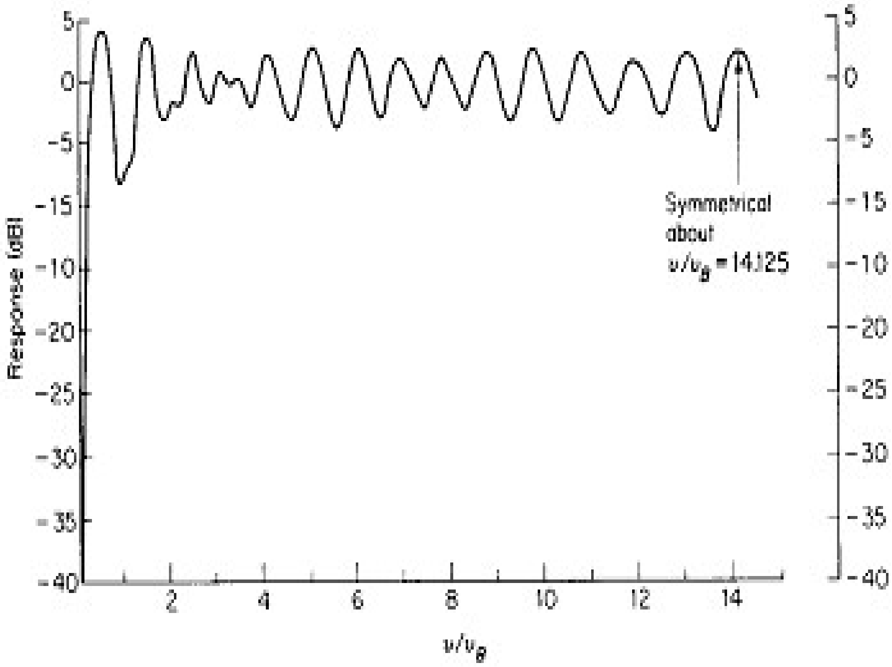


Figure 4.17 Frequency response of a five-pulse (four-period) stagger. (*From Shrader,[8] Courtesy McGraw-Hill Book Co.*)

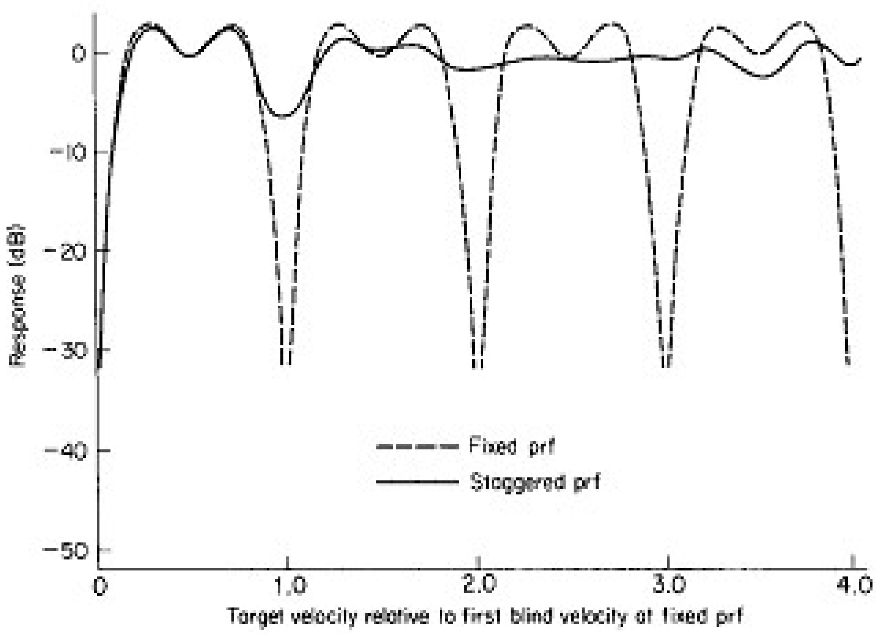


Figure 4.18 Response of a weighted five-pulse canceler. Dashed curve, constant prf; solid curve, staggered prf's. (*From Zverev,[22] Courtesy IEEE.*)

Note that the response at the first blind speed is down only 6.6dB. The disadvantage of staggered PRF is its inability to cancel second time around clutter echoes. Such clutter does not appear at the same range from pulse to pulse and produces uncancelled residue.

Second time around clutter can be removed by the use of constant PRF providing there is pulse to pulse coherence (i.e. power amplifier form of MTI). Constant PRF might be employed only over angular sectors where second time around clutter is expected, or by changing the PRF each time the antenna scans half a beam width, or by changing the PRF each scan period.

Range Gated Doppler Filters:

It is possible to use frequency domain band pass filters in MTI radar to sort the Doppler frequency shifted targets. A narrow band filter with a pass band designed to pass the Doppler components of a moving target will ring when excited by a short radar pulse.

The narrow band filters means the input pulse since the impulse response is the reciprocal of the filter bandwidth. This smearing destroys the range resolution.

If there is more than one target in the smeared region they can not be resolved. Even for one target, noise from other range cells will interfere (collapsing loss) resulting in reduction in sensitivity. The loss of range information and collapsing loss can be eliminated by quantizing the range into small intervals (range gating). The width of the range gate is usually of the order of the pulse width. The range resolution is established by range gating. Once the return is quantized, the output from each gate is applied to a narrow band filter since the pulse shape need no linger be preserved for range resolution. Collapsing loss does not take place since noise from other range intervals is excluded.

The output of the phase detector is sampled sequentially by range gates. The range gate acts as a switch which opens and closes at the proper time. Gates are activated once each PRI. The output from a stationary target is s series of pulses of constant amplitude. An echo from a moving target produces pulses which vary in amplitude according to the Doppler frequency.

The output from the range gates is stretched in boxcar generators (or sample and hold circuits). This helps the filtering and detection process by emphasizing the fundamental of the modulation frequency and eliminating harmonics of the PRF. Clutter rejection filter is a band pass filter whose bandwidth depends on the clutter spectrum.

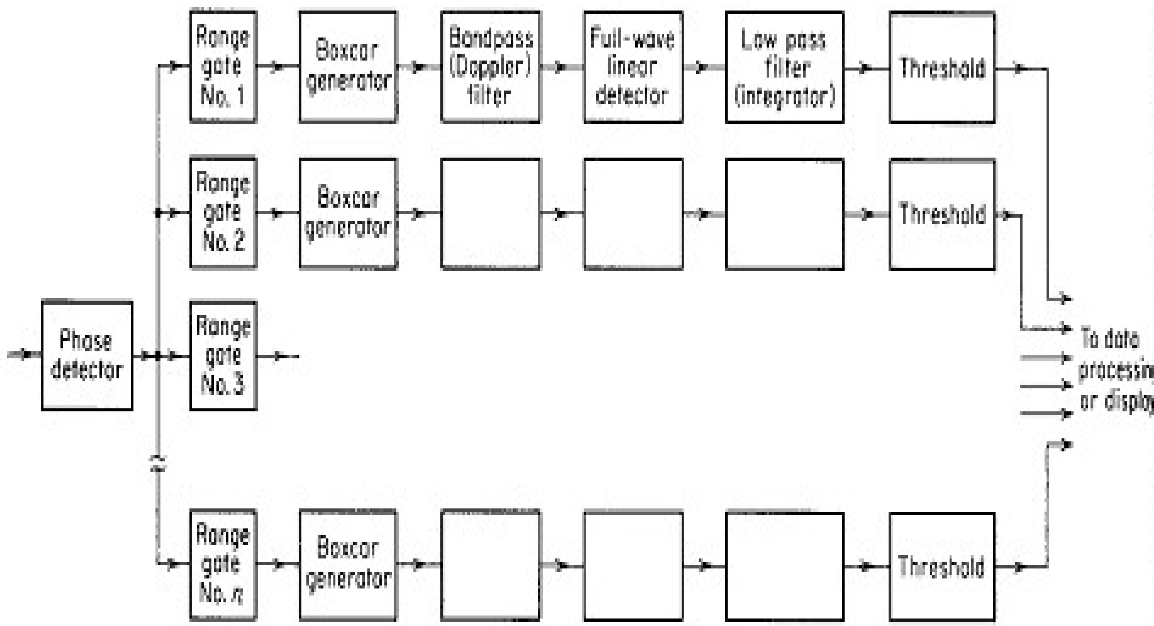


Figure 4.19 Block diagram of MTI radar using range gates and filters.

The full wave rectifier then converts the bipolar video to unipolar form. An integrator acts as a low pass filter. The signal is then applied to the threshold detector circuit.

Following threshold detection the outputs from each range channel must be properly combined for display on the PPI etc.

The band pass filter can be designed with a variable low frequency cutoff that can be varied to conform to prevailing clutter conditions (i.e. adaptive).

For example, a wide notch at DC is needed to remove echoes from birds but mayalso remove some desired targets. Since the appearance of birds varies with time of day and season the adaptive notch width might be important.

MTI using range gates is more complex than MTI with single delay line canceler. MTI using range gates improves performance and allows flexibility of range gates and filter bandwidth.

Difference between MTI Radar and CW Radar:

CW Radar	MTI Radar
1. Using this radar, we can not measure range	1. Using this radar, we can measure range of a
2. This radar can not have blind speeds and	2. This radar has blind speeds and phases.
3. It is difficult to avoid transmitter to receiver	3. It is easy to avoid transmitter to receiver
4. This radar has little inability to distinguish between approaching and receiding target.	4. This radar not only distinguishes the approaching and receiding target but also

Differences between blind speeds and blind phases:

Blind speeds	Blind phases
1.The relative velocities of the target at which MTI response is zero are called as blind speeds and are given by $v_n = \frac{n\lambda}{2T}$; n=0,1,2 or $v_n = \frac{n\lambda f_p}{2}$	1. The blind phases are due to the presence of sampling pulses at the same point in the Doppler cycle at each sampling instant.
2. Due to the presence of blind speeds within a Doppler frequency band, the capability of	2. When the Doppler frequency is half of the PRF, blind phases with single has serious
3. By operating with more than one PRF or operating at more than one RF frequency	3. By using quadrature, we can eliminate the blind phases.

4. Effect is more in MTI Radars.	4. Effect is more in MTI Radars.

Advantages of Time domain Delay-line Canceller over Conventional Frequency Domain Filter:

1. The delay-line canceller has been widely used in MTI Radar, in order to separating moving targets from stationary clutter ehile the filter configuration in conventional frequency domain filter is more complex.
2. The delay-line canceller rejects the DC component of clutter.
3. In time domain delay-line canceller, a single network operates at all frequency ranges, it does not require a separate fiter for each frequency range.
4. The gain through the delay-line canceller is unity.
5. Delay-line cancellers are used in digital signal processing through which great stability is achieved.
6. The delay-line canceller serves as an important tool for a MTI designer.
7. The time domain delay-line canceller provides a time delay equal to the pulse repetition period.
8. The digital processing technology using the delay line cancellers made an introduction of digital devices for computing the Fourier transform.

Digital Signal Processing:

Allows multiple delay line cancelers with tailored frequency response to be achieved. Is more dependable, requires less adjustment than analog cancelers, does not vary with temperature etc. Most advantages of digital MTI result from the use of digital storage in the place of analog delay lines. Here the IF signals is split into I and Q channels. The outputs of the two phase detectors are 90˚ out of phase.

The quadrature channel eliminates the effects of blind phases. Note that this is seldom done in analog delay line cancelers due to the complexity involved with additional analog delay lines. The baseband bipolar video is sampled at a rate to obtain one or more samples per range resolution cell. The A/D output is stored in the memory for one PRI and then subtracted from the words from the next sweep. The I and Q outputs are combined as either

$$\sqrt{I^2 + Q^2} \quad \text{or} \quad |I| + |Q|$$

.

The combined output is then converted to an analog signal with D/A for display.

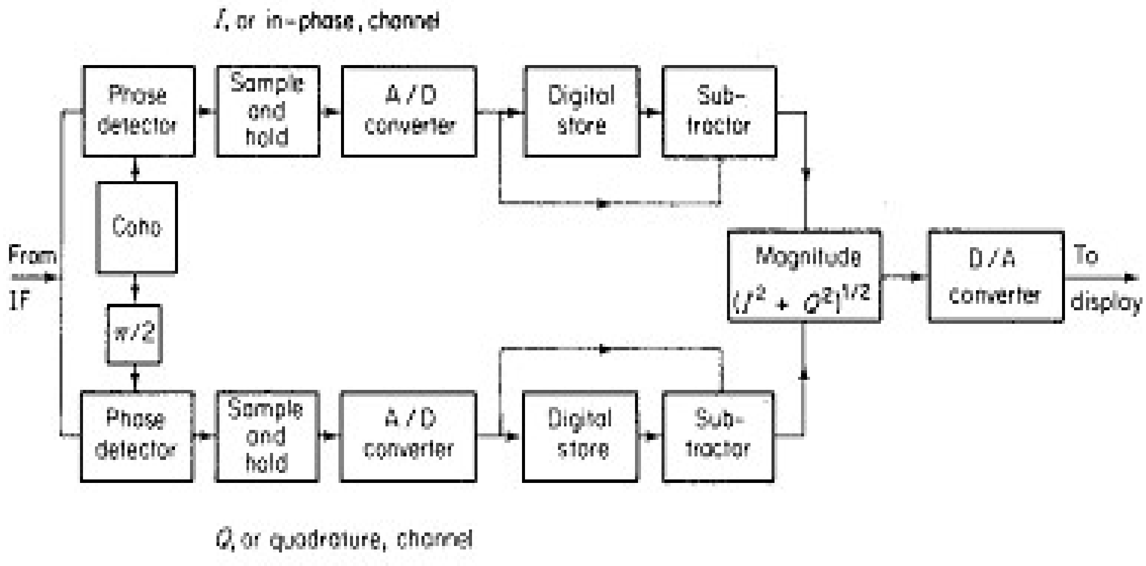


Figure 4.21 Block diagram of a simple digital MTI signal processor.

Note: more complex filtering schemes are normally implemented.

The memory can be RAM or a shift register. A/D must operate at high speeds to preserve the information content, and also have sufficient number of bits to quantize the signal to preserve precision. The number of bits determines the maximum improvement factor for the MTI radar. The A/D converter must cover the peak excursion of the phase detector output. To ensure this, a limiter may be necessary. The quantization noise introduced by the A/D imposes a limit to the improvement factor of

$$I_{QN} \approx 20\log[(2^N - 1)\sqrt{0.75}]$$

Where 2N - 1 is the number of discrete intervals. This is approximately 6dB / bit

Example: **A** 9 bit A/D has 511 levels. Thus the maximum resolution is 1 out of 511

$$I = 20\log\left[\frac{511}{1}\right] = 54 \text{ dB}$$

Note: I_{QN} = 52.9 which is close

Blind phase - this is different from blind speed. Blind speed occurs when pulse sampling occurs at the same point on the Doppler cycle at each sampling instant (Fig 4.22a).

Fig. 4.22b shows the I channel with the pulse train such that the signals are of the same amplitude and with spacing such that when pulse a_1 is subtracted from pulse a_2 the result is zero. However a residue is produced when a2 is subtracted from a_3, but not when a3 is subtracted from a_4.

Note: The pulse pairs that were lost in the I channel are recovered in the Q channel and vice versa.

The combination of I and Q channels results in a uniform signal with no loss. An extreme case, resulting in the complete loss of signal, occurs, with only one channel when the Doppler frequency is half of the PRF and the phase between the two is such that the echo pulses lie on the zeroes of the Doppler signal. However if the Q channel is added the all of the echo pulses occur at the peaks of the Doppler signal. Other phase relationships also give loss with single channel MTI.

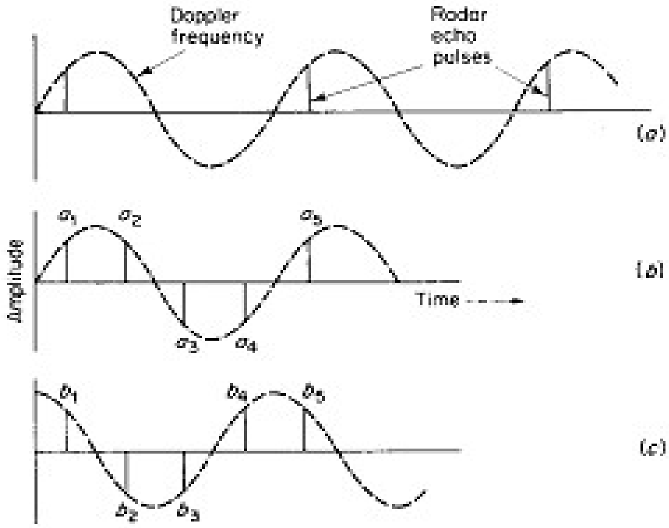


Figure 4.22 (*a*) Blind speed in an MTI radar. The target doppler frequency is equal to the prf. (*b*) Effect of blind phase in the *I* channel, and (*c*) in the *Q* channel.

Advantages of digital processing over analog delay lines:

1) Greater stability
2) Greater repeatability
3) Greater precision
4) Greater reliability
5) No special temperature control required
6) Greater dynamic range (i.e. no spurious response)
7) Avoids restriction that PRI and delay time must be equal
8) Different PRI can be used (without switching delay lines of various lengths in and out)
9) Greater freedom in selection of amplitude weighting for shaping filters
10) Allows easy implementation of I and Q channels (eliminates blind phases)

Digital Filter Banks and the FFT:

A transversal filter with N outputs (N-1 delays) can be made to form a bank of N contiguous filters covering the spectrum of 0 to PRF. Consider the transversal filter of Fig. 4.11 to have N-1 delays each with a delay of T = $1/f_p$. Let the weights at the outputs be

$$w_{ik} = e^{-j[2\pi(i-1)k/N]}$$

Where i = 1,2,... N represents the ith tap And k = an index from 1 to N-1

Each k value corresponds to a different set of N weights and to a different doppler

frequency response. The N filters generated by the index k constitute the filter bank. The impulse response of the transversal filter with the above weights is:

$$h_k(t) = \sum_{i=1}^{N} \delta[t-(i-1)T]e^{-j2\pi(i-1)k/N}$$

The Fourier transform yields the frequency response $$H_k(f) = e^{-j2\pi fT} \sum_{i=1}^{N} e^{j2\pi(i-1)[fT-k/N]}$$

The magnitude of the frequency response yields the amplitude characteristic

$$|H_k(f)| = \left|\sum_{i=1}^{N} e^{j2\pi(i-1)[fT-k/N]}\right| = \frac{\sin[\pi N(fT-k/N)]}{\sin[\pi(fT-k/N)]}$$

Note that the peak response occurs when the denominator is zero i.e.

$\pi(fT\text{-}k/N)= 0,\pi,2\pi$ ....

for k=0, the peak response occurs at f= 0,1/T,2/T etc.

This is a filter centred at DC, the PRF and harmonics of the PRF. This filter has no clutter rejection capability, b ut it is useful fro pro viding a map of clutter.

The first null occurs when the numerator is 0

For k=0 this is f= 1/NT

The bandwidth between nulls is then 2/NT and the half power bandwidth is: BW(3dB) ≈ 0.9/NT

When k=1, the peak response occurs at f = 1/NT, 1/T + 1/NT....

Hence each k defines a separate filter response, each with a bandwidth between nulls of 2/NT.

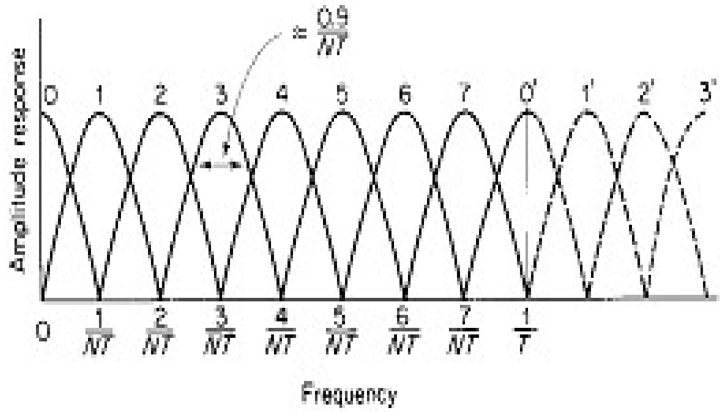


Figure 4.23 MTI doppler filter bank resulting from the processing of $N = 8$ pulses with the weights of Eq. (4.13), yielding the response of Eq. (4.16). Filter sidelobes not shown.

Since each filter occupies the 1/N bandwidth of a delayline canceler, its SNR will be greater than the delay line canceler. Also the dividing of the band by N filters allows a measure of the Doppler frequency to be made. If moving clutter (birds, weather) appears, the threshold of each filter maybe adjusted individuallyto adapt to the clutter in it. This selectivity allows clutter to be removed which would be passed by a delay line canceler. The first sidelobe level is at -

13dBc for the filters with the above weights. This can be reduced at the expense of wider bandwidth by adjusting the weights (windowing).

Two forms of windows are the Hamming and the Hanning.

It is not always convenient to display outputs of all of the Doppler filters. One approach is to connect filter outputs to a “greatest of” circuit, so that only one output is obtained (Note: the filter at DC would be excluded).

The improvement factor for each of the 8 filters of a 8 filter bank is shown in Fig. 4.24 as a function of the standard deviation of the clutter spectrum, for a spectrum represented as a Gaussian function. For comparison, the improvement factor for an N pulse canceler is shown in Fig. 4.25

Note: the improvement factor for a two pulse canceler is almost as good as that of the 8 pulse Doppler filter bank. However, maximizing the improvement factor is not the onlycriterion for judging the effectiveness of MTI Doppler processes. If a 2 or 3 pulse canceler is cascaded with a Doppler filter bank, better clutter rejection is obtained.

Fig. 4.26 gives I_c for a 3 pulse canceler cascaded with an 8 pulse filter bank. The upper figure assumes uniform amplitude weights (-13.2 dB sidelobes) and the lower figure gives the results for Chebyshev weights (-25 dB sidelobes). It is found that doubling the number of pulses

in the filter bank to 16 does not offer significant improvement (i.e. increased pulses ⇒ increase

in bandwidth as well as decreased sidelobes. If the sidelobes of the individual filters in the filter

bank can be made low, the inclusion of the delay line canceler ahead of it might not be necessary.

Other MTI Delay Lines:

Solid crystal has been mentioned previously.

Electrostatic storage tube

- Here the signals are stored on a mesh similar to that of a TV camera. The first sweep reads the signal on the storage tube. the next sweep written on the same space and generates the difference between the two. Two tubes are required; one to write and store the new sweep, the other to subtract the new sweep from the old. The electrostatic storage tube can be used with different PRFs.

The charge transfer device (CTD) is a sampled data, analog shift register which can be used as a delayline, transversal filter and recursive filter.

There are two types of CTD:

- ❖ The bucket brigade device (BBD) with capacitive storage elements separated byswitches.

  Here the data is transferred through the BBD at a rate determined bythe rate at which the Switches are operated. The delay time is determined by the number of stages and the switching (or clock) speed.

- ❖ Charge coupled devices (CCDs) are similar except that charge packets are transferred to adjacent potential wells by a clocked potential gradient

Notes: The sampling frequency must give one or two samples per resolution cell. No A/D or D/A required as for digital systems.

MTI cancelling can be done at IF. Here no blind phases occur. Finally the video can FM modulate a carrier f or acoustic delay. This avoids problems of amplitude stability of acoustic delay lines.

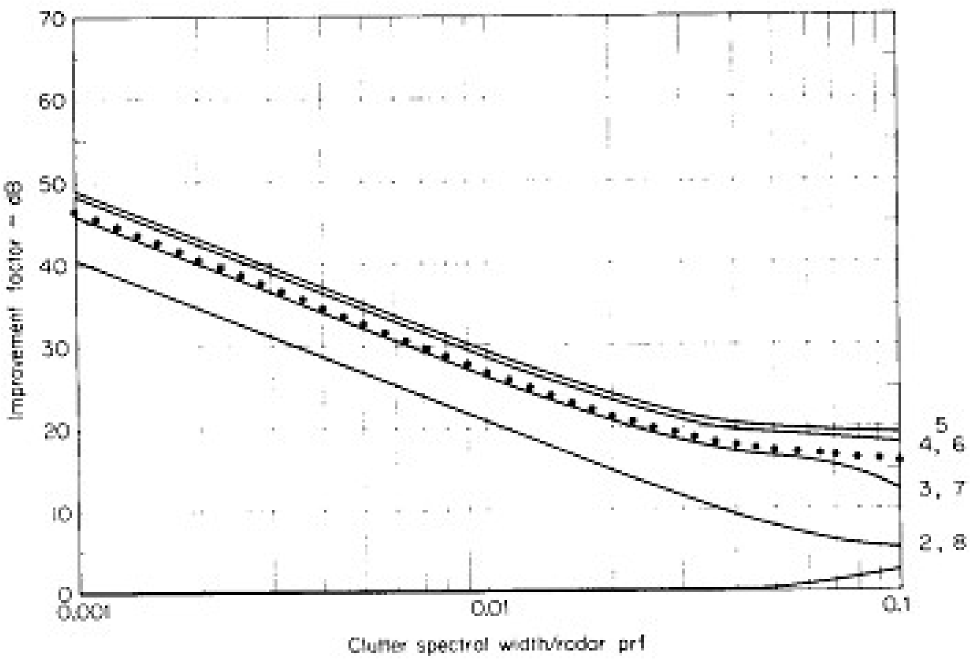


Figure 4.24 Improvement factor for each filter of an 8-pulse doppler filter bank with uniform weighting as a function of the clutter spectral width (standard deviation). The average improvement for all filters is indicated by the dotted curve. (*From Andrews.*[30])

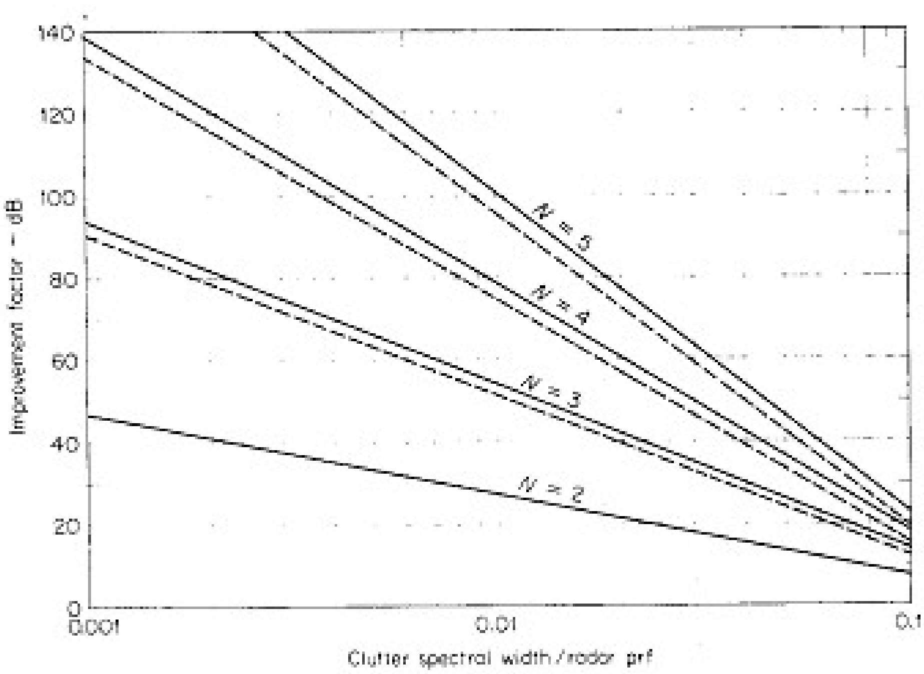
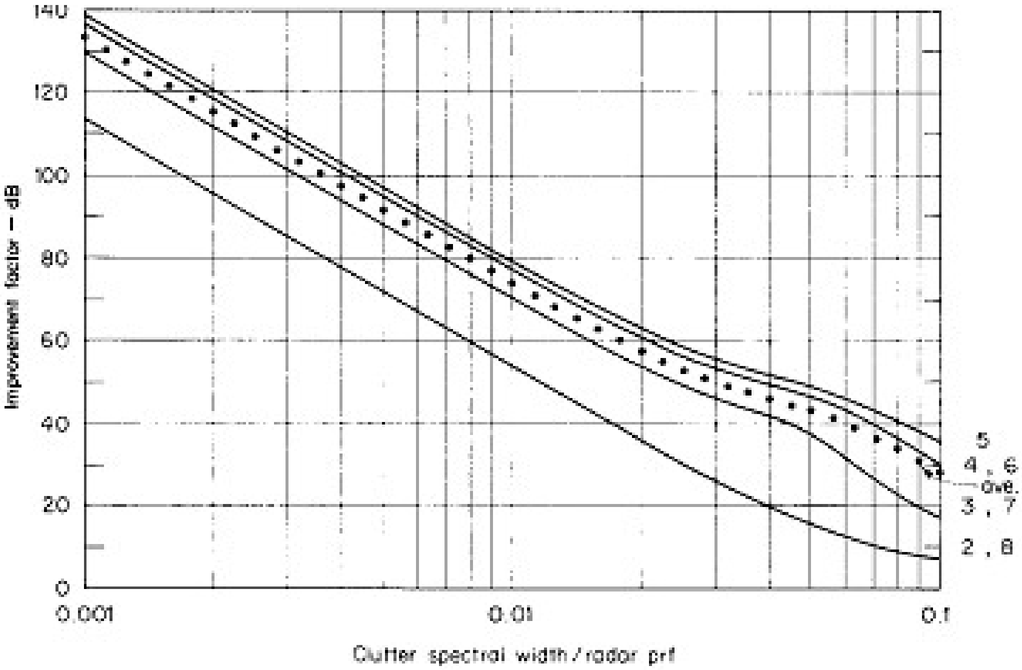
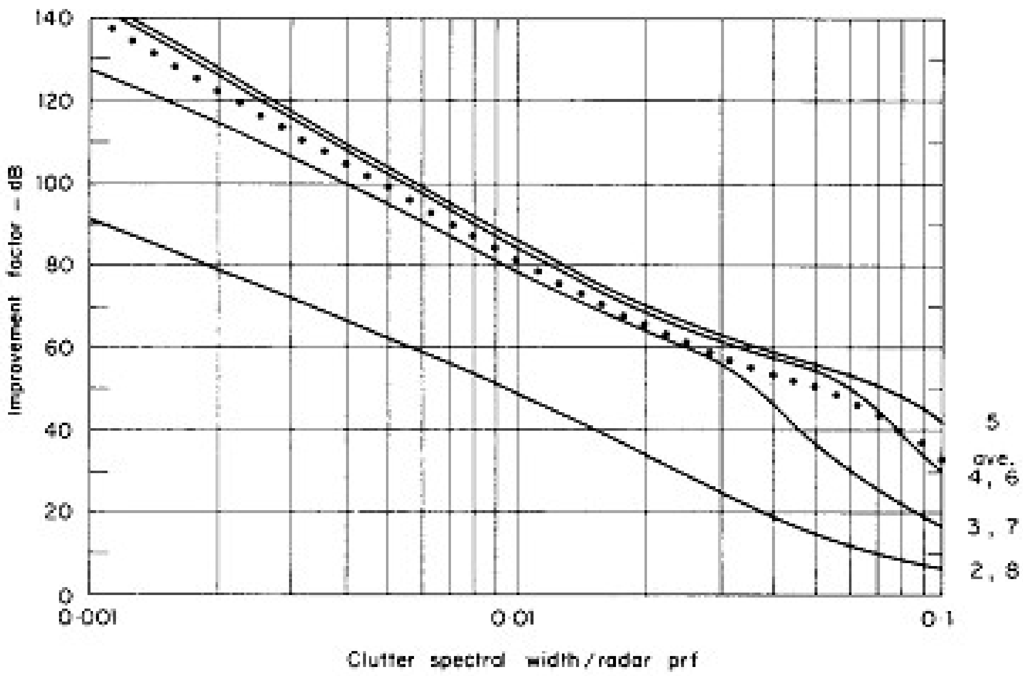


Figure 4.25 Improvement factor for an N-pulse delay-line canceler with optimum weights (solid curves) and binomial weights (dashed curves), as a function of the clutter spectral width. (*After Andrews.*[31,32])



(*a*)



(*b*)

Figure 4.26 Improvement factor for a 3-pulse (double-canceler) MTI cascaded with an 8-pulse doppler filter bank, or integrator. (*a*) Uniform amplitude weights and (*b*) 25-dB Chebyshev weights. The average improvement for all filters is indicated by the dotted curve. (*From Andrews.*[30])

Example of an MTI Radar Processor:

The moving target detector (MTD) was developed byLincoln Labs for the FAA's airport surveillance radars (ASR). The ASR is medium range (60 NM) radar located at most US airports. It has 2.7 to 2.9 GHz RF, 1μs pulse width, 1.4 ° azimuth beamwidth, antenna rotation rate of 12.5 to 15 RPM PRF from 700 to 1200 Hz and average power from 400 to 600 W.

The MTD processor is based on digital technology. Uses a 3 pulse canceler followed by an 8 pulse FFT Doppler filter bank with weighting to reduce sidelobes. The total system has alternating PRFs to eliminate blind speeds, adaptive thresholds and a clutter map used for detecting targets with zero radial speed.

The measured Ic of the MTD on the ASR was 45 dB (20 dB better than with a conventional 3 pulse MTI processor). The MTD also achieves a narrower notch at zero velocity blind speeds. The processor is preceded by a large dynamic range receiver to avoid reduction in Ic caused by limiting.

The IF is applied to I & Q detectors and the video is A/Dd with 10 bit resolution. The range totalling 47.5 NM is divided into 1/16 NM intervals and the azimuth into 0.75 ° intervals. In each 0.75° interval (≈ 1/2 beamwidth) 10 pulses are transmitted at constant PRF.

On receive the 10 pulses are processed by the delay line canceller and Doppler filter bank which forms 8 Doppler filters. Hence the radar output is divided into 2,920,000 range/azimuth/doppler cells. Each cell has its own adaptive threshold. In the alternate 0.75° azimuth intervals another 10 pulses are transmitted at a different PRF. Changing PRF every 10 pulses eliminates second time around clutter. The MTD processor is shown in Fig. 4.27. The 3 pulse canceler ahead of the filter bank eliminates stationary clutter ahead of the filter bank and reduces the dynamic range required of the Doppler filter bank.

Since the first two pulses of a three pulse canceler are useless, only the last 8 of the 10 pulses are passed to the filter bank (implemented using the FFT).

Following the filter bank, weighting is applied in the frequencydomain to reduce the filter sidelobes. This is necessary since the frequency spectrum of rain and wind blown clutter requires sidelobes lower than the -13.2 dB available from uniformly weighted samples.

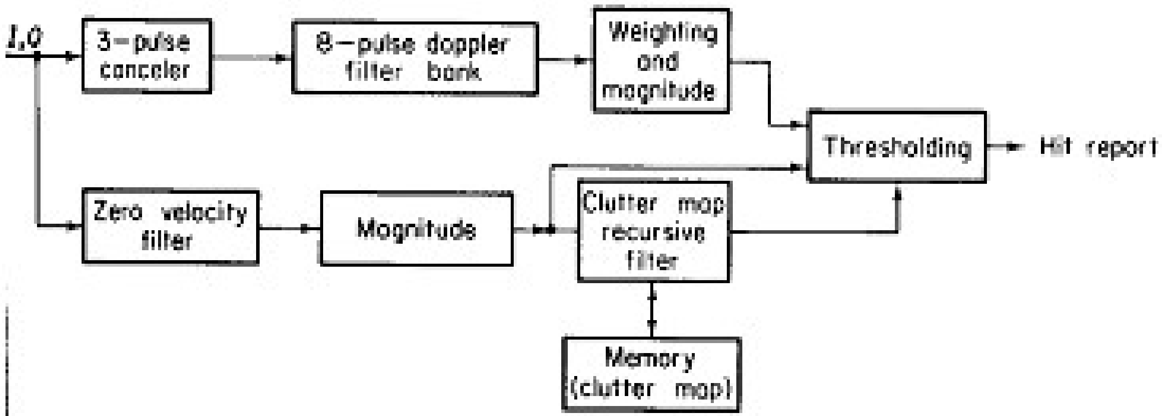


Figure 4.27 Simple block diagram of the Moving Target Detector (MTD) signal processor.

The magnitude operation forms $\sqrt{I^2 + Q^2}$

Separate thresholds are applied to each filter. For non zero velocityresolution cells the thresholds are established byscanning the detected outputs of the same velocityfilter in 16 range cells, eight on either side of the cell of interest. Hence each filter output is averaged over one mile in range to get the mean level of non zero velocityclutter.

The filter thresholds are determined by multiplying the mean levels by a constant to achieve the desired P_{fa}. This adaptive threshold on each doppler filter at each range cell provides a constant false alarm rate (CFAR) and results in subclutter visibility (for different radial velocities of aircraft and weather). A digital clutter map establishes the thresholds for zero velocity cells. The map is implemented with one word for each of the 365000 range/azimuth cells. The purpose of the zero velocity filters is to recover the clutter signal and use it to establish a threshold for targets with zero radial velocity.

In each cell of the ground clutter map is stored the average value of the output of the zero velocityfilter for the past 8 scans (32 seconds) On each scan 1/8 of the output is added to 7/8 of the value stored in the map. The map is built up recursively requiring 10 to 20 scans to establish steady state clutter values.

The values in the map are multiplied by a constant to establish a threshold for zero velocity targets. This allows elimination of usual MTI blind speed at zero relative velocity for targets with large cross section. The MTD filters are shown in Fig. 4.28.

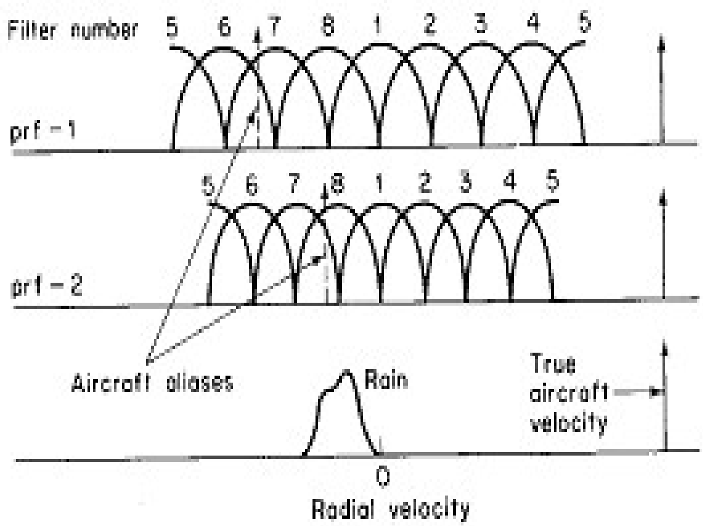


Figure 4.28 Detection of aircraft in rain using two prf's with a doppler filter bank, illustrating the effect of doppler foldover. (*From Muehe,*[43] *Courtesy IEEE.*)

Here the filters 3 through 7 establish the threshold from the mean signal in 16 range cells. Filter 1 establishes the threshold from the clutter map. Filters 2 and 8 adjacent to the zero velocity filter establish the threshold from the larger of the mean signal over 16 range cells and the clutter map.

Fig 4.28 also shows the advantage of using 2 PRFs to detect targets in rain. As shown, the PRFs are 20% different. Typical rain storm spectrum is shown o the bottom and has about 25 to 30 knots width centred any where from - 60 to +60 knots.

An aircraft return is also shown at the right of the figure and is shown as occupying filters 6 and 7 on PRF-1 and filters 7 and 8 on PRF-2 (the difference is due to frequency foldover). With PRF-2 the aircraft velocity is competing with the rain clutter, but at PRF-1 it is clear of clutter. Hence by using 2 PRFs alternating every 10 pulses, aircraft targets will appear in at least one filter free of rain (except when the aircraft's radial velocity is that of the rain).

Interference from other radars may appear as one large return among the ten returns processed. The MTD has an interference eliminator which compares the magnitude of each of the ten pulses against the average magnitude of the ten pulses. If anypulse is greater than 5 times the average, all information from that range/azimuth cell is discarded. The output of the MTD is a hit report containing azimuth, range, amplitude, filter number and PRF.

On one scan, a large aircraft might be reported from more than one doppler filter, from several coherent processing intervals and from adjacent range cells. Up to 20 hit reports might be generated from a single large target.

A post processor groups together all reports which appear to originate from the same target and interpolates to find the best azimuth, range, and amplitude and radar velocity. Target amplitude and Doppler are used to eliminate small cross section and low speed angle echoes before targets are delivered to the automatic tracking circuits.Tracking circuits eliminate false hit reports which do not form logical tracks. The output of the automatic tracker is what is displayed on the PPI.

Since the MTD processor eliminates large amounts of clutter and has a low false detection rate, its output can be remoted via narrow bandwidth telephone.

Limitations of MTI Performance:

Improvement in signal to clutter is affected by factors other than the Doppler spectrum. Instabilities in the transmitter and receiver, physical motion of the clutter, finite time on target and receiver limiting all affect I_c.

Definitions:

1) ***MTI Improvement Factor I:*** The signal to clutter ratio at the output of the MTI system divided by the signal to clutter ratio at the input, averaged over all of the target radial velocities of interest.

2) ***Subclutter Visibility (SCV):*** The ratio by which the target echo may be weaker than the coincident clutter echo power and still be detected with specified Pd and Pfa. All target radial velocities are assumed equally likely

Note: A typical value is 30 dB

Note: Two radars with the same subclutter visibility might not have the same ability to detect targets in clutter if the resolution cell of one is greater and accepts more clutter echo power

3) ***Clutter Visibility Factor Voc:*** The signal to clutter ratio after cancellation (or Doppler processing) that provides the stated P_d and P_{fa}.

4) ***Clutter Attenuation C_A:*** the ratio of the clutter power at the canceler input to the clutter residue at the output, normalized (divided by) to the attenuation of a single pulse passing through the unprocessed channel of the canceler

5) ***Cancellation Ratio:*** The ratio of canceler voltage amplification fro fixed target echoes received with a fixed antenna, to the gain for a single pulse passing through the unprocessed channel of the canceler.

Note: $I_c = (SCV)(V_{oc})$

Note: When the MTI is limited by noise like system instabilities, Voc should be chosen as the SNR for range equation calculations.

Note: when the MTI is limited by antenna scanning fluctuations, let Voc = 6dB for a single pulse.

Note: Once again, I is the preferred measure of MTI performance but does not account for possible poor performance at certain velocities

6) ***Interclutter Visibility:*** The ability of the MTI to detect moving targets in clear resolution cells between patches of strong clutter. Resolution cells can be range, azimuth or Doppler.

Note: the higher the radar resolution, the better the interclutter visibility. A medium resolution radar with 2μs pulse width and 1.5 ˚ beamwidth has sufficient resolution to achieve a 20 dB advantage over low resolution radars

7) ***Equipment Instabilities:*** The apparent frequency spectrum from perfectly stationary clutter can be broadened (and hence will degrade the MTI improvement factor) due to the following:

- Pulse to pulse changes in amplitude
- Pulse to pulse changes in frequency
- Pulse to pulse changes in phase
- Timing jitter on transient pulse
- Variations in time delay through the delay lines
- Changes in pulse width
- Changes in coho or stalo between time of transmit and time of receive

8) ***Consider the effect of phase variations***

Let the echo from a stationary clutter on pulse 1 and pulse 2 be

$$g_1 = A\cos\omega t$$
$$g_2 = A\cos(\omega t + \Delta\phi)$$

Where $\Delta\phi$ is the change in oscillator phase between the two pulses

9) On subtraction

$$g_1(t) - g_2(t) = 2A\sin\left(\frac{\Delta\phi}{2}\right)\sin\left(\omega t + \frac{\Delta\phi}{2}\right) \approx A\Delta\phi\sin\left(\omega t + \frac{\Delta\phi}{2}\right)$$

Hence the limitation on the improvement factor due to oscillator instability is

$$I = \frac{A^2}{(A\Delta\phi)^2} = \frac{1}{(\Delta\phi)^2}$$

Note: To achieve I > 40 dB requires $\Delta\phi < 0.01$ rad from pulse to pulse!. This applies to the coho locking or the phase change introduced by the HPA.

The limits to I imposed by pulse to pulse instability are:

Transmitter frequency	$(\pi\Delta f\tau.)^{-2}$
Stalo or coho frequency	$(2\pi\Delta fT)^{-2}$
Transmitter phase shift	$(\Delta\phi)^{-2}$
Coho locking	$(\Delta\phi)^{-2}$
Pulse timing	$\tau^2/(\Delta t)^2 2B\tau$
Pulse width	$\tau^2/(\Delta\tau)^2 B\tau$
Pulse amplitude	$(A/\Delta A)^2$

where Δf = interpulse frequency change
B = bandwidth
τ = pulse width
T = transmission time
$\Delta\phi$ = interpulse phase change
Δt = timing jitter
$\Delta\tau$ = pulse width jitter
ΔA = interpulse amplitude change

Note: the digital processor does not experience degradation due to timing jitter of the transmit pulse if the clock controlling the processor timing is started from the detected RF envelope of the transmitted pulse.

10) ***Internal Fluctuation of Clutter:*** Absolutely stationary clutter - buildings, water towers, mountains, bare hills

Dynamic clutter - trees, vegetation, sea, rain, chaff

Can limit the performance of MTI

Most fluctuating clutter targets can be represented by a model consisting of many independent scatterers located in the resolution cell.

Any motion of the scatterers relative to the radar results in a different vector sum from pulse to pulse. Fig. 4.29 shows the power spectral density of clutter for a 1 GHz carrier.

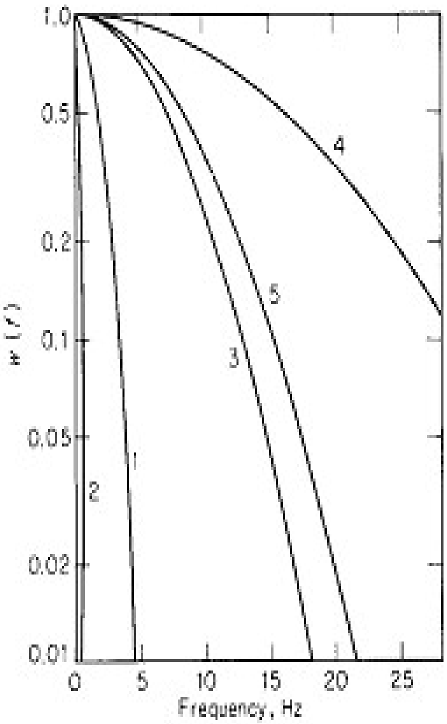


Figure 4.29 Power spectra of various clutter targets. (1) Heavily wooded hills, 20 mi/h wind blowing ($a = 2.3 \times 10^{17}$); (2) sparsely wooded hills, calm day ($a = 3.9 \times 10^{19}$); (3) sea echo, windy day ($a = 1.41 \times 10^{16}$); (4) rain clouds ($a = 2.8 \times 10^{15}$); (5) chaff ($a = 1 \times 10^{16}$). (*From Barlow,*[49] *Proc. IRE.*)

Experimentally measured spectra of clutter can be approximated by

$$W(f) = |g(f)|^2 = |g_0|^2 \exp\left[-a\left(\frac{f}{f_0}\right)^2\right]$$

Where W(f) = clutter power spectrum

g(f) = Fourier transform of the clutter waveform

f_0 = radar carrier frequency and a = a parameter (given by Fig 4.29)

This equation can be rewritten as: $W(f) = W_0 \exp\left[\frac{-f^2}{2\sigma_c^2}\right]$

Where σ_c is the RMS clutter frequency spread in Hz

or as $V(f) = W_0 \exp\left[\frac{-f^2\lambda^2}{8\sigma_v^2}\right]$

Where σ_v is the clutter velocity spread in m/s

Note:

$$\sigma_c = \frac{2\sigma_v}{\lambda}$$

$$a = \frac{c^2}{8\sigma_v^2}$$

The improvement factor can be written as

$$I = \left(\frac{S_0/C_0}{S_i/C_i}\right) = \left(\frac{S_0}{S_i}\right)_{ave} CA$$

Where $C_A = C_i/C_0$ - clutter attenuation averaged over all doppler frequencies

For a single delay line canceler

$$CA = \frac{\int_0^\infty W(f)\,df}{\int_0^\infty W(f)|H(f)|^2\,df}$$

Where H(f) is the frequency response of the canceler

For a single delay line canceler, $h(f) = 1 - \exp(-j2\pi fT)$

$$= 2j\sin(\pi fT)\exp(-j\pi fT)$$

Therefore,

$$CA = \frac{\int_0^\infty W_0(f)\exp[-f^2/2\sigma_c^2]\,df}{\int_0^\infty W_0\exp[-f^2/2\sigma_c^2]4\sin(\pi fT)^2\,df}$$

Assuming $\sigma_c \ll 1/T$ yields, $CA = \dfrac{0.5}{1 - \exp[-2\pi^2 T^2 \sigma_c{}^2]}$

If the exponential is small, we can use the first two terms of its series expansion

$$CA = \frac{f_p^2}{4\pi^2 \sigma_c^{\ 2}}$$

or

$$CA = \frac{f_p^2 \lambda^2}{16\pi^2 \sigma_v^{\ 2}}$$

or

$$CA = \frac{a f_p^2}{2\pi^2 f_0^{\ 2}}$$

Now the average gain $(S_0/S_i)_{avg}$ for the single delay line canceler is 2

Hence, $$I_{1c} = \frac{f_p^2}{2\pi^2 \sigma_c^2} = \frac{f_p^2 \lambda^2}{8\pi^2 \sigma_v^2} = \frac{a f_p^2}{\pi^2 f_0^2}$$

Similarly for a double canceler, whose average gain is 6, $$I_{2c} = \frac{f_p^4}{8\pi^4 \sigma_c^4} = \frac{f_p^4 \lambda^4}{128\pi^4 \sigma_v^4} = \frac{a^2 f_p^4}{2\pi^4 f_0^4}$$

A plot of above equation is shown in Fig 4.30 with $f_p\lambda$ as a parameter. Several representative examples of clutter are indicated at particular values of σ_v.

Note: σ_v (spectral spread in velocity) is with respect to the mean velocity which for ground clutter is zero.

Rain, sea echo and chaff have non zero mean velocity which must be accounted for

Note: the frequency dependence of equations 25 and 26 for the clutter spectrum can not be extended over too great a frequencyrange since no account is taken of variation of cross section of individual scatterers as a function of frequency.

For example the leaves and branches of trees have considerably different reflecting

properties at Ka band where their dimensions are comparable to λ, than at UHF frequencies.

The general form for an N pulse canceler with N_l = N-1 delay lines is

$$I_{NC} = \frac{2^{N_l}}{N_l!}\left(\frac{f_p}{2\pi\sigma_c}\right)^{2N_l}$$

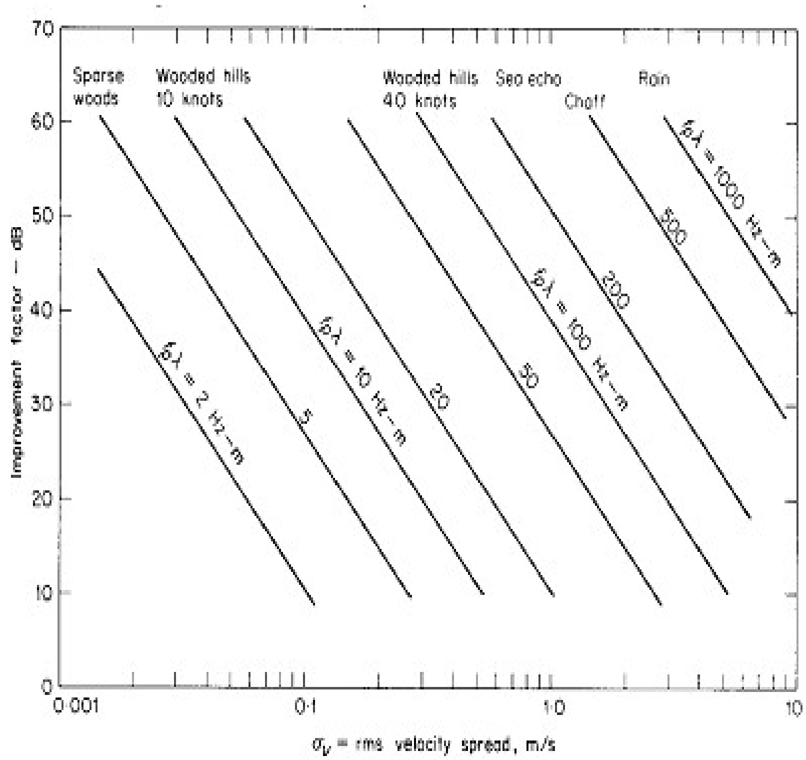


Figure 4.30 Plot of double-canceler clutter improvement factor [Eq. (4.26)] as a function of σ_v = rms velocity spread of the clutter. Parameter is the product of the pulse repetition frequency (f_p) and the radar wavelength (λ).

Antenna Scanning Modulation:

A scanning antenna observes a target for time t_0

$$t_0 = \frac{n_B}{f_p} = \frac{\theta_B}{\dot{\theta}_S}$$

Where n_B is the number of hits received

The received pulse train of finite duration t0 has a frequency spectrum which is proportional t_o. Hence even if the clutter were perfectly stationary, the clutter spectrum would have a finite width due to the finite time on target.

If the clutter spectrum is too wide due to too short an observation interval, the improvement factor will be affected. This limitation is called scanning fluctuation or scanning modulation.

To find the limitation on Ic we first find the clutter attenuation C_A

$$CA = \frac{\int_0^\infty W_S(f)\,df}{\int_0^\infty W_S(f)\,|H(f)|^2\,df}$$

Here $W_S(f)$ describes the spectrum produced by the finite time on target and H(f) is the MTI processor frequency response.

If the antenna main beam is approximated bya Gaussian shape, the spectrum will also be

Gaussian. Hence the results previously derived for a Gaussian clutter spectrum can be applied.

Thus equations I_{1c} and I_{2c} apply for the antenna scanning fluctuations with the correct

interpretation of σ_c (the RMS spread of the spectrum about the mean).

Here the voltage waveform for the clutter is modulated bythe antenna power pattern (equal to the two way field strength pattern), as it is rotated

Now, $$G(\theta) = G_0 \exp\left[\frac{-2.776\theta^2}{\theta_B^2}\right]$$ Where θ and θ_B are in degrees

Therefore

$$S_a = G_0 \exp\left[\frac{-2.776\left(\frac{\theta}{\dot{\theta}_s}\right)^2}{\left(\frac{\theta_B}{\dot{\theta}_s}\right)^2}\right]$$

Where $S_a(t)$ is the modulation of the received signal due to the antenna pattern and is the scan rate in °/s.

$$\text{now } \frac{\theta}{\dot{\theta}_s} = t \text{ and } \frac{\theta_B}{\dot{\theta}_s} = t_0$$

$$\text{hence} \qquad S_a = K \exp\left[\frac{-2.776t^2}{t_0^2}\right]$$

The spectrum is found by taking the Fourier transform

$$S_a = K \int_{-\infty}^{\infty} \exp\left[\frac{-2.776t^2}{t_0^2}\right] \exp[-j2\pi ft]\,dt$$

$$= K_1 \exp\left[\frac{-\pi^2 f^2 t_0^2}{2.776}\right]$$

Since this is a Gaussian function we must have

$$\frac{-\pi^2 f^2 t_0^2}{2.776} = \frac{f^2}{2\sigma_f^2}$$

Hence, $\sigma_f = \frac{1.178}{\pi t_0}$

Note: This is for the voltage spectrum

For the power spectrum we have, $\sigma_s = \frac{\sigma_f}{\sqrt{2}}$

Hence the σs due to the antenna scanning is $\sigma_s = \frac{1}{3.77 t_0}$

Substituting this for σ_c , we get,

$$I_{1s} = \frac{n_B^2}{1.388} \qquad \text{single canceler}$$

and $$I_{2s} = \frac{n_B^4}{3.853}$$

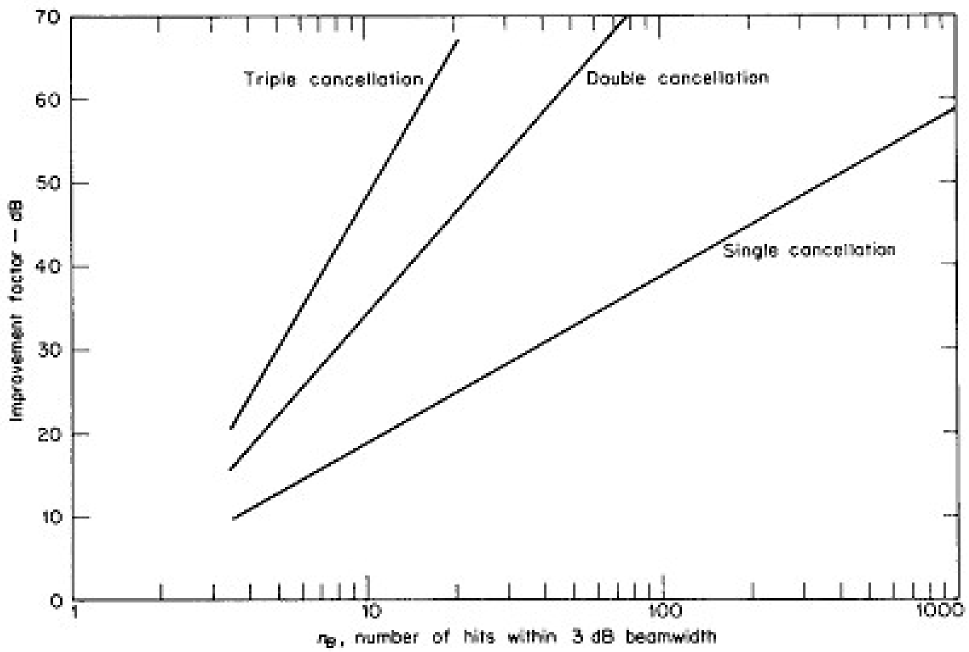


Figure 4.31 Limitation to improvement factor due to a scanning antenna. Antenna pattern assumed to be of gaussian shape.

Note: the stepped scan antenna also is limited in MTI performance by the finite time on target. The time waveform is rectangular which gives a different improvement factor.

Limiting in MTI Radar:

A limiter is used just before the MTI processor to prevent the residue from large clutter echoes from saturating the display. Ideally an MTI radar should reduce clutter to a level comparable to noise. However, when I is not great enough, the clutter residue will appear on the displayand prevent the detection of aircraft whose radar cross section is larger than the clutter residue. This condition can be prevented bysetting the limit L relative to the noise N equal to the MTI improvement factor I.

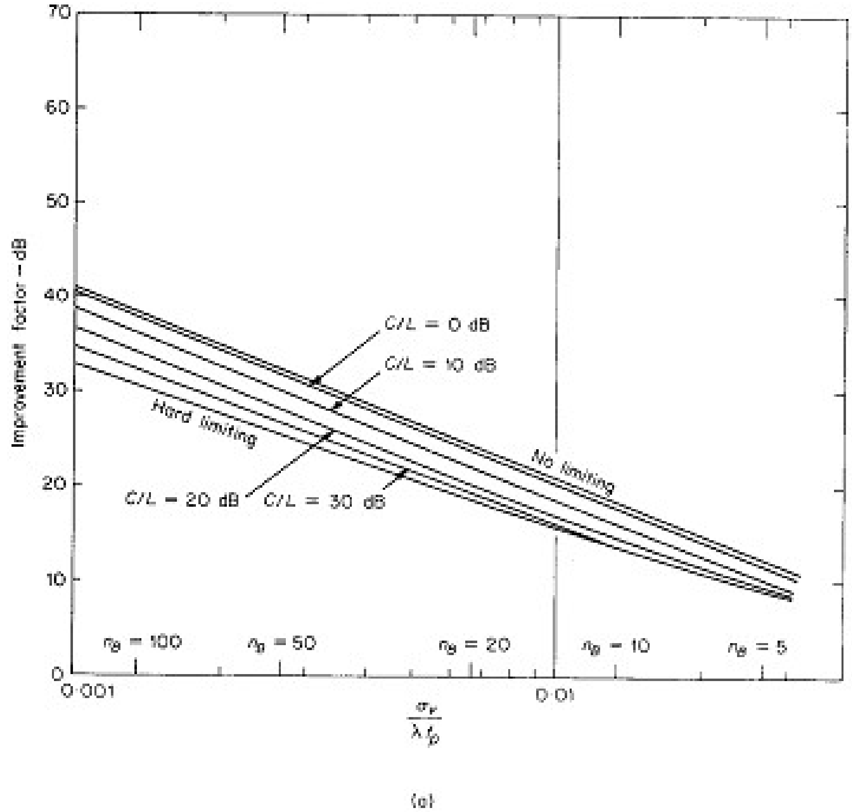
$$\frac{L}{N} = I$$

If the limit le vel is set too high, clutter residue obscures part of the display. If the limit is set too lo w, there may be a black hole effect. The limiter provides a CFAR (Constant False Alarm Rate). A nonlinear limiter however causes the spectrum of strong clutter to spread into the canceler pass band and results in the generation of additional residue which degrades the MTI performance.

Figure 4.32 plots I for 2 pulse and 3 pulse cancelers with various levels of limiting. The abscissa applies to Gaussian clutter spectrum generated by clutter motion (σ_V) or by antenna scanning modulation (n_B). Here C/L is the ratio of RMS clutter power to the IF limit level.

Note: The loss of I increases with the complexity of the canceler. Limiting in a 3 pulse canceler will cause a 15 to 25 dB reduction in performance. A 4 pulse canceler with limiting is only 2 dB better than a 3 pulse canceler. Hence adding complexity is not justified in a limiting environment

Limiting is not needed if processor I is large enough to reduce the largest clutter to the noise level (typically requires I ≈ 60 dB). This is difficult to achieve since it requires the receiver to have a linear dynamic range of at least 60 dB, the A/D must have at least 11 bits, the equipment must be stable, the processor must be designed for I = 60 dB and the number of pulses processed must be sufficient to reduce the antenna scanning modulation.



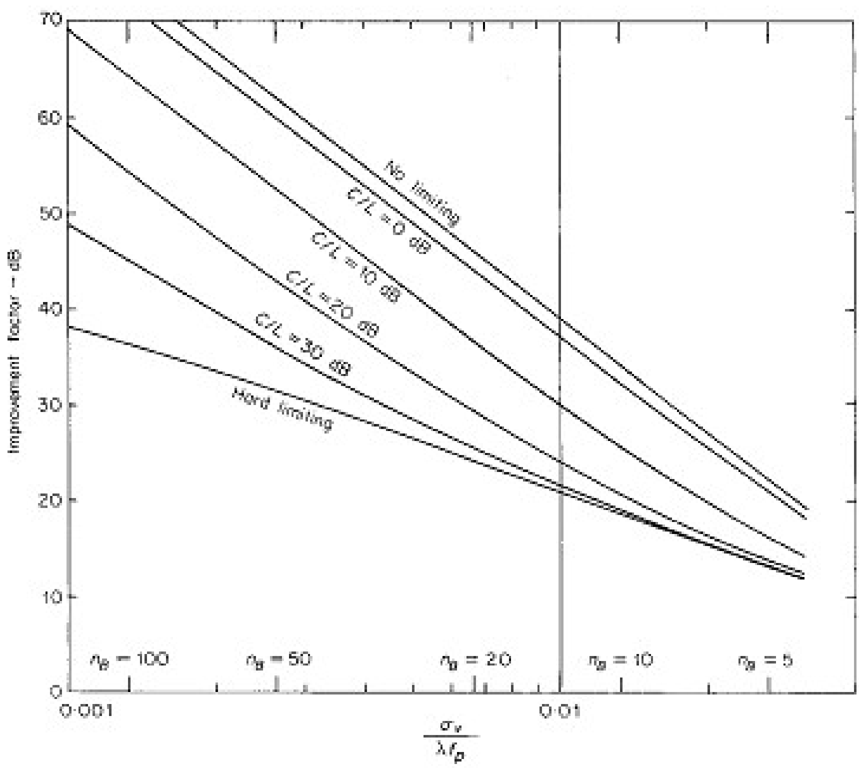


Figure 4.32 Effect of limit level on the improvement factor for (*a*) two-pulse delay-line canceler and (*b*) three-pulse delay-line canceler. C/L = ratio of rms clutter power to limit level. (*From Ward and Shrader,*[53] *Courtesy IEEE.*)

Tracking with Surveillance Radar:

The track of a target can be determined with surveillance radar from the coordinates of the target measured from scan to scan. The quality of such a track depends on the time between observations, the location accuracy of each observation and the number of extraneous targets present in the vicinity of the tracked target.

A surveillance radar which develops tracks on targets it has detected is called a “track while scan” (TWS) radar. Tracks can be obtained by having an operator mark the location of a target on the face of the PPI with a grease pencil on each scan.

A single operator however can not handle more than about 6 target tracks when the radar has a twelve second scan rate. Also an operator’s effectiveness in detecting new targets decreases rapidly after 1/2 hour of operation. These problems are avoided by automating the target detection and tracking process. This is called automatic detection and tracking (ADT)

The ADT performs the following functions:

- Target detection
- Track initiation
- Track association
- Track update
- Track smoothing

- Track termination

The automatic detection quantizes the range into intervals equal to the range resolution. At each range interval the detector integrates n pulses (n is the expected number of hits expected from the target due to the antenna scan rate). The integrated pulses are compared with the threshold to determine the presence or absence of a target.

An example is the “moving window detector” which examines continuouslythe last n samples within each quantized range and announces the presence of a target if m out of n of these samples crosses a preset threshold. By locating the centre of the n pulses, an estimate of the

target’s angular direction can be obtained. This is called “beam splitting”.

If there is only one target present within the radar’s coverage, then the detection on two scans is all that is required to establish a tar get track and estimate its’ v elocity. However there are usuallyother targets plus clutter echoes and hence three or more detections are necessary to establish a track reliably without the generation of false tracks.

A computer can recognize and reject false tracks, however too many false tracks can overload a computer. Hence the radar using ADT should exclude unwanted signals from clutter and interference. A good ADT system therefore requires a good MTI and a good CFAR receiver.

A clutter map generated by the radar is sometimes used to reduce the load on the tracking computer by blanking the clutter areas and removing detections associated with large point clutter sources not rejected by the MTI. Slowly moving echoes that are not of interest can also be removed by the clutter map.

The availability of distinctive target characteristics such as altitude might prove of help when performing track association. When a new detection is received, an attempt is made to associate it with existing tracks. This is aided byestablishing for each track a small search region (gate) within which a new detection is predicted based on the estimate of the target speed and direction. It is desired to make the gate as small as possible to avoid having more than one echo fall within it when traffic density is high, or when two tracks are close. However a large gate is required if the tracker is to follow target maneuvres. Hence more than one gate size may be used.

The size of the gate depends on the accuracy of the track. When a target does not appear in the small gate, a larger gate could be used. The size of the second gate would depend on the maximum acceleration expected of the target. On the basis of past detections, the track while scan radar must make smoothed estimates of the target position and velocity as well as the predicted position and velocity.

One method of computing this information is the $\alpha-\beta$ tracker (also called the g-h tracker). Here the present smoothed target position $\bar{x}$ and velocivty $\bar{v}$ are computed using

$$\bar{x}_n = x_{pn} + \alpha(x_n - x_{pn})$$

$$\bar{v}_n = \overline{v_{n-1}} + \frac{\beta}{T_s}(x_n - x_{pn})$$

Where x_{pn} = the predicted position of the target at the nth scan x_n = the measured position at the nth scan

α = the position smoothing parameter β = the velocity smoothing parameter T_s = the time between observations

The predicted position at the n+1 scan is $x_{pn+1} = \overline{x_n} + \overline{v_n} T_s$

When acceleration is important, a third equation can be added to describe an $\alpha-\beta-\gamma$ tracker where γ is the acceleration smoothing parameter.

If $\alpha = \beta = 0$ then the tracker uses no current data, only smoothed data.

If $\alpha = \beta = 1$ then no smoothing is included at all

The $\alpha-\beta$ filter is designed to minimize the mean square error in the smoothed (filtered) position and velocity assuming small velocity changes between observations. To minimize the output noise variance at steady state and the transient response to a maneuvering target modeled by a ramp function the $\alpha-\beta$ coefficients are related by

$$\beta = \frac{\alpha^2}{2-\alpha}$$

The choice of α between 0 and 1 depends on the system application and in particular on the tracking bandwidth.

A compromise must be made between good smoothing (narrow bandwidth) and rapid response to maneuver (wide bandwidth).

A criterion for selecting the α,β coefficients is based on the best linear track fitted to the radar data in the least squares sense. This gives values

$$\alpha = \frac{2(2n-1)}{n(n+1)} \qquad n>2$$

$$\beta = \frac{6}{n(n+1)} \qquad n>2$$

Where n = the number of scans or target observations

A standard $\alpha - \beta$ tracker does not handle the maneuvering target. An adaptive $\alpha - \beta$ tracker is one which varies the two smoothing parameters to achieve a variable bandwidth so as to be able to follow maneuvres.

The value of α can be set by observing the measurement error x_n - $x_{pn.}$ At the start of the tracking, the bandwidth is set to be wide and is narrowed down if the target moves in a straight line. As the target maneuvres, the bandwidth is widened to keep the tracking error small.

A Kalman filter is similar to an adaptive $\alpha - \beta$ tracker except that it inherently provides for dynamical targets. Here a model for the measurement error has to be assumed, as well a model of the target trajectoryand the disturbance of the trajectory.

Disturbances might be due to neglect of higher order derivatives in the model of the dynamics, random motions due to atmospheric turbulence, and target maneuvres.

Kalman filters can use a wide varietyof models for measurement noise and trajectorydisturbance, however these are usually assumed to be white noise with zero mean.

A maneuvering target does not always fit this model since it produces correlated observations. Proper inclusion of realistic dynamical models increases the complexity of the calculations.

The Kalman filter is sophisticated but costly to implement. Its chief advantage over the classical $\alpha - \beta$ tracker is its inherent ability to account for maneuvre statistics. If Kalman filters were restricted to modelling the target trajectoryas a straight line and if the measurement noise and trajectory disturbance noise were modeled as white Gaussian noise with zero mean, then the Kalman filter equations reduce to $\alpha - \beta$ tracker equations, with the α and β computed sequentially by the Kalman filter procedure.

The classical $\alpha - \beta$ is easy to implement. To handle maneuvering targets, some means maybe included to detect maneuvres and to change the values of α and β. The data rate might also be increased during target maneuvres in some radar systems. As means for choosing α and β become more sophisticated, the optimal $\alpha - \beta$ tracker becomes equivalent to a Kalman filter even for a target trajectory with error. Here the computation of α and β require a knowledge of the statistics of the measurement errors and the prediction errors, and are determined in a recursive manner in that theydepend on previous estimates of the mean square error in the smoothed position and velocity.

Note: The concept of the $\alpha - \beta$ tracker or the Kalman filter apply also to a continuous single target tracker where the error signal is processed digitally. The equations describing the $\alpha - \beta$ tracker are equivalent to a Type II servo system

Of the track while scan radar does not receive target information on a particular scan, the smoothing and predicting operations can be continued by accounting for the missing data. When data to update a track is missing for a sufficient number of consecutive scans, the track is terminated.

Example: if the criterion for establishing a track is 3 target reports, 5 consecutive misses is suitable for termination.

ADT effectively reduces the bandwidth in the radar output allowing the data to be transmitted to a remote site via narrow band phone lines. This allows the outputs from several radars to be communicated to a central control point economically.

The adaptive thresholding of the automatic detector can cause worsening of the range resolution. It would seem that two targets might be resolved if their separation is about 0.8 pulse width. However with automatic detection, the probabilityof resolving targets in range onlye xceeds 0.9 if theyare separated by 2.5 pulse widths.

To achieve this resolution, a log video receiver should be used and the threshold should be proportioned to the smaller of the two means calculated from a number of reference cells on either side of the test cell. This resolution limit assumes that the shape of the return pulse is not known. (If it were, it would be possible to resolve targets within one pulse width).

When more than one radar are covering the same volume of space, it is sometimes desirable to combine their outputs to form a single track file (Automatic Detection and Integrated Tracking, ADIT). ADIT has the advantage of a greater data rate than a single radar and reduces the likelihood of a loss of target detections caused by antenna lobing, fading, interference and clutter. The integrated processing allows a favourable weighting of better data and lesser weighting of the poorer data.

OBJECTIVE TYPE QUESTIONS

1. Radar range primarily depends upon [ ]

a. peak transmitted power b. average transmitted

power c. independent of transmitted power d. resolution of radar

2. In radar system which of the following is used for transmitter output tubes [ ]

a. parameter amplifier b. RC coupled amplifier

c. klystron only d. magnetron or travelling wave tube

3. A radar is to have maximum range of 60km. The maximum allowable pulse repetition frequency for unambiguous reception should be [ ]

a. 25 pps b. 250 pps c. 2500 pps d. 25,000 pps

4. An MTI radar operates at 10 GHz with PRF of 3000pps. The lowest blind speed will be

90km/hr

a. 40 km/hr b. 66 km./hr c. 81 km/hr d. 162 km/hr

5. In which of the following case the lowest blind speed will be 90 km/hr [ ]

a. frequency 1 GHz an PRF 300 pps b. frequency 3 GHz and PRF 500 pps c.

frequency 5 GHz and PRF 700 pps d. frequency 7 GHz and PRF 1000 pps

6. If a given maximum range of radar is to be doubled, all other factors remaining constant the peak power must be increased [ ]

a. four fold b. eight fold c. sixteen fold d. thirty fold

7. The maximum range of a radar depends on all of the following except [ ]

a. peak transmitted pulse power b. direction of movement of

target c. target area d. capture are

8. The antenna used for radar is [09M03] [ ]

a. paraboloidal antenna b. isotropic radiator

c. resonant antenna d. whip antenna

9. Noise figure of radar receiver is 12 dB and its bandwidth is 2.5 MHz. the value of

Pmin for this radar will be [ ]

a. 1.59 X 10^{-9} watt b. 1.59 X 10^{-13} watt c. 1.59 X 10^{-15}watt d. 1.59 X 10^{-17} watt

10. IN radar system the lobe switching technique is used to [09S01] [ ]

a. scan the area b. move antenna in the direction fo the object

c. locate the target accurately d. move the weapon in the required direction

11. Radar display is [ ]

a. A scope display `b. PPI c. MTI d. CRO

12. The minimum receivable signal in radar receiver which has an IF bandwidth of 1.5

MHz and is 9 dB noise figure will be [ ]

a. 4.17 X 10 -10 watt b. 4.17 X 10 -12 watt

c. 4.17 X 10-14 watt d. 4.17 X 10-16 watt

13. The advance and retard switches for the circuit are needed due to [ ]

a. the gate generators being unstable b. scanner and deflection coil misalignment c. in accuracies in the power supply d. failure of the switch circuits

14. When real transmitted power on a radar system, is increased by a factor of 16,. The maximum range will be increased by a factor of [ ]

a. 2 b. 4 c. 8 d. 16

15. The signal arriving from the transmitter to the display unit is the [ ]

a. trigger b. echoes c. heading marker d.bearing information

16. The delay unit section of the VRM/ delay unit can be used to [ ]

a. extend the range of radar` b. reduce radar

interference c. expand an area for examination

d. extend the range of radar and expand an area for examination

17. The trigger circuit [ ]

a. is a switch connecting high voltage through to

magnetron b. is a master timing device of the radar

c. is microwave frequency oscillator

d. receives bearing information from the scanner.]

18. An AFC system produces a control voltage to control the --- frequency [ ]

a. Magnetrons b. Local oscillator's

c. PRF oscillator's d. Tabulator

19. The local oscillator's frequency is [ ]

a. 60 MHz above the echo frequency b. 60 MHz below the echo frequency

c. 30 MHz above the echo frequency d. 30 MHz below the echo frequency

20Sea clutter returns occur [10M05] [ ]

a. due to reflections from rain clouds b. at short ranges

c. due to land reflections d. due to satellite reflections

Answers:

1.a	2.d	3.c	4.d	5.b	6.c	7.b	8.a	9.b	10.c
11.a	12.c	13.b	14.a	15.a	16.d	17.b	18.b	19.b	20.b

ESSAY TYPE QUESTIONS

1. What is a delay line canceller? Illustrate the concept of blind speeds based on the frequency response of a single delay line canceller.
2. Discuss the factors limiting the performance of an MTI system.
3. What are blind speeds? Suggest a method to reduce the effect of blind speeds for unambiguous detection of a moving target.
4. Calculate the lowest blind speed of an MTI system operating at 3.6 cm wave length and transmitting at a pulse repetition time of 330 µS.
5. Explore the possibility of broadening the clutter rejection null using a second dlay line canceller in the MTI radar system.
6. Describe automatic tracking of a target through range gating technique for unambiguous detection of a moving target.
7. Calculate the lowest blind speed of an MTI system operating at 4.2 cm wave length and transmitting at a pulse repetition time of 286 µS.
8. Explore the possibility of broadening the clutter rejection null using a second delay line canceller in the MTI radar system.
9. With the help of necessary blo ck diagram explain the operation of an MTI radar system with a power amplifier in the transmitter.
10. Compare and contrast the situations with a power amplifier and a power oscillator in the transmitter of an MTI system
11. Describe the method of staggering pulse repetition frequency to reduce the effect of blind speeds in an MTI system.

12. Explain the following limitations of MTI radar.

(a) Equipment instabilities.

(b) Scanning modulation.

(c) Internal fluctuation of clutter.

13. MTI radar is operated at 9GHz with a PRF of 3000 pps. Calculate the first two lowest blind speeds for this radar. Derive the formula used.

14. Discuss the liminations of non-cohorent MTI Radar systems.

15. Write the description of Range gate Doppler filters.

16. Explain the operation of MTI radar with 2 pulse repetition frequencies.

17. Explain the function of time domain filter with an example.

18. MTI radar operates at 10GHz with a PRF of 300 pps. Calculate the lowest blind speed?

UNIT-VI

TRACKING

RADAR

Explain tracking principles:

A tracking-radar system

(1) Measures the coordinates of a target and

(2) Provides data which may be used to determine the target path and to predict its future position.

All or only part of the available radar data-range, elevation angle, azimuth angle, and Doppler frequency shift may be used in predicting future position; that is, radar might track in range, in angle, in doppler, or with any combination. Almost any radar can be considered tracking radar provided its output information is processed properly. But, in general, it is the method by which angle tracking is accomplished that distinguishes what is normal normally considered tracking radar from any other radar. It is also necessary to distinguish between continuous tracking radar and a track-while-scan (TWS) radar.

The continuous tracking radar supplies continuous tracking data on a particular target, while the track-while-scan supplies sampled data on one or more targets. In general, the continuous tracking radar and the TWS radar employ different types of equipment.

The antenna beam in the continuous tracking radar is positioned in angle by a servomechanism actuated by an error signal. The various methods for generating the error signal may be classified as sequential lobing, conical scan, and simultaneous lobing or monopulse. The range and Doppler frequency shift can also be continuously tracked, if desired, by a servocontrol loop actuated by an error signal generated in the radar receiver.

Explain sequential lobing:

The antenna pattern commonly employed with tracking radars is the symmetrical pencil beam in which the elevation and azimuth beamwidths are approximately equal. However, a simple pencil-beam antenna is not suitable for tracking radars unless means are provided for determining the magnitude and direction of the target's angular position with respect to some reference direction, usually the axis of the antenna. The difference between the target position and the reference direction is the angular error. The tracking radar attempts to position the antenna to make the angular error zero. When the angular error is zero, the target is located along the reference direction.

One method of obtaining the direction and the magnitude of the angular error in one coordinate is by alternately switching the antenna beam between two positions (Fig.). This is called lobe switching, sequential switching, or sequential lobing. Fig (a) is a polar representation of the antenna beam (minus the sidelobes) in the two switched positions. A plot in rectangular coordinates is shown in Fig (b), and the error signal obtained from a target not on the switching axis (reference direction) is shown in Fig (c).

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 sign of the difference determines the direction the antenna must be moved in order to align the switching axis with the direction of the target. When the voltages in the two switched positions are equal, the target is on axis and, its position may be determined from the axis direction.

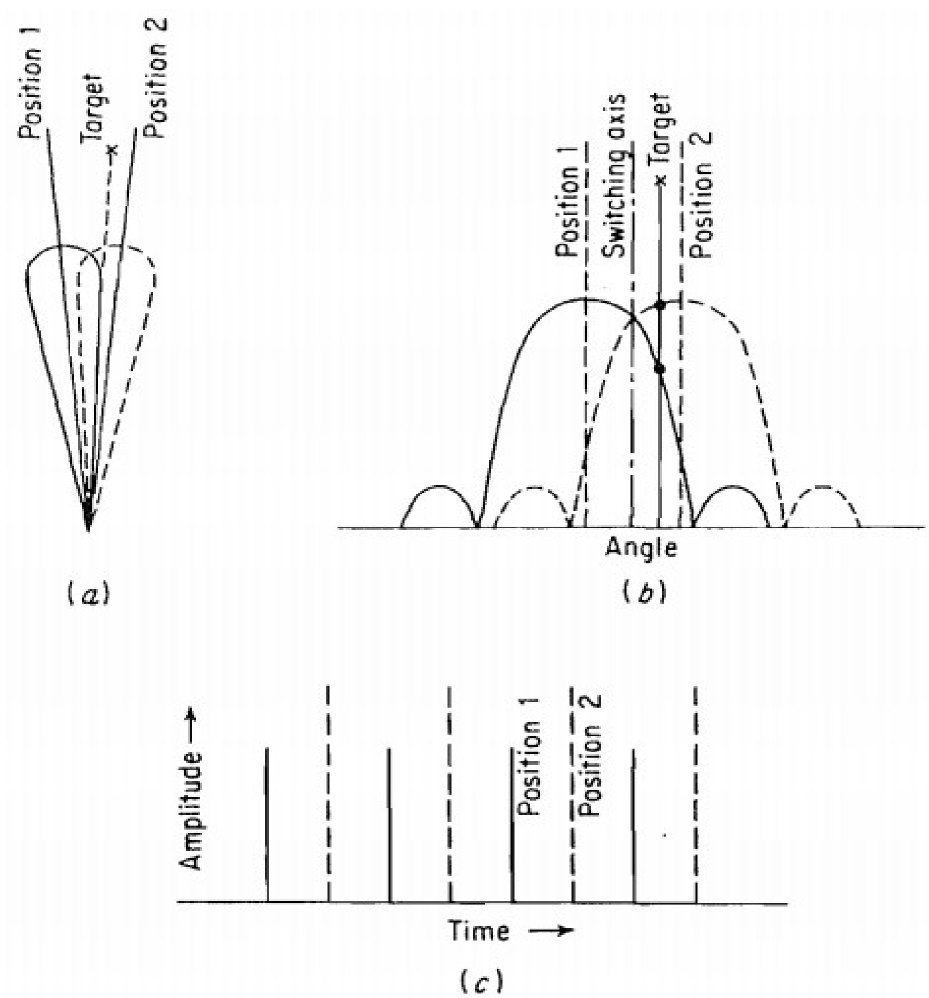


Fig: Lobe-switching antenna patterns and error signal (one dimension). (a) Polar representation of switched antenna patterns (b) rectangular representation (c) error signal.

Two additional switching positions are needed to obtain the angular error in the orthogonal coordinate. Thus a two-dimensional sequentially lobing radar might consist of a cluster of four feed horns illuminating a single antenna, arranged so that the right-left, up-down sectors are covered by successive antenna positions. Both transmission and reception are accomplished at each position. A cluster of five feeds might also be employed, with the central feed used for transmission while the outer four feeds are used for receiving. High-power RF switches are not needed since only the receiving beams, and not the transmitting beam, are stepped in this five-feed arrangement.

One of the limitations of a simple unswitched nonscanning pencil-beam antenna is that the angle accuracy can be no better than the size of the antenna beamwidth. An important feature of sequential lobing (as well as the other tracking techniques to be discussed) is that the targetposition accuracy can be far better than that given by the antenna beamwidth. The accuracy depends on how well equality of the signals in the switched positions can be determined. The fundamental limitation to accuracy is system noise caused either by mechanical or electrical fluctuations.

Sequential lobing, or lobe switching, was one of the first tracking-radar techniques to be employed. Early applications were in airborne-interception radar, where it provided directional information for homing on a target, and in ground-based antiaircraft fire-control radars. It is not used as often in modern tracking-radar applications.

Explain conical scanning method:

The logical extension of the sequential lobing technique is to rotate continuously an offset antenna beam rather than discontinuously step the beam between four discrete positions. This is known as conical scanning (Fig). The angle between the axis of rotation (which is usually, but not always, the axis of the antenna reflector) and the axis of the antenna beam is called the squint angle.

Consider a target at position A. The echo signal will be modulated at a frequency equal to the rotation frequency of the beam. The amplitude of the echo-signal modulation will depend upon the shape of the antenna pattern, the squint angle and the angle between the target line of sight and the rotation axis. The phase of the modulation depends on the angle between the target and the rotation axis. The conical scan modulation is extracted from the echo signal and applied to a servo-control system which continually positions the antenna on the target. When the antenna is on target, as in B of Fig, the line of sight to the target and the rotation axis coincide, and the conical-scan modulation is zero.

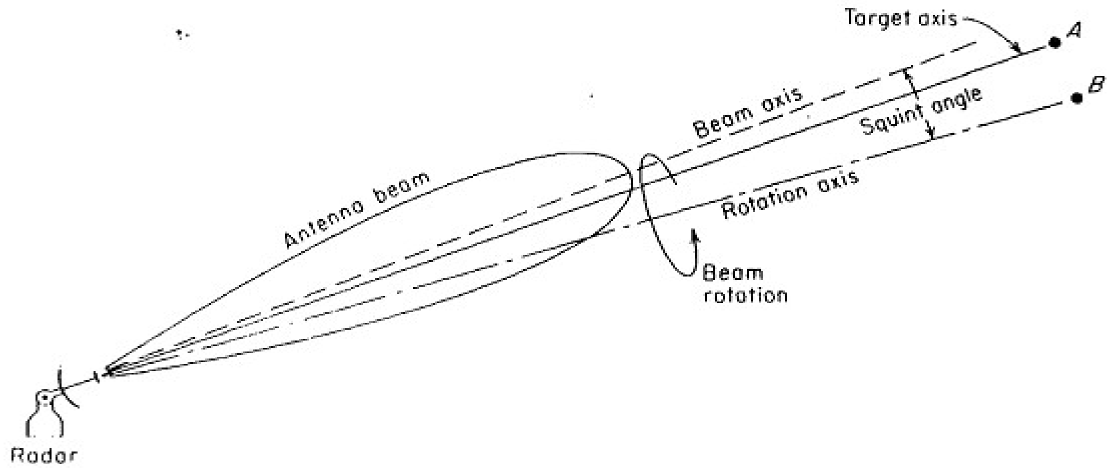


Fig: Conical-scan tracking

A block diagram of the angle-tracking portion of typical conical-scan tracking radar is shown in Fig. The antenna is mounted so that it can be positioned in both azimuth and elevation by separate motors, which might be either electric- or hydraulic-driven. The antenna beam is offset by tilting either the feed or the reflector with respect to one another.

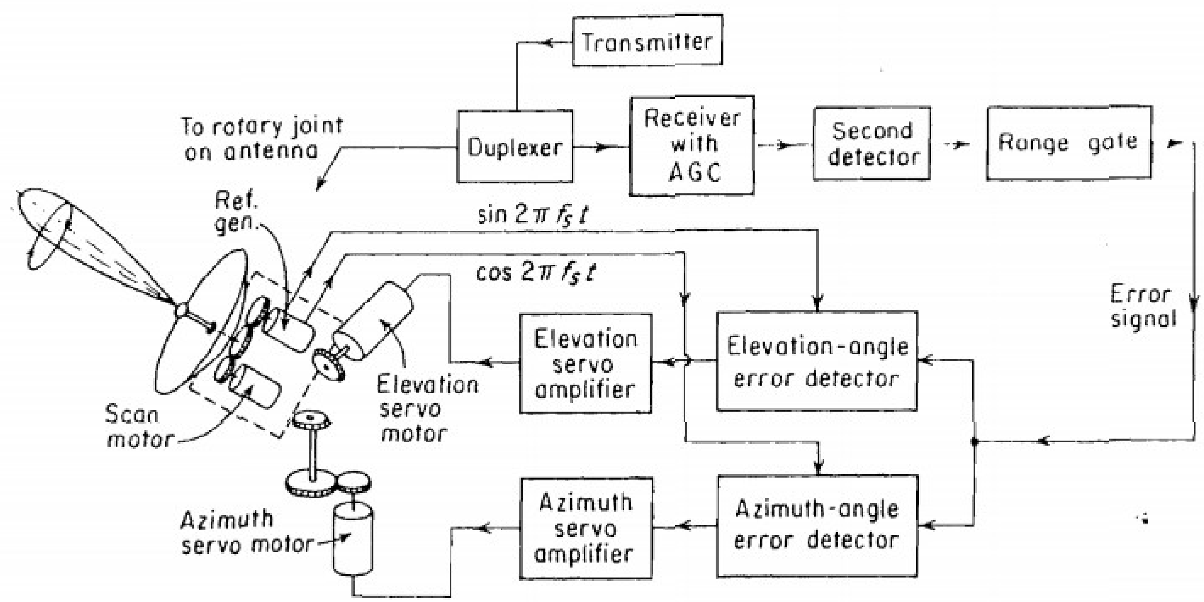


Fig: Block diagram of conical-scan tracking radar

One of the simplest conical-scan antennas is a parabola with an offset rear feed rotated about the axis of the reflector. If the feed maintains the plane of polarization fixed as it rotates, it is called a nutating feed. A rotating feed causes the polarization to rotate. The latter type of feed requires a rotary joint. The nutating feed requires a flexible joint. If the antenna is small, it may be easier to rotate the dish, which is offset, rather than the feed, thus avoiding the problem of a rotary or flexible RF joint in the feed.

A typical conical-scan rotation speed might be 30 r/s. The same motor that provides the conical-scan rotation of the antenna beam also drives a two phase reference generator with two outputs 90° apart in phase. These two outputs serve as a reference to extract the elevation and azimuth errors. The received echo signal is fed to the receiver from the antenna via two rotary joints (not shown in the block diagram). One rotary joint permits motion in azimuth, the other, in elevation.

The receiver is a conventional superheterodyne except for features peculiar to the conical scan tracking radar. One feature not found in other radar receivers is a means of extracting the conical-scan modulation, or error signal. This is accomplished after the second detector in the video portion of the receiver.

The error signal is compared with the elevation and azimuth reference signals in the angle-error detectors, which are phase-sensitive detectors. A phasesensitive detector is a nonlinear device in which the input signal (in this case the angle-error signal) is mixed with the reference signal. The input and reference signals are of the same frequency. The output d-c voltage reverses polarity as the phase of the input signal changes through 180°.

The magnitude of the d-c output from the angle-error detector is proportional to the error, and the sign (polarity) is an indication of the direction of the error. The angle-errordetector outputs are amplified and drive the antenna elevation and azimuth servo motors.

The angular position of the target may be determined from the elevation and azimuth o the antenna axis. The position can be read out by means of standard angle transducers such as synchros, potentiometers, or analog-to-digital-data converters.

Explain the block diagram of the AGC portion of tracking radar receiver:

The echo-signal amplitude at the tracking-radar receiver will not be constant but will vary with time. The three major causes of variation in amplitude are (1) the inverse-fourth-power relationship between the echo signal and range, (2) the conical- scan modulation (angle-error signal), and (3) amplitude fluctuations in the target cross section.

The function of the automatic gain control (AGC) is to maintain the d-c level of the receiver output constant and to smooth or eliminate as much of the noise like amplitude fluctuations as possible without disturbing the extraction of the desired error signal at the conical-scan frequency.

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

An example of the AGC portion of a tracking-radar receiver is shown in Fig. 6.4. A portion of the video-amplifier output is passed through a low-pass or smoothing filter and fed back to control the gain of the IF amplifier. The larger the video output, the larger will be the feedback signal and the greater will be the gain reduction.

The filter in the AGC loop should pass all frequencies from direct current to just below the conical-scan-modulation frequency. The loop gain of the AGC filter measured at the conical-scan frequency should be low so that the error signal will not be affected by AGC action.

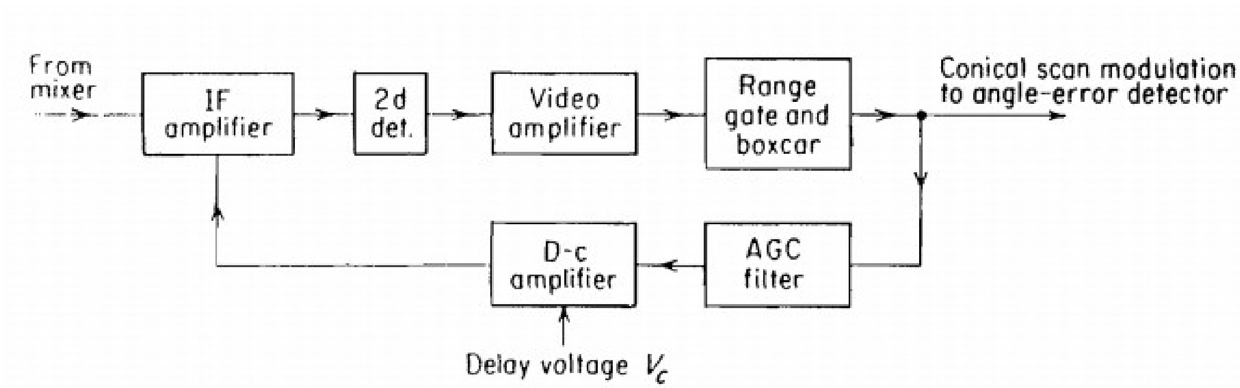


Fig: Block diagram of the AGC portion of a tracking-radar receiver

The output of the feedback loop will be zero unless the feedback voltage exceeds a prespecified minimum value V_c. In the block diagram the feedback voltage and the voltage V_c are compared in the d-c amplifier. If the feedback voltage exceeds V_c, the AGC is operative, while if it is less, there is no AGC action. The voltage V_c is called the delay voltage. The purpose of the delay voltage is to provide a reference for the constant output signal and permit receiver gain for weak signals. If the delay voltage were zero, any output which might appear from the receiver would be due to the failure of the AGC circuit to regulate completely.

In many applications of AGC the delay voltage is actually zero. This is called undelayed AGC. In such cases the AGC can still perform satisfactorily since the loop gain is usually low for small signals. Thus the AGC will not regulate weak signals. The effect is similar to having a delay voltage, but the performance will not be as good.

Explain the Block diagram of amplitude-comparison monopulse radar for single angular coordinate and explain its operation:

The amplitude-comparison monopulse employs two overlapping antenna patterns (Fig (a)) to obtain the angular error in one coordinate. The two overlapping antenna beams may be generated with a single reflector or with a lens antenna illuminated by two adjacent feeds. (A cluster of four feeds may be used if both elevation- and azimuth-error signals are wanted.) The sum of the two antenna patterns of Fig (a) is shown in Fig (b), and the difference in Fig (c). The sum patterns are used for transmission, while both the sum pattern and the difference pattern are used on reception.

The signal received with the difference pattern provides the magnitude of the angle error. The sum signal provides the range measurement and is also used as a reference to extract the sign of the error signal. Signals received from the sum and the difference patterns are amplified separately and combined in a phase-sensitive detector to produce the error-signal characteristic shown in Fig (d).

A block diagram of the amplitude-comparison-monopulse tracking radar for a single angular coordinate is shown in Fig. The two adjacent antenna feeds are connected to the two arms of a hybrid junction such as a magic T, a rat race, or a short-slot coupler. The sum and difference signals appear at the two other arms of the hybrid. On reception, the outputs of the sum arm and the difference arm are each heterodyned to an intermediate frequency and amplified as in any superheterodyne receiver.

The transmitter is connected to the sum arm. Range information is also extracted from the sum channel. A duplexer is included in the sum arm for the protection of the receiver. The output of the phase-sensitive detector is an error signal whose magnitude is proportional to the angular error and whose sign is proportional to the direction.

The output of the monopulse radar is used to perform automatic tracking. The angular error signal actuates a servo-control system to position the antenna, and the range output from the sum channel feeds into an automatic-range-tracking unit.

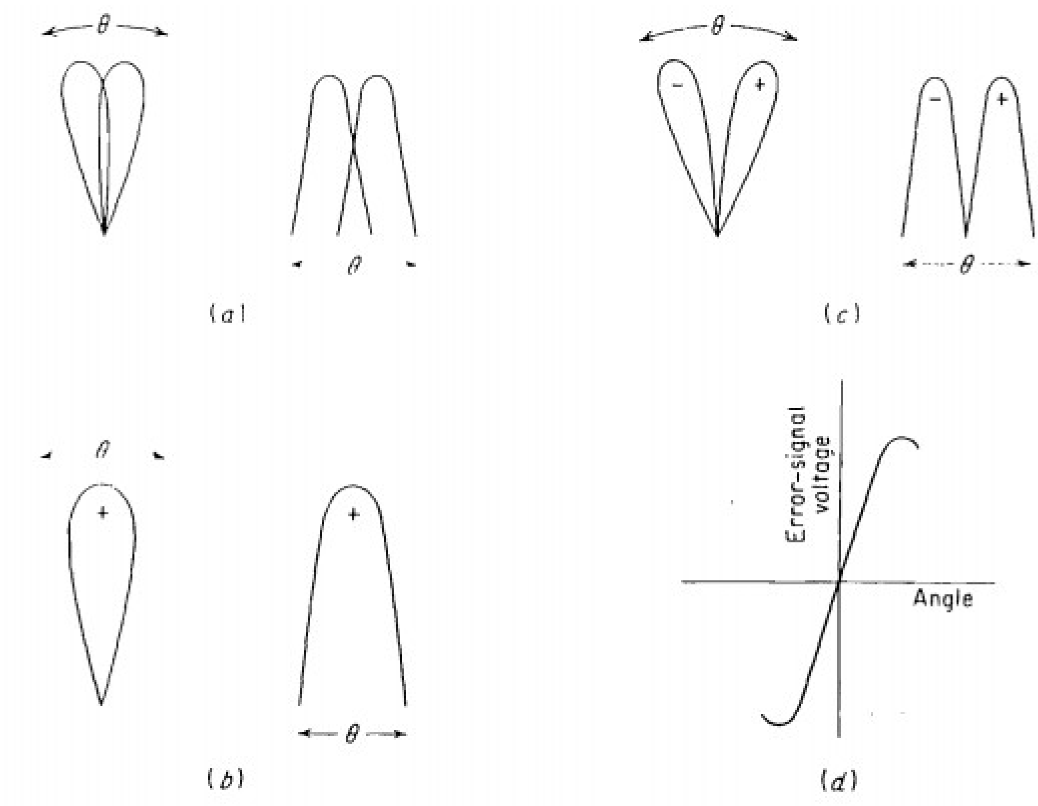


Fig: Monopulse antenna patterns and error signal

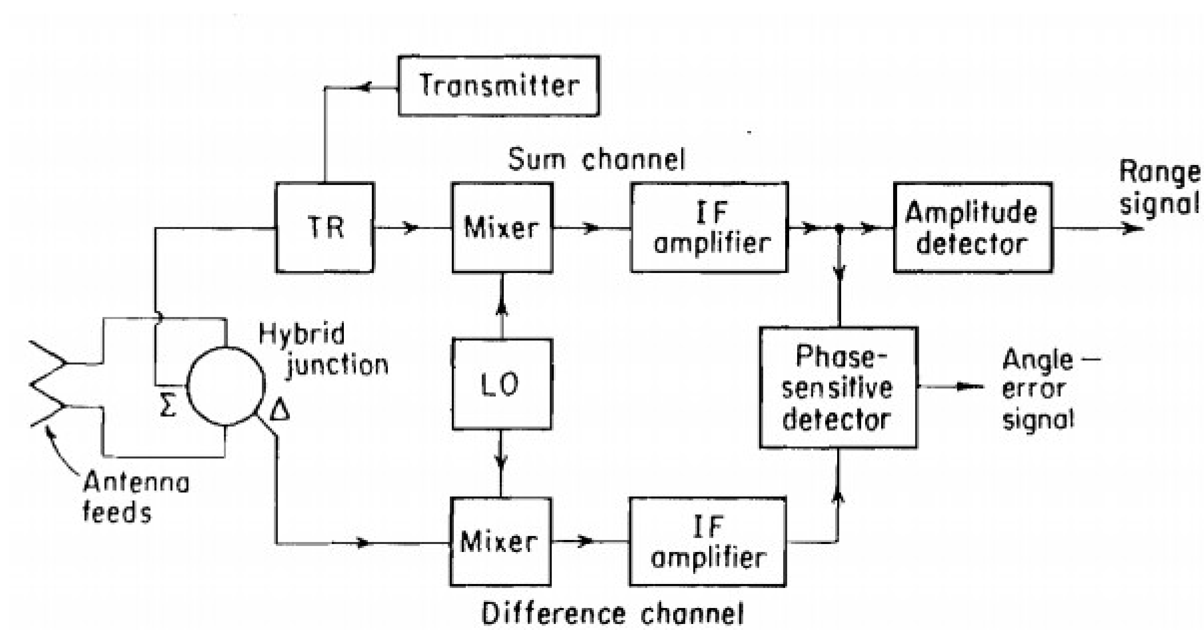


Fig: Block diagram of amplitude-comparison monopulse radar

The sign of the difference signal (and the direction of the angular error) is determined by comparing the phase of the difference signal with the phase of the sum signal. If the sum signal in the IF portion of the receiver were As cos ωIF t, the difference signal would be either - A_d cos (ω_{IF}t) or – A_d cos (ω_{IF}t) (A_s > 0, A_d >0), depending on which side of center is the target. Since A_d

cos ($\omega_{IF}t$) = – A_d cos [$\omega_{IF}(t + \pi)$], the sign of the difference signal may be measured by determining whether the difference signal is in phase with the sum or 1800 out of phase.

Explain the Block diagram of amplitude-comparison monopulse radar for extracting error signals in both elevation and azimuth:

A block diagram of monopulse radar with provision for extracting error signals in both elevation and azimuth is shown in Fig. The cluster of four feeds generates four partially overlapping antenna beams. The feeds might be used with a parabolic reflector, Cassegrain antenna, or a lens. All four feeds generate the sum pattern. The difference pattern in one plane is formed by taking the sum of two adjacent feeds and subtracting this from the sum of the other two adjacent feeds. The difference pattern in the orthogonal plane is obtained by adding the differences of the orthogonal adjacent pairs. A total of four hybrid junctions generate the sum channel, the azimuth difference channel, and the elevation difference channel. Three separate mixers and IF amplifiers are shown, one for each channel. All three mixers operate from a single local oscillator in order to maintain the phase relationships between the three channels. Two phase-sensitive detectors extract the angle-error information, one for azimuth, the other for elevation. Range information is extracted from the output of the sum channel after amplitude detection.

Since a phase comparison is made between the output of the sum channel and each of the difference channels, it is important that the phase shifts introduced by each of the channels be almost identical. The phase difference between channels must be maintained to within 25° or better for reasonably proper performance. The gains of the channels also must not differ by more than specified amounts.

The purpose in using one- or two-channel monopulse receivers is to ease the problem associated with maintaining identical phase and amplitude balance among the three channels of the conventional receiver. Two-channel monopulse receivers have also been used by combining the sum and the two difference signals in a manner such that they can be again resolved into three components after amplification.

The approximately "ideal" feed-illuminations for a monopulse radar is shown in Fig. This has been approximated in some precision tracking radars by a five-horn feed consisting of one horn generating the sum pattern surrounded by four horns generating the difference patterns.

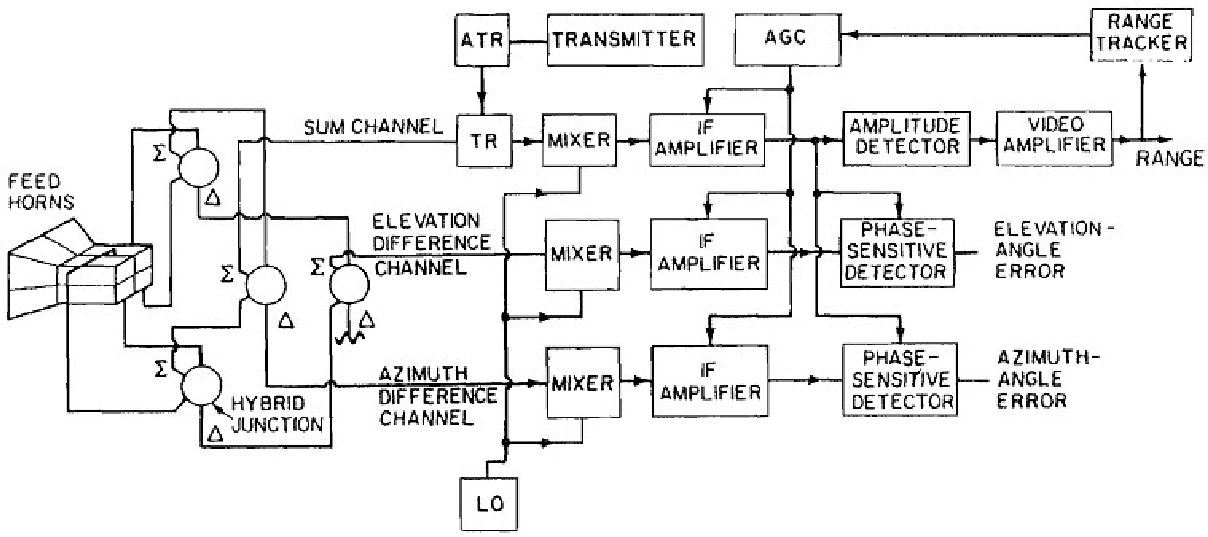


Fig: Block diagram of two-coordinate (azimuth and elevation) amplitude-comparison monopulse tracking radar

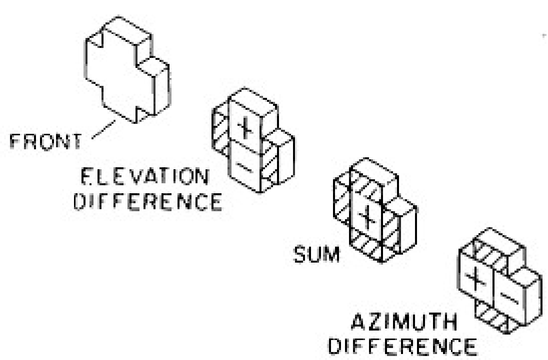


Fig: approximately ideal feed-aperture illumination for monopulse sum and difference channels

Explain Phase-comparison monopulse tracking radar technique:

The measurement of angle of arrival by comparison of the phase relationships in the signals from the separated antennas of a radio interferometer has been widely used by the radio astronomers for precise measurements of the positions of radio stars. The interferometer as used by the radio astronomer is a passive instrument, the source of energy being radiated by the target itself. Tracking radar which operates with phase information is similar to an active interferometer and might be called an interfero meter radar. It has also been called Simultaneous phase comparison radar, or phase-comparison monopulse.

In Fig. two antennas are shown separated by a distance d. The distance to the target is R

and is assumed large compared with the antenna separation d. The line of sight to the target makes an angle θ to the perpendicular bisector of the line joining the two antennas. The distance

from antenna 1 to the target is

$$R_1 = R+(d\sin\theta)/2$$

And the distance from antenna 2 to the target is

$$R_2 = R-(d\sin\theta)/2$$

The phase difference between the echo signals in the two antennas is approximately

$$\Delta\phi = 2\pi d \sin\theta/\lambda$$

For small angles where $\sin\theta = 0$, the phase difference is a linear function of the angular error and

may be used to position the antenna via a servo-control loop.

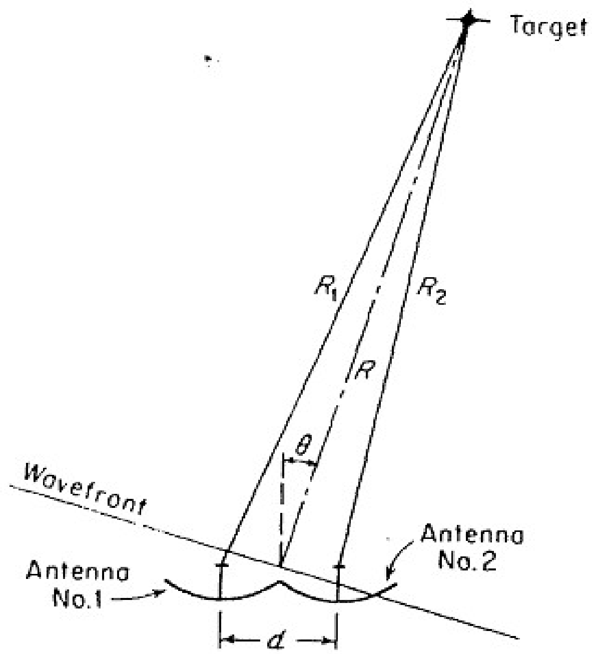


Fig: Wavefront phase relationships in phase- comparison monopulse radar

In the early versions of the phase-comparison monopulse radar, the angular error was determined by measuring the phase difference between the outputs of receivers connected to each antenna.

The output from one of the antennas was used for transmission and for providing the range information. With such an arrangement it was difficult to obtain the desired aperture illuminations and to maintain a stable boresight. A more satisfactory method of operation is to form the sum and difference patterns in the RF and to process the signals as in conventional amplitude-comparison monopulse radar.

Explain how tracking in range is achieved using split range gates:

The technique for automatically tracking in range is based on the split range gate. Two range gates are generated as shown in Fig. One is the early gate, and the other is the late gate. The echo pulse is shown in Fig., the relative position of the gates at a particular instant in Fig., and the error signal in Fig. The portion of the signal energy contained in the early gate is less than that in the late gate. If the outputs of the two gates are subtracted, an error signal will result which may be used to reposition the center of the gates. The magnitude of the error signal is a measure of the difference between the center of the pulse and the center of the gates. The sign of the error signal determines the direction in which the gates must be repositioned by a feedback- control system. When the error signal is zero the range gates are centered on the pulse.

The range gating necessary to perform automatic tracking offers several advantages as by products. It isolates one target excluding targets at other ranges. This permits the boxcar generator to be employed. Also range gating improves the signal-to-noise ratio since it eliminates the noise from the other range intervals. Hence the width of the gate should be sufficiently narrow to minimize extraneous noise. On the other hand, it must not be so narrow that an appreciable fraction of the signal energy is excluded. A reasonable compromise is to make the gate width of the order of the pulse width.

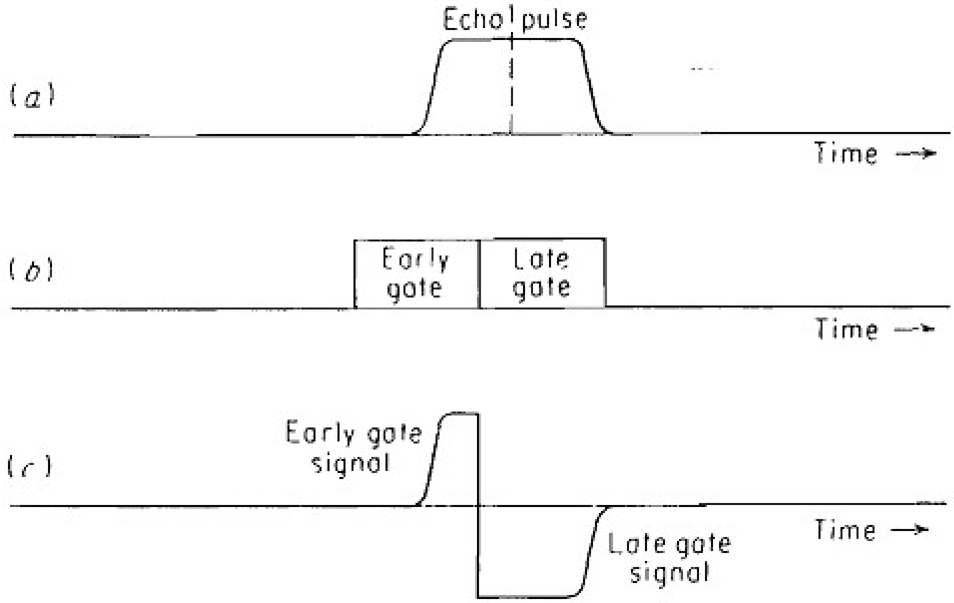


Fig: Split-range-gate tracking (a) Echo pulse; (h) early-late range gates; (c) difference signal between early and late range gates.

A target of finite length can cause noise in range-tracking circuits in an analogous manner to angle-fluctuation noise (glint) in the angle-tracking circuits. Range-tracking noise depends or the length of the target and its shape. It has been reported that the rms value of the range noise is approximately 0.8 of the target length when tracking is accomplished with a video split-range- gate error detector.

What are the various methods of acquisition before tracking a target with Radar?

A tracking radar must first find and acquire its target before it can operate as a tracker. Therefore it is usually necessary for the radar to scan an angular sector in which the presence of the target is suspected. Most tracking radars employ a narrow pencil-beam antenna. Examples of the common types of scanning patterns employed with pencil-beam antennas are illustrated in

Fig.

In the helical scan, the antenna is continuously rotated in azimuth while it is simultaneously raised or lowered in elevation. It traces a helix in space. Helical scanning was employed for the search mode of the SCR-584 fire-control radar, developed during World War II for the aiming of antiaircraft-gun batteries. The SCR-584 antenna was .rotated at the rate of 6 rpm and covered a 20" elevation angle in 1 min.

The Palmer scan derives its name from the familiar penmanship exercises of grammar school days. It consists of a rapid circular scan (conical scan) about the axis of the antenna, combined with a linear movement of the axis of rotation. When the axis of rotation is held stationary, the Palmer scan reduces to the conical scan. Because of this property, the Palmer scan is sometimes used with conical-scan tracking radars which must operate with a search as well as a track mode since the same mechanisms used to produce conical scanning can also be used for Palmer scanning.

Some conical-scan tracking radars increase the squint angle during search in order to reduce the time required to scan a given volume. The conical scan of the SCR-584 was operated during the search mode and was actually a Palmer scan in a helix. In general, conical scan is performed during the search mode of most tracking radars.

The Palmer scan is suited to a search area which is larger in one dimension than another. The spiral scan covers an angular search volume with circular symmetry. Both the spiral scan and the Palmer scan suffer from the disadvantage that all parts of the scan volume do not receive the same energy unless the scanning speed is varied during the scan cycle. As a consequence, the number of hits returned from a target when searching with a constant scanning rate depends upon the position of the target within the search area.

The raster, or TV, scan, unlike the Palmer or the spiral scan, paints the search area in a uniform manner. The raster scan is a simple and convenient means for searching a limited sector, rectangular in shape. Similar to the raster scan is the nodding scan produced by oscillating the antenna beam rapidly in elevation and slowly in azimuth. Although it may be employed to cover a limited sector-as does the raster scan-nodding scan may also be used to obtain hemispherical coverage, that is, elevation angle extending to 900 and the azimuth scan angle to 3600.

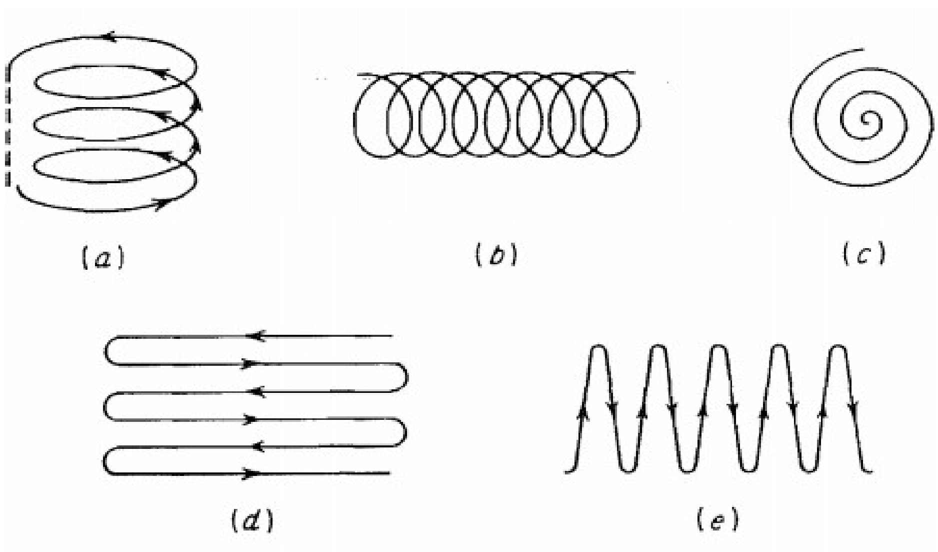


Fig: Examples of acquisition search patterns. (a) Trace of helical scanning beam; (b) Palmer scan; (c) spiral scan; (d) raster, or TV, scan; (e) nodding scan.

The helical scan and the nodding scan can both be used to obtain hemispheric coverage with a pencil beam. The nodding scan is also used with height-finding radars. The Palmer, spiral, and raster scans are employed in fire-control tracking radars to assist in the acquisition of the target when the search sector is of limited extent.

Explain in detail about the limitations to tracking accuracy:

Main limitations to tracking accuracy of radar are,

1. Amplitude fluctuations**.**

2. Angle fluctuations.

3. Receiver and servo noise

Amplitude fluctuations:

A complex target such as an aircraft or a ship may be considered as a number of independent scattering elements. The echo signal can be represented as the vector addition of the contributions from the individual scatterers. If the target aspect changes with respect to the radar- as might occur because of motion of the target, or turbulence in the case of aircraft targets-the relative phase and amplitude relationships of the contributions from the individual scatterers also change. Consequently, the vector sum, and therefore the amplitude change with changing target

aspect.

Amplitude fluctuations of the echo signal are important in the design of the lobeswitching radar and the conical-scan radar but are of little consequence to the monopulse tracker. Both the conical-scan tracker and the lobe-switching tracker require a finite time to obtain a measurement of the angle error.

To reduce the effect of amplitude noise on tracking, the conical-scan frequency should be chosen to correspond to a low value of amplitude noise. If considerable amplitude fluctuation noise were to appear at the conical-scan or lobe-switching frequencies, it could not be readily eliminated with filters or AGC. A typical scan frequency might be of the order of 30 Hz. Higher frequencies might also be used since target amplitude noise generally decreases with increasing frequency.

It has been found experimentally that the tracking accuracy of radars operating with pulse repetition frequencies from 1000 to 4000 Hz and a lobing or scan rate one-quarter of the prf are not limited by echo amplitude fluctuations.

Angle fluctuations:

Changes in the target aspect with respect to the radar can cause the apparent center of radar reflections to wander from one point to another. (The apparent center of radar reflection is the direction of the antenna when the error signal is zero.) In general, the apparent center of reflection might not correspond to the target center.

The random wandering of the apparent radar reflecting center gives rise to noisy or jittered angle tracking. This form of tracking noise is called angle noise, angle scintillations, angle fluctuations, or target glint. The angular fluctuations produced by small targets at long range may be of little consequence in most instances. However, at short range or with relatively large targets (as might be seen by a radar seeker on a homing missile), angular fluctuations may be the chief factor limiting tracking accuracy. Angle fluctuations affect all tracking radars whether conical-scan, sequential lobing, or monopulse.

Receiver and servo noise:

Another limitation on tracking accuracy is the receiver noise power. The accuracy of the angle measurement is inversely proportional to the square root of the signal-to-noise power ratio. Since the signal-to-noise ratio is proportional to 1/R4 (from the radar equation), the angular error due to receiver noise is proportional to the square of the target distance. Servo noise is the hunting action of the tracking servomechanism which results from backlash and compliance in the gears, shafts, and structures of the mount. The magnitude of servo noise is essentially independent of the target echo and will therefore be independent of range.

Explain low angle tracking:

Radar that tracks a target at a low elevation angle, near the surface of the earth, can receive two echo signals from the target, Fig. 6.11. One signal is reflected directly from the target, and the ther arrives via the earth's surface. The direct and the surface-reflected signals combine at the radar to yield angle measurement that differs from the true measurement that would have been made with a single target in the absence of surface reflections. The result is an error in the measurement of elevation. The surface-reflected signal may be thought of as originating from the image of the target mirrored by the earth's surface. Thus, the effect on tracking is similar to the two-target model used to describe glint. The surface-reflected signal is sometimes called a multipath signal.

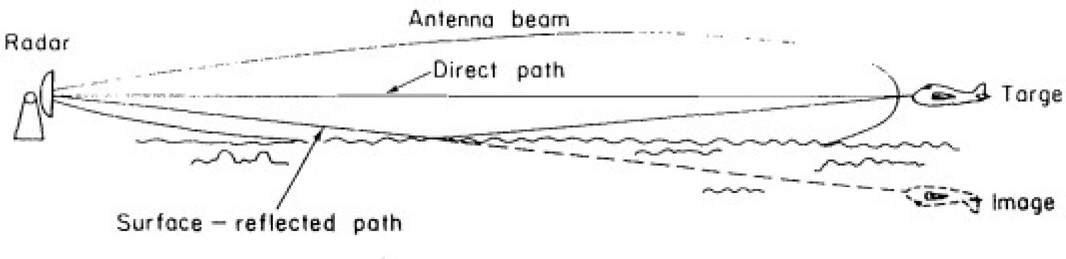


Fig: Low-angle tracking

The surface-reflected signal travels a longer path than the direct signal so that it may be possible in some cases to separate the two in time (range). Tracking on the direct signal avoids the angle errors introduced by the multipath. The range-resolution required to separate the direct from the ground-reflected signal is

Where,

h_a = radar antenna
height, h_t = target height,

R = range to the target.

For a radar height of 30 m, a target height of 100 m and a range of 10 km, the

rangeresolution must be 0.6 m, corresponding to a pulse width of 4 ns. This is a much shorter pulse than is commonly employed in radar. Although the required range-resolutions for groundbased radar are achievable in principle, it is usually not applicable in practice. The use of frequency diversity can also reduce the multipath tracking error.

OBJECTIVE TYPE QUESTIONS

1. One method commonly employed to extract Doppler information in a form suitable for display on the PPI scope is with a [ ]

   a. Power Amplifier b. A-Scope display

   c. Delay line canceller d. Coherent oscillator

2. The characteristic feature of coherent MIT Radar is that the [ ]

   a. Transmitted signal must be out of phase with reference signal in receiver

   b. The transmitted signal must be equal in the magnitude with reference signal

   c. The transmitted signal must be coherent with the reference signal in the receiver d. Transmitted signal must not be equal to reference signal in the receiver

3. In the following which are produce, with time a butterfly effect on the 'A' scope [ ]

a. Fixed Targets b. PPI scope c. Moving Targets d.Phase Detector

4. The stalo, coho and the mixer in which they are combined plus any low-level amplification are called the [ ]

a. Transmitter-Oscillator b. Transmitter-Exciter c. Receiver-Amplifier d.Receiver-exciter

5. The Doppler frequency shift produced by a moving target may be used in pulse radar

to a. Combine moving targets from desired stationery objects

b. Determine the relative velocity of a target

c. Separate desired moving targets from desired stationery

objects d. Determine the displacement of a target

6. To operate with unambiguous Doppler pulse repetition frequency is usually [ ]

a. Low b. Very low c. High d. Very High

7. MTI stands for [ ]

a. Moving Transmitter Indicator b. Moving target interval

c. Moving target indication d. Modulation Transmitting Interval

8. Echoes from fixed targets [ ]

a. Vary in amplitude b. Vary in frequency c.
Vary in pulse interval d. Remains constant

9. The limitation of pulse MTI radar which do not occur with cw radar [ ]

a. Blind speeds b. Delay lines

c. Requires more operating powers d. Requires complex circuitry

10. The presence of blind speeds within the Doppler frequency band reduces the [ ]

a. Output of the radar b. Detection capabilities of the radar

c. Unambiguous range d. Ambiguous range

11. The capability of delay line canceller depends on the [ ]

a. Quality of signal b. Pulse Interval

c. Quality of the medium used as the delay line. d. Delay time of the delay line

12. The output of the MTI receiver phase detector be quantized into a sequence of digital words by Using [ ]

a. Digital quantizer b. Digital Phase detector c. Digital delay lines d. Digital filter

13. A transmitter which consists of a stable low power oscillator followed by a power amplifier is called

a. POMA b. MOPA c. MTI Radar d. CW radar

14. A simple MTI delay line canceller is an example of a [ ]

a. Frequency domain filters b. High pass filter

c. Active filter d. Time domain filter

15. The delay line must introduce a time delay equal to the [ ]

a. Time interval b. Pulse repetition interval c. Pulse width d. Phase shift

16. The delay line canceller [ ] a. Rejects the ac component of clutter b. Rejects the dc component of clutter c. Allows the ac as well as dc d. It rejects all components

17. In pulse MTI radar, Doppler is measured by [ ]

a. Continuous signals b. discrete samples c. Constant period d.Constant amplitude

18. The output of the two single delay line cancellers in cascade is the [ ]

a. Double of that from a single canceller b. U times that from a single canceller

c. Square of that from a single canceller d. Same as that from a single canceller

19. To operate MTI radar with high pulse repetition frequencies [ ]

a. λ fp must be small b. λ fp must be unity c. λ fp must be zero d. λ fp must be large

20. If the first blind speed were 600 knots, the maximum unambiguous range would be ----- at a frequency of 300MHz [ ]

a. 140 nautical miles b. 600 knots c. 140 knots d. 130 nautical miles

Answers:

1.c	2.d	3.c	4.d	5.b	6.c	7.c	8.d	9.a	10.b
11.c	12.c	13.b	14.d	15.b	16.b	17.b	18.c	19.d	20.d

ESSAY TYPE QUESTIONS

1. Discuss the e ect of surface quality and reaction characteristics of a target on the angular tracking accuracy of a tracking radar.
2. Describe the phase comparison mono pulse tracking technique in a radar system with the help of necessary block diagram.
3. With the help of a suitable block diagram, Sequential lobing type of tracking technique in a tracking radar system.
4. Compare and contrast conical scan and sequential lobing type tracking techniques
5. Describe the pro cess of acquiring a moving target prior to tracking it along with the patterns used for acquisition.
6. Describe automatic tracking of a target through range gating technique.

7. Describe sequential lobing type of error signal generation to track a target automatically.

8. List the merits and demerits of monopulse tracker over conical scan type tracker.

9. Draw the block diagram of an amplitude comparison monopulse tracking radar in azimuth and elevation directions. Explain the functioning of this two dimensional tracking radar.

10. Write the simpli er version of radar range equation and explain how this equation does not adequately describe the performance of practical radar?

11. What are the speci c bands assigned by the ITU for the Radar? What the corresponding frequencies?

12. A low power, short range radar is solid-state throughout, including a low-noise RF ampli er which gives it an overall noise gure of 4.77dB. If the antenna diameter is 1m, the IF

bandwidth is 500kHz, the operating frequency is 8GHz and the radar set is supposed to be capable of detecting targets of 5m2 cross sectional area at a maximum distance of 12 km, what must be the peak transmitted pulse power?

13. The average false alarm time is a more signi cant parameter than the false-alarm probability. Give the reasons.

14. Why post detection integration is not as e cient as pre-detection integration of radar pulses?

15. Draw and explain block diagram of Conical-scan tracking radar.

16. Why does a tracking radar have poor accuracy at low elevation angles? Explain

17. Draw and explain the following with respect to Tracking in range:

i. Echo pulse

ii. Early-late range gates

iii. Difference signal between early and late range gates.

18. Limitation of automatic detection and tracking.

19. Explain the block diagram of amplitude comparison mono pulse for extracting error signals in both elevation and azimuth.

20. With diagrams explain Split-range-gate tracking.

UNIT-VII

DETECTION OF RADAR SIGNALS IN NOISE

Matched-Filter Receiver:

A network whose frequency-response function maximizes the output peak-signal-to-mean-noise (power) ratio is called a matched filter. This criterion, or its equivalent, is used for the design of almost all radar receivers.

The frequency-response function, denoted H(f), expresses the relative amplitude and phase of the output of a network with respect to the input when the input is a pure sinusoid. The magnitude |H(f) | of the frequency-response function is the receiver amplitude passband characteristic.

If the bandwidth of the receiver passband is wide compared with that occupied by the signal energy, extraneous noise is introduced by the excess bandwidth which lowers the output signal-to-noise ratio. On the other hand, if the receiver bandwidth is narrower than the bandwidth occupied by the signal, the noise energy is reduced along with a considerable part of the signal

energy.

The net result is again a lowered signal-to-noise ratio. Thus there is an optimum bandwidth at which the signal-to-noise ratio is a maximum. This is well known to the radar receiver designer.

The rule of thumb quoted in pulse radar practice is that the receiver bandwidth B should be approximately equal to the reciprocal of the pulse width τ. This is a reasonable approximation for pulse radars with conventional superheterodyne receivers. It is not generally valid for other waveforms, however, and is mentioned to illustrate in a qualitative manner the effect of the receiver characteristic on signal-to-noise ratio.

The exact specification of the optimum receiver characteristic involves the frequency-response function and the shape of the received waveform.

The receiver frequency-response function, is assumed to apply from the antenna terminals to the output of the IF amplifier. (The second detector and video portion of the well-designed radar superheterodyne receiver will have negligible effect on the output signal-to-noise ratio if the receiver is designed as a matched filter.) Narrow banding is most conveniently accomplished in the IF.

The bandwidths of the RF and mixer stages of the normal superheterodyne receiver are usually large compared with the IF bandwidth. Therefore the frequency-response function of the portion of the receiver included between the antenna terminals to the output of the IF amplifier is taken to be that of the IF amplifier alone. Thus we need only obtain the frequency-response function that maximizes the signal-to-noise ratio at the output of the IF. The IF amplifier may be considered as a filter with gain. The response of this filter as a function of frequency is the property of interest. For a received waveform s(t) with a given ratio of signal energy E to noise energy No (or noise power per hertz of bandwidth), North showed that the frequency-response function of the linear, time-invariant filter which maximizes the output peak-signal-to-mean-noise (power) ratio for a fixed input signal-tonoise (energy) ratio is

$$H(f) = G_a S^*(f) \exp(-j2\pi f t_1)$$

where $S(f) = \int_{-\infty}^{\infty} s(t) \exp(-j2\pi ft)\, dt$ = voltage spectrum (Fourier transform) of input signal

$S^*(f)$ = complex conjugate of $S(f)$

t_1 = fixed value of time at which signal is observed to be maximum

G_a = constant equal to maximum filter gain (generally taken to be unity)

The noise that accompanies the signal is assumed to be stationary and to have a uniform spectrum (white noise). It need not be gaussian. The filter whose frequency-response function is given by Eq. above has been called the North filter, the conjugate filter, or more usually the matched filter. It has also been called the Fourier transform criterion. It should not be confused with the circuit-theory concept of impedance matching, which maximizes the power transfer rather than the signal-to-noise ratio.

The frequency-response function of the matched filter is the conjugate of the spectrum of the received waveform except for the phase shift exp (- j2Πf$_{t1}$). This phase shift varies uniformly with frequency. Its effect is to cause a constant time delay. A time delay is necessary in the specification of the filter for reasons of physical realizability since there can be no output from the filter until the signal is applied.

The frequency spectrum of the received signal may be written as an amplitude spectrum

|S(f) | (and a phase spectrum exp [- jϕ_s (f)]. The matched- filter frequency-response function may similarly be written in terms of its amplitude and phase spectra |H(f) | and exp [-jϕ_m(f)]. Ignoring the constant G_a, Eq. above for the matched filter may then be written as

$$|H(f)| \exp[-j\phi_m(f)] = |S(f)| \exp\{j[\phi_s(f) - 2\pi f t_1]\}$$

or

$$|H(f)| = |S(f)|$$

and

$$\phi_m(f) = -\phi_s(f) + 2\pi f t_1$$

Thus the amplitude spectrum of the matched filter is the same as the amplitude spectrum of the signal, but the phase spectrum of the matched filter is the negative of the phase spectrum of the signal plus a phase shift proportional to frequency.

The matched filter may also be specified by its impulse response h(t), which is the inverse Fourier transform of the frequency-response function.

$$h(t) = \int_{-\infty}^{\infty} H(f) \exp(j2\pi ft)\, df$$

Physically, the impulse response is the output of the filter as a function of time when the input is an impulse (delta function).

$$h(t) = G_a \int_{-\infty}^{\infty} S^*(f) \exp[-j2\pi f(t_1 - t)]\, df$$

Since $S^*(f) = S(-f)$, we have

$$h(t) = G_a \int_{-\infty}^{\infty} S(f) \exp[j2\pi f(t_1 - t)]\, df = G_a s(t_1 - t)$$

A rather interesting result is that the impulse response of the matched filter is the image of the received waveform; that is, it is the same as the received signal run backward in time starting from the fixed time t_1. Figure 1 shows a received waveform s(t) and the impulse response h(t) of its matched filter. The impulse response of the filter, if it is to be realizable, is not defined for t< 0. (One cannot have any response before the impulse is applied.) Therefore we must always have $t < t_1$. This is equivalent to the condition placed on the transfer function H(f) that there be a phase shift exp $(-j2\Pi f_{t1})$. However, for the sake of convenience, the impulse response of the matched filter is sometimes written simply as s(- t).

Derivation of the matched-filter characteristic:

The frequency-response function of the matched filter has been derived by a number of authors using either the calculus of variations or the Schwartz inequality. We shall derive the matched-filter frequency-response function using the Schwartz inequality.

We wish to show that the frequency-response function of the linear, time-invariant filter which maximizes the output peak-signal-to-mean-noise ratio is

$$H(f) = G_a S^*(f) \exp(-j2\pi f t_1)$$

When the input noise is stationary and white (uniform spectral density). The ratio we wish to maximize is

$$R_f = \frac{|s_o(t)|^2_{max}}{N}$$

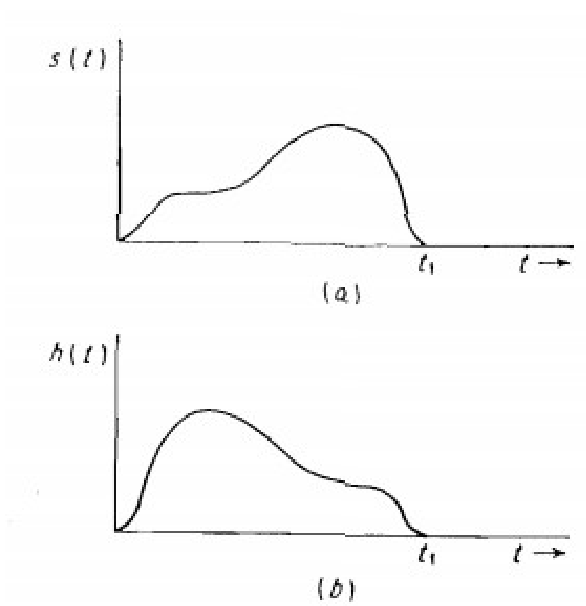


Fig.1 (a) Received waveform s(t); (b) impulse response h(t) of the matched filter.

Where $|s_o(t)|_{max}$ = maximum value of output signal voltage and N = mean noise power at receiver output. The ratio R_f is not quite the same as the signal-to-noise ratio which has been considered in the radar equation. The output voltage of a filter with frequency-response function H(f) is

$$|s_o(t)| = \left| \int_{-\infty}^{\infty} S(f)H(f) \exp(j2\pi ft)\, df \right|$$

Where S(f) is the Fourier transform of the input (received) signal. The mean output noise power is

$$N = \frac{N_o}{2} \int_{-\infty}^{\infty} |H(f)|^2\, df$$

Where N_o is the input noise power per unit bandwidth. The factor appears before the integral because the limits extend from - ∞ to +∞, whereas No is defined as the noise power per cycle of bandwidth over positive values only. Assuming that the maximum value of $|s_0(t)|^2$ occurs at time $t = t_1$, the ratio R_f becomes

$$R_f = \frac{\left| \int_{-\infty}^{\infty} S(f) H(f) \exp(j2\pi f t_1)\, df \right|^2}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2\, df}$$

Schwartz's inequality states that if P and Q are two complex functions, then

$$\int P^* P\, dx \int Q^* Q\, dx \geq \left| \int P^* Q\, dx \right|^2$$

The equality sign applies when P = kQ, where k is a constant. Letting

$$P^* = S(f) \exp(j2\pi f t_1) \quad \text{and} \quad Q = H(f)$$

and recalling that

$$\int P^* P\, dx = \int |P|^2\, dx$$

We get, on applying the Schwartz inequality to the numerator of Eq. earlier, we get

$$R_f \leq \frac{\int_{-\infty}^{\infty} |H(f)|^2\, df \int_{-\infty}^{\infty} |S(f)|^2\, df}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2\, df} = \frac{\int_{-\infty}^{\infty} |S(f)|^2\, df}{\frac{N_0}{2}}$$

From Parseval's theorem,

$$\int_{-\infty}^{\infty} |S(f)|^2\, df = \int_{-\infty}^{\infty} s^2(t)\, dt = \text{signal energy} = E$$

Therefore we have

$$R_f \leq \frac{2E}{N_0}$$

The frequency-response function which maximizes the peak-signal-to-mean-noise ratio

R_f may be obtained by noting that the equality sign in Eq. applies when P = kQ, or

$$H(f) = G_a S^*(f) \exp(-j2\pi f t_1)$$

Where the constant k has been set equal to $1/G_a$.

Relation between the matched filter characteristics and correlation function:

The matched filter and the correlation function. The output of the matched filter is not a replica of the input signal. However, from the point of view of detecting signals in noise, preserving the shape of the signal is of no importance. If it is necessary to preserve the shape of the input pulse rather than maximize the output signal-to-noise ratio, some other criterion must be employed.

The output of the matched filter may be shown to be proportional to the input signal cross-correlated with a replica of the transmitted signal, except for the time delay t_1. The crosscorrelation function R(t) of two signals y(λ) and s(λ), each of finite duration, is defined as

$$R(t) = \int_{-\infty}^{\infty} y(\lambda)s(\lambda - t)\, d\lambda$$

The output $y_o(t)$ of a filter with impulse response h(t) when the input is $y_{in}(t) = s(t) + n(t)$ is

$$y_0(t) = \int_{-\infty}^{\infty} y_{in}(\lambda)h(t - \lambda)\, d\lambda$$

If the filter is a matched filter, then $h(\lambda) = s(t_1 - \lambda)$ and Eq. above becomes

$$y_0(t) = \int_{-\infty}^{\infty} y_{in}(\lambda)s(t_1 - t + \lambda)\, d\lambda = R(t - t_1)$$

Thus the matched filter forms the cross correlation between the received signal corrupted by noise and a replica of the transmitted signal. The replica of the transmitted signal is "built in" to the matched filter via the frequency-response function. If the input signal yin (t) were the same as the signal s(t) for which the matched filter was designed (that is, the noise is assumed negligible), the output would be the autocorrelation function. The autocorrelation function of a rectangular pulse of width τ is a triangle whose base is of width 2τ.

Efficiency of non-matched filters:

In practice the matched filter cannot always be obtained exactly. It is appropriate, therefore, to examine the efficiency of non matched filters compared with the ideal matched filter. The measure of efficiency is taken as the peak signal-to-noise ratio from the non matched filter divided by the peak signal-to-noise ratio ($2E/N_o$) from the matched filter. Figure. Plots the efficiency for a single-tuned (RLC) resonant filter and a rectangular-shaped filter of half-power bandwidth B_τ when the input is a rectangular pulse of width τ. The maximum efficiency of the single-tuned filter occurs for $B_\tau \approx 0.4$. The corresponding loss in signal-to-noise ratio is 0.88 dB as compared with a matched filter.

Table lists the values of B_τ which maximize the signal-to-noise ratio (SNR) for various combinations of filters and pulse shapes. It can be seen that the loss in SNR incurred by use of these non-matched filters is small.

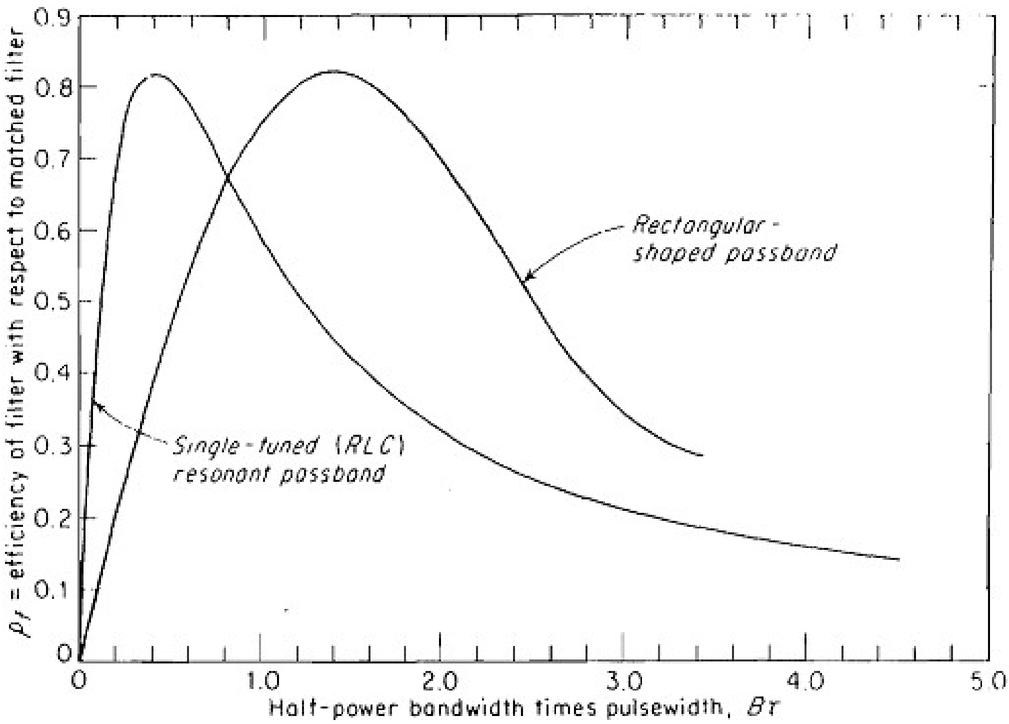


Fig: Efficiency, relative to a matched filter, of a single-tuned resonant filter and a rectangular shaped filter, when the input signal is a rectangular pulse of width τ. B = filter bandwidth.

Input signal	Filter	Optimum $B\tau$	Loss in SNR compared with matched filter, dB
Rectangular pulse	Rectangular	1.37	0.85
Rectangular pulse	Gaussian	0.72	0.49
Gaussian pulse	Rectangular	0.72	0.49
Gaussian pulse	Gaussian	0.44	0 (matched)
Rectangular pulse	One-stage, single-tuned circuit	0.4	0.88
Rectangular pulse	2 cascaded single-tuned stages	0.613	0.56
Rectangular pulse	5 cascaded single-tuned stages	0.672	0.5

Table: Efficiency of nonmatched filters compared with the matched filter

Matched filter with nonwhite noise:

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

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

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

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

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

Correlation Detection:

$$y_0(t) = \int_{-\infty}^{\infty} y_{\text{in}}(\lambda) s(t_1 - t + \lambda)\, d\lambda = R(t - t_1)$$

Equation above describes the output of the matched filter as the cross correlation between the input signal and a delayed replica of the transmitted signal. This implies that the matched-filter receiver can be replaced by a cross-correlation receiver that performs the same mathematical operation as shown in Fig.5. The input signal y (t) is multiplied by a delayed replica of the transmitted signal s(t - Tr), and the product is passed through a low-pass filter to perform the integration. The cross-correlation receiver of Fig.5 tests for the presence of a target at only a single time delay Tr. Targets at other time delays, or ranges, might be found by varying T_r. However, this requires a longer search time. The search time can be reduced by adding parallel channels, each containing a delay line corresponding to a particular value of Tr, as well as a multiplier and low-pass filter. In some applications it may be possible to record the signal on some storage medium, and at a higher playback speed perform the search sequentially with different values of T_r. That is, the playback speed is increased in proportion to the number of time-delay intervals T_r that are to be tested.

Since the cross-correlation receiver and the matched-filter receiver are equivalent

mathematically, the choice as to which one to use in a particular radar application is determined by which is more practical to implement. The matched-filter receiver, or an approximation, has been generally preferred in the vast majority of applications.

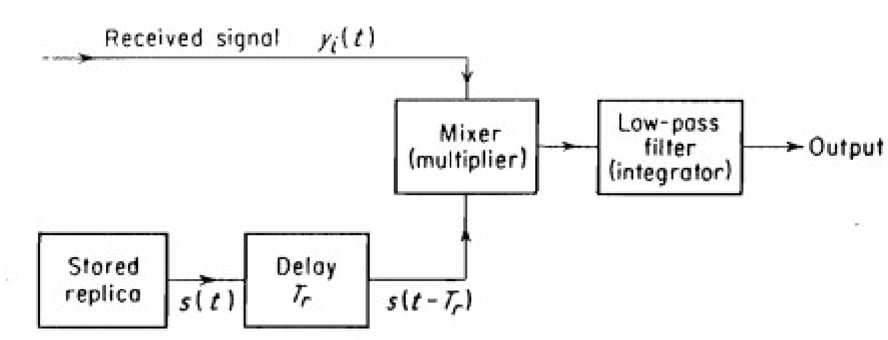


Fig: Block diagram of a cross-correlation receiver.

OBJECTIVE TYPE QUESTIONS

1. The maximum unambiguous range [ ]

a. Runamb = cτ/2 b. Runamb = (cτ) c. Runamb = cτ/4 d. Runamb = (cτ/2)

2. To operate MTI radar at low frequencies [ ]

a. λ fp must be small b. λ fp must be zero c. λ fp must be large d. λ fp must be unity

3. The effect of blind speed can be significantly reduced in [ ]

a. Pulse MTI radar b. Delay line canceller c. Staggered - prf MTI d. Pulse canceller

4. The blind speeds are present in pulse radar because [ ]

a. Doppler is measured by discrete samples at the prf

b. Doppler is measured by continuous signal c. Doppler is assumed to be zero d. Doppler frequency remains constant

5. If the first blind speed is to be greater than the maximum radial velocity expected from the

target, The product λ fp must be [ ]

a. Small b. Zero c. Large d. Infinity

6. The clutter-rejection notches may be widened by passing the output of the delay line canceller through a [ ]

a. Coho b. Stalo c. Second delay line canceller d.Pulse canceller

7. The frequency response of double delay line canceller is [ ]

a. 4 Sin πfd1T b. 4 π Sin πfd1T c. 4 Sin2 πfd1T d. $(2 \pi \text{ Sin } \pi fd1T)^2$

8. MTI radar primarily designed for the detection of aircraft must usually operate with [ ]

a. Unambiguous Doppler b. Unambiguous blind speed

c. Ambiguous Doppler d. Ambiguous range

9. The blind speeds of two independent radars operating at the same frequency will be different if Their [ ]

a. Amplitudes are different b. Blind speeds are different

c. Pulse repetition frequencies are different d. Pulse intervals are different

10. A disadvantage of the staggered prf is its inability to [ ]

a. Cancel second-time around echoes b. Cancel second-time around clutter

echoes c. Provide variable prf d. Provide pulse to pulse incoherence

11. Second-time around clutter echoes can be removed by use of a [ ]

a. Stalo b. Coho c. Constant prf d. Delay canceller

12. The loss of range information and the collapsing loss may be eliminated by [ ]

a. Sampling the range b. Shaping the range

c. Quantizing the range d. Keeping constant range

13. Range gating is a process of [ ]

a. Sampling the range into various samples b.Quantizing the range into small intervals c. Getting constant range d. Removing the range intervals

14. When the switching is pulse to pulse. It is known as a [ ]

a. Delay canceller b. Staggered prf c. MTI radar d. Pulse radar

15. Pulse to pulse coherence is provided by use of [ ]

a. Stalo b. Coho c. Constant prf d. Delay canceller

16. The output of the range gates is stretched in a circuit called [ ]

a. Clutter rejection filter b. Clutters filter c. Box car generator d. Sampler

17. The clutter rejection filter is a [ ]

a. Band stop filter   b. Band pass filter   c. Low pass filter   d. High pass filter

18. The bandwidth of clutter rejection filter depends upon the extent of the   [   ]

a. Spectrum   b. Expected clutter spectrum

c. Filter characteristic   d. Clutter characteristic

19. By quantizing the range [05D01]   [   ]

a. Loss of range Information is eliminated   b. Loss of range information is increased

c. Range becomes constant   d. Range becomes sampled

20. Clutter visibility factor provides   [   ]

a. Attenuation   b. Probabilies of detection and false alarm

c. Cancellation ratio   d. Decrease in clutter residue

Answers:

1.a	2.c	3.c	4.a	5.c	6.c	7.c	8.c	9.c	10.b
11.c	12.c	13.b	14.b	15.c	16.c	17.b	18.b	19.a	20b

ESSAY TYPE QUESTIONS

1. Explain the principle behind the operation of duplexers and receiver protectors.

2. Explain how a circulator can be utilized for a radar receiver protection.

3. Define noise figure and noise temperature of a receiver system.

4. Derive the impulse response of a matched filter that is commonly used in a a radar system.

5. Describe the principle behind the operation of a phased array antenna in a radar system.

6. Substantiate the requirement of duplexers in e cient radar systems. Describe the operation of branch and balanced type duplexers with necessary diagrams.

7. Describe any two types of duplexers used in radar receivers.

8. Define noise figure and equivalent noise temperature of a radar receiver.

9. A radar receiver is connected to a 50 ohm resistance antenna that has an equivalent noise resistance of 30 ohms. Calculate the noise figure of the receiver and the equivalent noise temperature of the receiver.

10. Discuss in detail about Matched-filter Receiver with necessary expressions.

11. Explain the function of time domain filter with an example.

12. An MTI radar operates at 10GHz with a PRF of 300 pps. Calculate the lowest blind speed?

13. Derive the impulse response of a matched filter that is commonly used in a radar receiver.

14. Explain how a threshold level is selected in threshold detection?

15. How to find the number of pulses that returned from a point target as the radar antenna scans through its beam width?

16. Why most of the radar receivers are considered as envelop detectors while calculating the SNR?

17. Discuss the relation between the matched filter characteristics and correlation detection

18. What is the difference between matched filter and non-matched filter?

19. Discuss the efficiency of non matched filters.

RADAR RECIEVERS

The function of the radar receiver is to detect desired echo signals in the presence of noise, interference, or clutter. It must separate wanted from unwanted signals, and amplify the wanted signals to a level where target information can be displayed to an operator or used in an automatic data processor. The design of the radar receiver will depend not only on the type of waveform to be detected, but on the nature of the noise, interference, and clutter echoes with which the desired echo signals must compete.

Noise can enter the receiver via the antenna terminals along with the desired signals, or it might be generated within the receiver itself. At the microwave frequencies usually used for radar, the external noise which enters via the antenna is generally quite low so that the receiver sensitivity is usually set by the internal noise generated within the receiver. The measure of receiver internal noise is the noise-figure.

Good receiver design is based on maximizing the output signal-to-noise ratio. To maximize the output signal-to-noise ratio, the receiver must be designed as a matched filter, or its equivalent. The matched filter specifies the frequency response function of the IF part of the radar receiver. Obviously, the receiver should be designed to generate as little internal noise as possible, especially in the input stages where the desired signals are weakest. Although special attention must be paid to minimize the noise of the input stages, the lowest noise receivers are not always desired in many radar applications if other important receiver properties must be sacrificed.

Receiver design also must be concerned with achieving sufficient gain, phase, and amplitude stability, dynamic range, tuning, ruggedness, and simplicity. Protection must be provided against overload or saturation, and burnout from nearby interfering transmitters. Timing and reference signals are needed to properly extract target information. Specific applications such as MTI radar, tracking radar, or radars designed to minimize clutter place special demands on the receiver. Receivers that must operate with a transmitter whose frequency can drift need some means of automatic frequency control (AFC). Radars that encounter hostile counter-measures need receivers that can minimize the effects of such interference. Thus there can be many demands placed upon the receiver designer in meeting the requirements of modern high-quality radar systems. The receiver engineer has responded well to the challenge, and there exists a highly refined state of technology available for radar applications. Radar receiver design and implementation may not always be an easy task; but in tribute to the receiver designer, it has seldom been an obstacle preventing the radar systems engineer from eventually accomplishing the desired objectives.

Although the super regenerative, crystal video and tuned radio frequency (TRF) receivers have been employed in radar systems, the super heterodyne has seen almost exclusive application because of its good sensitivity, high gain, selectivity, and reliability. No other receiver type has been competitive to the super heterodyne.

Noise Figure:

The noise figure of a receiver was described as a measure of the noise produced by a practical receiver as compared with the noise of an ideal receiver. The noise figure F_n of a linear network may be defined as

$$F_n = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{N_{out}}{kT_0 B_n G}$$

where S_{in} = available input signal power
N_{in} = available input noise power (equal to $kT_0 B_n$)
S_{out} = available output signal power

"Available power" refers to the power which would be delivered to a matched load. The available gain G is equal to S_{out}/S_{in}, k = Boltzmann's constant = 1.38 x 10^{-23} J/deg.

$$F_n = \frac{kT_0 B_n G + \Delta N}{kT_0 B_n G} = 1 + \frac{\Delta N}{kT_0 B_n G}$$

where N is the additional noise introduced by the network itself. The noise figure is commonly expressed in decibels, that is, 10 log F_n . The term noise factor is also used at times instead of noise figure. The two terms are synonymous.

The definition of noise figure assumes the input and output of the network are matched. In some devices less noise is obtained under mismatched, rather than matched, conditions. Inspite of definitions, such networks would be operated so as to achieve the maximum output signal-to-noise ratio.

Noise temperature:

The noise introduced by a network may also be expressed as an effective noise temperature, T_e, defined as that (fictional) temperature at the input of the network which would account for the noise N at the output. Therefore N = k T_e B $_n$ G and

$$F_n = 1 + \frac{T_e}{T_0}$$

$$T_e = (F_n - 1)T_0$$

The system noise temperature T_s is defined as the effective noise temperature of the receiver system including the effects of antenna temperature T_a. (It is also sometimes called the system operating noise temperature) If the receiver effective noise temperature is T_e, then

$$T_s = T_a + T_e = T_0 F_s$$

where F_s is the system noise-figure including the effect of antenna temperature. The effective noise temperature of a receiver consisting of a number of networks in cascade is

$$T_e = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \cdots$$

Where T_i and G_i , are the effective noise temperature and gain of the ith network. The effective noise temperature and the noise figure both describe the same characteristic of a network. In general, the effective noise temperature has been preferred for describing low-noise devices, and the noise figure is preferred for conventional receivers.

Noise figure of networks in cascade:

Consider two networks in cascade, each with the same noise bandwidth B_n but with different noise figures and available gain (Fig.3). Let F_1, G_1 be the noise figure and available gain, respectively, of the first network, and F_2, G_2 be similar parameters for the second network. The problem is to find F_o, the overall noise-figure of the two circuits in cascade. From the definition of noise figure the output noise N_o of the two circuits in cascade is

$$N_o = F_o G_1 G_2 k T_0 B_n = \text{noise from network 1 at output of network 2} + \text{noise } \Delta N_2 \text{ introduced by network 2}$$

$$N_o = kT_0 B_n F_1 G_1 G_2 + \Delta N_2 = kT_0 B_n F_1 G_1 G_2 + (F_2 - 1) kT_0 B_n G_2$$

or

$$F_o = F_1 + \frac{F_2 - 1}{G_1}$$

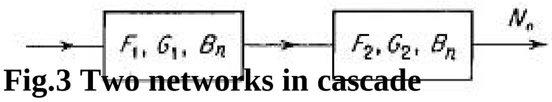


Fig.3 Two networks in cascade

The contribution of the second network to the overall noise-figure may be made negligible if the gain of the first network is large. This is of importance in the design of multistage receivers. It is not sufficient that only the first stage of a low-noise receiver have a small noise figure. The succeeding stage must also have a small noise figure, or else the gain of the first stage must be high enough to swamp the noise of the succeeding stage if the first network is not an amplifier but is a network with loss (as in a crystal mixer), the gain G_I should be interpreted as a number less than unity.
The noise figure of N networks in cascade may be shown to be

$$F_o = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots + \frac{F_N - 1}{G_1 G_2 \cdots G_{N-1}}$$

Balanced mixers: Noise that accompanies the local-oscillator (LO) signal can appear at the IF frequency because of the nonlinear action of the mixer. The LO noise must be removed if receiver sensitivity is to be maximized. One method for eliminating LO noise that interferes with the desired signal is to insert a narrow-bandpass RF filter between the local oscillator and the mixer. The center frequency of the filter is that of the local oscillator, and its bandwidth must be narrow so that LO noise at the signal and the image frequencies do not appear at the mixer. Since the receiver is tuned by changing the LO frequency, the narrowband filter must be tunable also.

A method of eliminating local-oscillator noise without the disadvantage of a narrow- bandwidth filter is the balanced mixer (Fig. 4.1). A balanced mixer uses a hybrid junction, a magic T, or an equivalent. These are four-port junctions. Figure 4.1 illustrates a magic T in which the LO and RF signals are applied to two ports. Diode mixers are in each of the remaining two arms of the magic T. At one of the diodes the sum of the RF and LO signals appears, and at the other diode the difference of the two is obtained. (In a magic T the LO would be applied to the H-plane arm, and the RF signal would be applied to the E-plane arm. The diode mixers would be mounted at equal distances in each of the collinear arms.) The two diode mixers should have identical characteristics and be well matched. The IF signal is

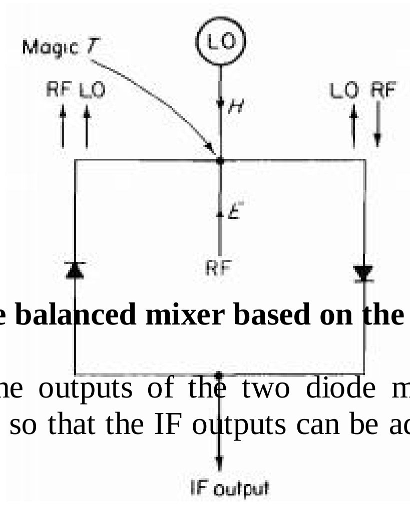


Fig.4.1 Principle of the balanced mixer based on the magic T

Recovered by subtracting the outputs of the two diode mixers. In Fig.4.1 the balanced diodes are shown reversed so that the IF outputs can be added. Local-oscillator noise at the two diode mixers will be in phase and will be canceled at the output. It is only the AM noise of the local oscillator which is canceled. The FM noise inserted by the local oscillator is unaffected by the balanced mixer.

In a single-ended mixer, the mixing action generates all harmonics of the RF and LO frequencies, and combinations thereof. The output is designed to filter out the frequency of interest, usually the difference frequency. A balanced mixer suppresses the even harmonics of the LO signal. A double balanced mixer is basically two single-ended mixers connected in parallel and 180˚ out of phase. It suppresses even harmonics of both the RF and the LO signals.

Reactive image termination: If the image frequency of a mixer is presented with the Proper reactive termination (such as an open or a short circuit), the conversion loss and the noise figure can be 1 to 2 dB less "broadband" mixer in which the image frequency is terminated in a matched reactive termination causes energy converted to the image frequency to be reflected back into the mixer and reconverted to IF Both the sum and the image frequencies can be reflected back to the diode mixer to achieve a lower conversion loss, but a number of adjustments are required for good results. One method for terminating the image in a reactive load is to employ a narrow bandpass filter, or preselector, at the RF signal frequency. A limitation of this approach is that it is not suitable for very wide bandwidths. The filter has to be retuned if the mixer must operate at another frequency. Also, the high Q of the filter introduces a loss which will increase the system noise-figure.

A method for achieving a reactive termination without narrow-bandwidth components is the image-recovery mixer shown in Fig.4.2. This has also been called an image-enhanced mixer, or product return mixer. (It is similar to the image-reject mixer whose purpose is to reject the image response.) The RF hybrid junction on the left of the circuit produces a 90˚ phase difference between the LO inputs to the two mixers. The IF hybrid junction on the right imparts another 90˚ phase differential in such a manner that the images cancel, but the IF signals from the two mixers add in phase. The two mixers in Fig.4.2 may be single-ended, balanced, or double-balanced mixers. This mixer is capable of wide bandwidth, and is restricted only by the frequency sensitivity of the structure of the microwave circuit. The noise figure of an image-recovery circuit is competitive with other receiver frontends. The image-recovery mixer is attractive as a receiver front-end because of its high dynamic range, low intermediation products, less susceptibility to burnout, and less cost as compared to other front-ends.

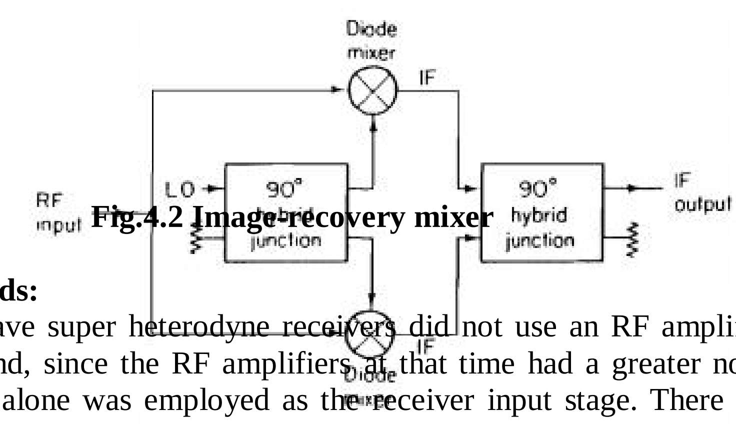


Fig.4.2 Image-recovery mixer

Low-Noise Front-Ends:

Early microwave super heterodyne receivers did not use an RF amplifier as the first stage, or front-end, since the RF amplifiers at that time had a greater noise figure than when the mixer alone was employed as the receiver input stage. There are now a number of RF amplifiers that can provide a suitable noise figure. Figure 5 plots noise figure as a function of frequency for the several receiver front-ends used in radar applications. The parametric amplifier has the lowest noise figure of those devices described here, especially at the higher microwave frequencies. However, it is generally more complex and expensive compared to the other front-ends.

The transistor amplifier can be applied over most of the entire range of frequencies of interest to radar. The silicon bipolar-transistor has been used at the lower radar frequencies (below L band) and the gallium arsenide field-effect transistor (Ga As FET) is preferred at the higher frequencies. The transistor is generally used in a multistage configuration with a typical gain per stage decreasing from 12 dB at VHF to 6 dB at K_u band. In the Ga As FET, the thermal noise contribution is greater than the shot noise. Cooling the device will therefore improve the noise figure.

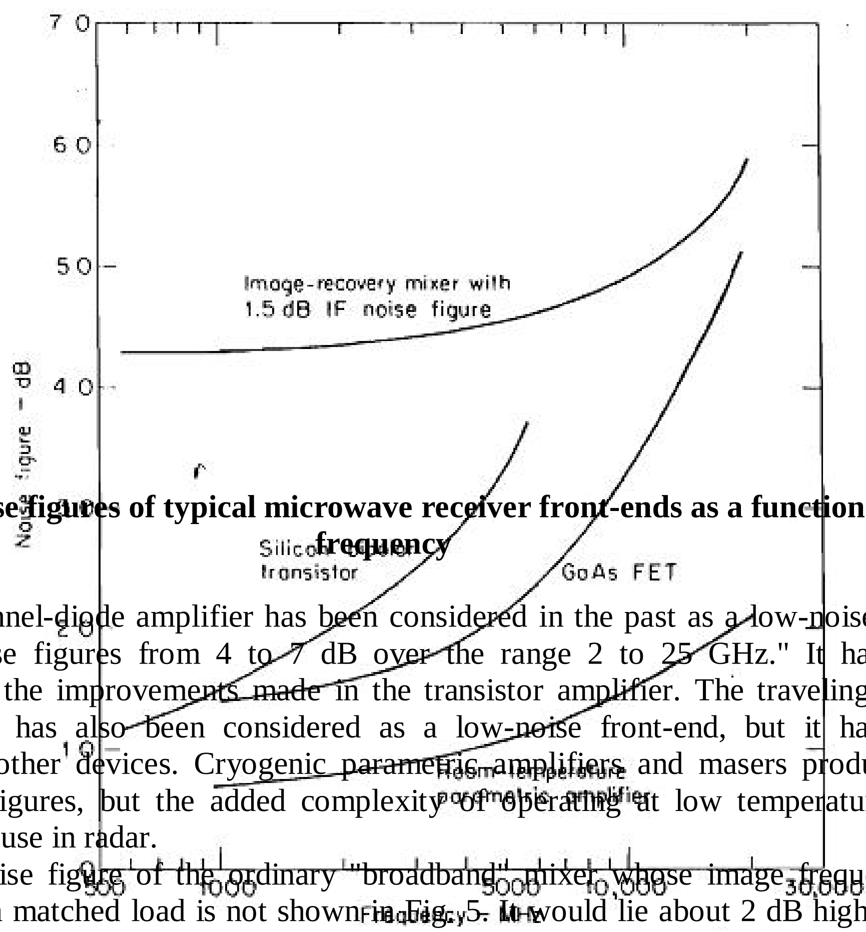


Fig.5 Noise figures of typical microwave receiver front-ends as a function of frequency

The tunnel-diode amplifier has been considered in the past as a low-noise front-end, with noise figures from 4 to 7 dB over the range 2 to 25 GHz." It has been supplanted by the improvements made in the transistor amplifier. The traveling-wave-tube amplifier has also been considered as a low-noise front-end, but it has been overtaken by other devices. Cryogenic parametric amplifiers and masers produce the lowest noise figures, but the added complexity of operating at low temperatures has tempered their use in radar.

The noise figure of the ordinary "broadband" mixer whose image frequency is terminated in a matched load is not shown in Fig. 5. It would lie about 2 dB higher than the noise figure shown for the image-recovery mixer.

There are other factors beside the noise figure which can influence the selection of a receiver front-end. Cost, burnout, and dynamic range must also be considered. The selection of a particular type of receiver front-end might also be influenced by its instantaneous band-width, tuning range, phase and amplitude stability, and any special requirements for cooling. The image-recovery mixer represents a practical compromise which tends to balance its slightly greater noise figure by its lower cost, greater ruggedness, and greater dynamic range.

Utility of low-noise front-ends:

The lower the noise figure of the radar receiver, the less need be the transmitter power and/or the antenna aperture. Reductions in the size of the transmitter and the antenna are always desirable if there are no concomitant reductions in performance. A few decibels improvement in receiver noise-figure can be obtained at a relatively low cost as compared to the cost and complexity of adding the same few decibels to a high-power transmitter.

There are, however, limitations to the use of a low-noise front-end in some radar applications. As mentioned above, the cost, burnout, and dynamic range of low-noise devices might not be acceptable in some applications. Even if the low-noise device itself is of large dynamic range, there can be a reduction of the dynamic range of the receiver as compared to a receiver with a mixer as its front-end. Dynamic range is usually defined as the ratio of the maximum signal that can be handled to the smallest signal capable of detected. if the mixer rather than the IF amplifier is what limits the total dynamic range of the receiver, the introduction of the low-noise front-end will cause a sacrifice in the mixer dynamic range and, consequently, the dynamic range of the entire receiver.

In military radar, a low-noise receiver car1 make the radar more susceptible to the effects of deliberate electronic countermeasures (ECM). When practical, it may be preferred to deliberately employ a conventional receiver with modest sensitivity and to make up for the reduced sensitivity by larger transmitter power. This is not the most economical way to build radar, but it does make the task of the hostile ECM designer more difficult.

A variety of low-noise radar receivers are available to the radar system designer. The well-recognized benefits of low-noise receivers, combined with their relative affordability, make them an attractive feature in modern radar design. However, low noise receivers are sometimes accompanied by other less desirable properties that tend to result in a compromise in receiver performance. Thus a low-noise receiver might not always be the obvious selection, if properties other than sensitivity are important.

Displays:

The purpose of the display is to visually present in a form suitable for operator interpretation and action the information contained in the radar echo signal. When the display is connected directly to the video output of the receiver, the information displayed is called raw video. This is the "traditional" type of radar presentation. When the receiver video output is first processed by an automatic detector or automatic detection and tracking processor (ADT), the output displayed is sometimes called synthetic video.

The cathode-ray tube (CRT) has been almost universally used as the radar display. There are two basic cathode-ray tube displays. One is the dejection-modulated CRT, such as the A-scope, in which a target is indicated by the deflection of the electron beam.

On the other hand, intensity-modulated displays have the advantage of presenting data in a convenient and easily interpreted form. The deflection of the beam or the appearance of an intensity-modulated spot on a radar display caused by the presence of a target is commonly referred to as a blip. .

Types of display presentations: The various types of CRT displays which might be used for surveillance and tracking radars are defined as follows:

- **A-scope:** A deflection-modulated display in which the vertical deflection is proportional to target echo strength and the horizontal coordinate is proportional to range.
- **B-scope:** An intensity-modulated rectangular display with azimuth angle indicated by the horizontal coordinate and range by the vertical coordinate.
- **C-scope:** An intensity-modulated rectangular display with azimuth angle indicated by the horizontal coordinate and elevation angle by the vertical coordinate.
- **D-scope:** A C-scope in which the blips extend vertically to give a rough estimate of distance.
- **E-scope:** An intensity-modulated rectangular display with distance indicated by the horizontal coordinate and elevation angle by the vertical coordinate. Similar to the RHI in which target height or altitude is the vertical coordinate.
- **F-Scope:** A rectangular display in which a target appears as a centralized blip when the radar antenna is aimed at it. Horizontal and vertical aiming errors are respectively indicated by the horizontal and vertical displacement of the blip.
- **G-Scope:** A rectangular display in which a target appears as a laterally centralized blip when the radar antenna is aimed at it in azimuth, and wings appear to grow on the pip as the distance to the target is diminished; horizontal and vertical aiming errors are respectively indicated by horizontal and vertical displacement of the blip.
- **H-scope:** A B-scope modified to include indication of angle of elevation. The target appears as two closely spaced blips which approximate a short bright line, the slope of which is in proportion to the sine of the angle of target elevation.
- **I-scope:** A display in which a target appears as a complete circle when the radar antenna is pointed at it and in which the radius of the circle is proportional to target distance; incorrect aiming of the antenna changes the circle to a segment whose arc length is inversely proportional to the magnitude of the pointing error, and the position of the segment indicates the reciprocal of the pointing direction of the antenna.
- **J-scope:** A modified A-scope in which the time base is a circle and targets appear as radial deflections from the time base.

- **K-scope:** A modified A-scope in which a target appears as a pair of vertical deflections. When the radar antenna is correctly pointed at the target, the two deflections are of equal height, and when not so pointed, the difference in deflection amplitude is an indication of the direction and magnitude of the pointing error.

- **L-scope:** A display in which a target appears as two horizontal blips, one extending to the right from a central vertical time base and the other to the left; both blips are of equal amplitude when the radar is pointed directly at the target, any inequality representing relative pointing error, and distance upward along the baseline representing target distance.

- **M-scope:** A type of A-scope in which the target distance is determined by moving an adjustable pedestal signal along the baseline until it coincides with the horizontal position of the target signal deflections; the control which moves the pedestal is calibrated in distance.

- **N-scope:** A K-scope having an adjustable pedestal signal, as in the M-scope, for the measurement of distance.

- **O-scope:** An A-scope modified by the inclusion of an adjustable notch for measuring distance.

- **PPI or Plan Position Indicator (also called P-scope):** An intensity-modulated circular display on which echo signals produced from reflecting objects are shown in plan position with range and azimuth angle displayed in polar (rho-theta) coordinates, forming a map-like display. An offset, or off center, PPI has the zero position of the time base at a position other than at the center of the display to provide the equivalent of a larger display for a selected portion of the service area. A delayed PPI is one in which the initiation of the time base is delayed.

- **R-scope:** An A-scope with a segment of the time base expanded near the blip for greater accuracy in distance measurement.

- **RHI or Range-Height Indicator:** An intensity modulated display with height (altitude) as the vertical axis and range as the horizontal axis.

(i) Branch-type duplexers:

The branch-type duplexer, diagrammed in Fig.7.1 was one of the earliest duplexer configurations employed. It consists of a TR (transmit-receive) switch and an ATR (anti-transmit receive) switch, both of which are gas-discharge tubes. When the transmitter is turned on, the TR and the ATR tubes ionize; that is, they break down, or fire. The TR in the fired condition acts as a short circuit to prevent transmitter power from entering the receiver. Since the TR is located a quarter wavelengths from the main transmission line, it appears as a short circuit at the receiver but as an open circuit at the transmission line so that it does not impede the flow of transmitter power. Since the ATR is displaced a quarter wavelengths from the main transmission line, the short circuit it produces during the fired condition appears as an open circuit on the transmission line and thus has no effect on transmission.

During reception, the transmitter is off and neither the TR nor the ATR is fired. The open circuit of the ATR, being a quarter wave from the transmission line, appears as a short circuit across the line. Since this short circuit is located a quarter wave from the receiver branch-line, the transmitter is effectively disconnected from the line and the echo signal power is directed to the receiver. The diagram of Fig.7.1 is a parallel configuration. Series or series-parallel configurations are possible.

The branch-type duplexer is of limited bandwidth and power-handling capability, and has generally been replaced by the balanced duplexer and other protection devices. It is used, inspite of these limitations, in some low-cost radars.

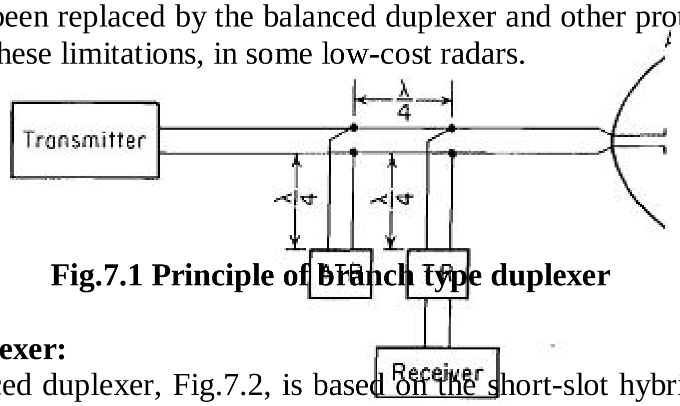


Fig.7.1 Principle of branch type duplexer

(ii) Balanced duplexer:

The balanced duplexer, Fig.7.2, is based on the short-slot hybrid junction which consists of two sections of waveguides joined along one of their narrow walls with a slot cut in the common narrow wall to provide coupling between the two. The short-slot hybrid may be considered as a broadband directional coupler with a coupling ratio of 3 dB. In the transmit condition (Fig. 7.2 a) power is divided equally into each waveguide by the first short slot hybrid junction. Both TR tubes break down and reflect the incident power out the antenna arm as shown

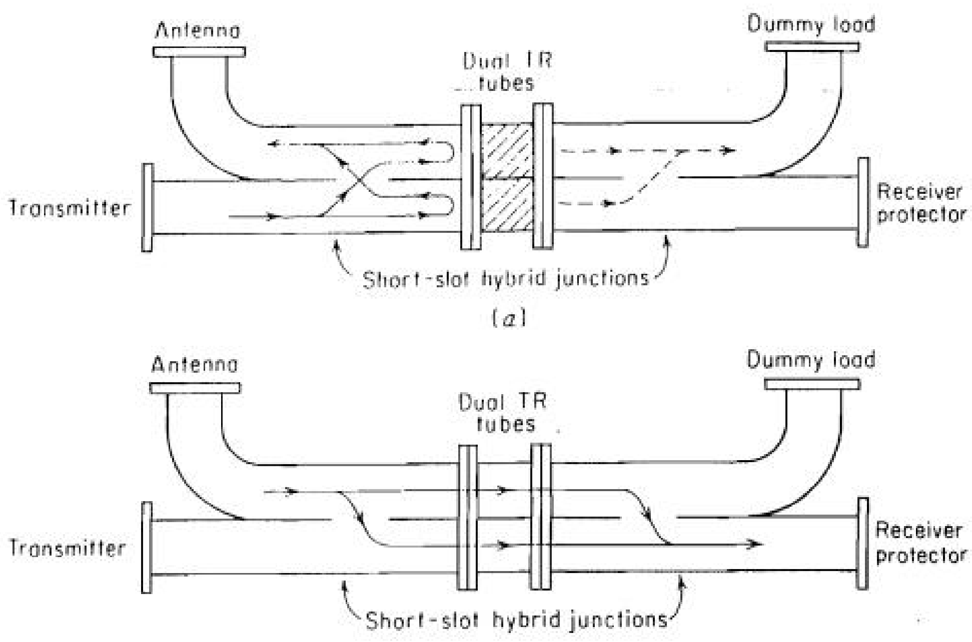


Fig.7.2. balanced duplexer using dual TR tubes and two short-slot hybrid junctions (a) Transmit condition; (b) receive condition.

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

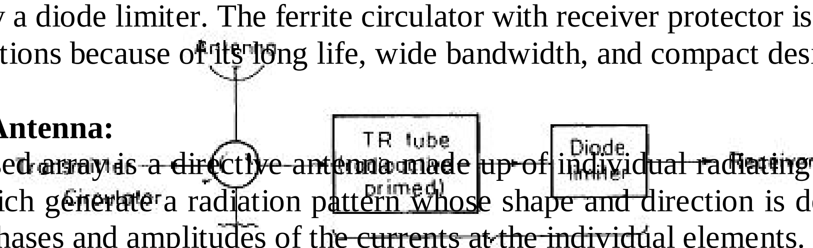
On reception the TR tubes are unfired and the echo signals pass through the duplexer and into the receiver as shown in Fig. 7.2 b. The power splits equally at the first junction and because of the 90˚ phase advance on passing through the slot, the energy recombines in the receiving arm and not in the dummy-load arm.

The power-handling capability of the balanced duplexer is inherently greater than that of the branch-type duplexer and it has wide bandwidth, over ten percent with proper design. A receiver protector is usually inserted between the duplexer and the receiver for added protection.

Circulator and receiver protector:

The ferrite circulator is a three-or four-port device that can, in principle, offer separation of the transmitter and receiver without the need for the conventional duplexer configurations. The circulator does not provide sufficient protection by itself and requires a receiver protector ..

The isolation between the transmitter and receiver ports of a circulator is seldom sufficient to protect the receiver from damage. However, it is not the isolation between transmitter and receiver ports that usually determines the amount of transmitter power at the receiver, but the impedance mismatch at the antenna which reflects transmitter power back into the receiver. The VSWR is a measure of the amount of power reflected by the antenna. For example, a VSWR of 1.5 means that about 4 percent of the transmitter power will be reflected by the antenna mismatch in the direction of the receiver, which corresponds to an isolation of only 14 dB. About 11 percent of the power is reflected when the VSWR is 2.0, corresponding to less than 10 dB of isolation. Thus, a receiver protector is almost always required. It also reduces to safe level radiations from nearby transmitters. The receiver protector might use solid-state diodes for an all solid-state configuration, or it might be a passive TR-limiter consisting of a radioactive primed TR-tube followed by a diode limiter. The ferrite circulator with receiver protector is attractive for radar applications because of its long life, wide bandwidth, and compact design.



Phased Array Antenna:

The phased array is a directive antenna made up of individual radiating antennas, or elements, which generate a radiation pattern whose shape and direction is determined by the relative phases and amplitudes of the currents at the individual elements.

By properly varying the relative phases it is possible to steer the direction of the radiation. The radiating elements might be dipoles open-ended waveguides, slots cut in waveguide, or any other type of antenna. The inherent flexibility offered by the phased-array antenna in steering the beam by means of electronic control is what has made it of interest for radar. It has been considered in those radar applications where it is necessary to shift the beam rapidly from one position in space to another, or where it is required to obtain information about many targets at a flexible, rapid data rate. The full potential of a phased-array antenna requires the use of a computer that can determine in real time, on the basis of the actual operational situation, how best to use the capabilities offered by the array.

Radiation pattern for Phased array Antenna: Consider a linear array made up of N elements equally spaced a distance d apart (Fig.9). The elements are assumed to be isotropic point sources radiating uniformly in all directions with equal amplitude and phase. Although isotropic elements are not realizable in practice, they are a useful concept in array theory, especially for the computation of radiation patterns. The array is shown as a receiving antenna for convenience, but because of the reciprocity principle, the results obtained apply equally well to a transmitting antenna. The outputs of all the elements are summed via lines of equal length to give a sum output voltage E_a. Element 1 will be taken as the reference signal with zero phases.

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

The sum of all the voltages from the individual elements, when the phase difference between adjacent elements is Ψ, can be written

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

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

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

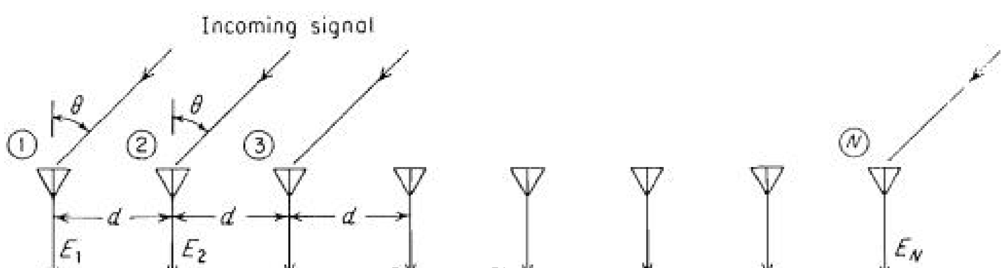


Fig.9 N-element linear array

The first factor is a sine wave of frequency ω with a phase shift $(N - 1)\Psi/2$ (if the phase reference were taken at the center of the array, the phase shift would be zero), while the second term represents an amplitude factor of the form $\sin(N\Psi/2)/\sin(\Psi/2)$. The field intensity pattern is the magnitude of the Eq. above or

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

The pattern has nulls (zeros) when the numerator is zero. The latter occurs when

$N\Pi\ (d/\lambda) \sin\theta = 0, \pm \Pi, \pm 2\Pi..., \pm n\Pi$, where n = integer. The denominator, however, is zero when $\Pi\ (d/\lambda) \sin\theta, = 0, \pm \Pi, \pm 2\Pi..., \pm n\Pi$. Note that when the denominator is zero, the numerator is also zero. The value of the field intensity pattern is indeterminate when both the denominator and numerator are zero. However, by applying L'Hopital's rule (differentiating numerator and denominator separately) it is found that

$|E_a|$ is a maximum whenever $\sin\theta = \pm n\lambda/d$. These maxima all have the same value and are equal to N. The maximum at $\sin\theta = 0$ defines the main beam. The other maxima are called grating lobes. They are generally undesirable and are to be avoided. If the spacing between elements is a half-wavelength ($d/\lambda = 0.5$), the first grating lobe ($n = \pm 1$) does not appear in real space since $\sin\theta > 1$, which cannot be.

Grating lobes appear at ± 90° when $d = \lambda$. For a non scanning array (which is what is considered here) this condition ($d = \lambda$) is usually satisfactory for the prevention of grating lobes. Equation discussed above applies to isotropic radiating elements, but practical antenna elements that are designed to maximize the radiation at $\theta = 0°$, generally have negligible radiation in the direction $\theta = \pm 90°$. Thus the effect of a realistic element pattern is to suppress the grating lobes at ±90°. It is for this reason that element spacing equal to one wavelength can be tolerated for a nonscanning array.

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

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

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

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

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

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

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

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

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

Beam steering phased array antennas:

The beam of an array antenna may be steered rapidly in space without moving large mechanical masses by properly varying the phase of the signals applied to each element.

Consider an array of equally spaced elements. The spacing between adjacent elements is d, and the signals at each element are assumed of equal amplitude. If the same phase is applied to all elements, the relative phase difference between adjacent elements is zero and the position of the main beam will be broadside to the array at an angle $\theta = 0$. The main beam will point in a direction other than broadside if the relative phase difference between elements is other than zero. The direction of the main beam is at an angle θ_0**,** when the phase difference is $Ø = 2\pi(d/\lambda) \sin \theta_0$**.** The phase at each element is therefore $Ø_c = mØ$, where m = 0, 1, 2, . . . , (N - 1), and $Ø_c$ **,** is any constant phase applied to all elements. The normalized radiation pattern of the array when the phase difference between adjacent elements is Ø is given by The maximum of the radiation pattern occurs when $\sin \theta = \sin \theta_o$

The above equation states that the main beam of the antenna pattern may be positioned to an angle θ_0 by the insertion of the proper phase shift *Ø* at each element of the array. If variable rather than fixed, phase shifters are used, the beam may be steered as the relative phase between elements is changed.

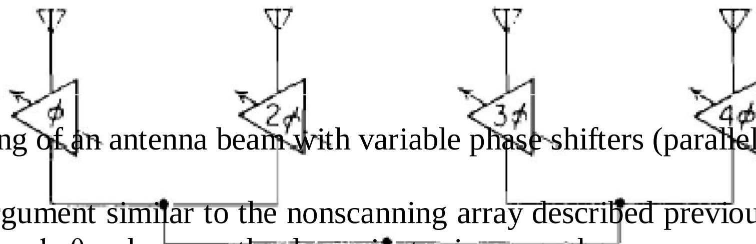


Figure 10.1 Steering of an antenna beam with variable phase shifters (parallel-fed array).

Using an argument similar to the nonscanning array described previously, grating lobes appear at an angle θ_o whenever the denominator is zero, when

$$\pi \frac{d}{\lambda}(\sin \theta_g - \sin \theta_0) = \pm n\pi$$

$$|\sin \theta_g - \sin \theta_0| = \pm \frac{\lambda}{d}$$

If a grating lobe is permitted to appear at -90° when the main beam is steered to +90°, it is found from the above that $d = \lambda / 2$. Thus the element spacing must not be larger than a half wavelength the beam is to be steered over a wide angle without having undesirable grating lobes appear. Practical array antennas do not scan ±90°. If the scan is limited to ±60°, the element spacing should be less than 0.54λ.

Beam width of a phased array antenna will vary with steering angle.

- **Change of beam width with steering angle:** The half-power beam width in the plane of scan increases as the beam is scanned off the broadside direction. The beam width is approximately inversely proportional to $\cos\theta_o$, where θ_o is the angle measured from the normal to the antenna. This may be proved by assuming that the sine in the denominator of Eq. discussed earlier can be replaced by its argument, so that the radiation pattern is of the form $(\sin^2 u)/u^2$,
- Where $u = N\Pi (d/\lambda)(\sin\theta - \sin\theta_o)$.
- 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 $\theta > \theta_o$, and , the angle corresponding to the half-power point when $\theta < \theta_o$; that is, θ_+ corresponds $+0.443\Pi$ and θ_- to $u = -0.443\Pi$. The $\sin\theta - \sin\theta_o$, term in the expression for u can be written

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

- The second term on the right-hand side of Eq. above can be neglected when θ_o is small (beam is near broadside), so that

$$\sin \theta - \sin \theta_0 \approx \sin (\theta - \theta_0) \cos \theta_0$$

- Using the above approximation, the two angles corresponding to the 3-dB points of the antenna pattern are

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

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

o The half-power beamwidth is

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

- Therefore, when the beam is positioned an angle θ_o off broadside, the beamwidth in the plane of scan increases as $(\cos \theta_o)^{-1}$. The change in beamwidth with angle θ_o, as derived above is not valid when the antenna beam is too far removed from broadside. It certainly does not apply when the energy is radiated in the end fire direction.

- Equation above applies for a uniform aperture illumination. With a cosine-on-a-pedestal aperture illumination of the form $A_n = a_o + 2a_1 \cos 2\Pi n/N$, the beamwidth is

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

- The parameter n in the aperture illumination represents the position of the element. Since the illumination is assumed symmetrical about the center element, the parameter n takes on values of n = 0, ±1, ± 2 . . . ± (N - 1)/2. The range of interest is $0 \leq 2a_1 \leq a_o$ which covers the span from uniform illuminations to a taper so severe that the illumination drops to zero at the ends of the array. (The array is assumed to extend a distance d/2 beyond each end element.)

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

o **Mechanically scanned array*:***

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

- **Linear array with frequency scan:**
  - The frequency-scanned, linear array feeding a parabolic cylinder or a planar array of slotted waveguides has seen wide application as a 3D air-surveillance radar. In this application, a pencil beam is scanned in elevation by use of frequency and scanned in azimuth by mechanical rotation of the entire antenna.
- **Linear array with phase scan:**
  - Electronic phase steering, instead of frequency scanning, in the 3D air-surveillance radar is generally more expensive, but allows the use of the frequency domain for purposes other than beam steering.
  - The linear array configuration is also used to generate multiple, contiguous fixed beams (stacked beams) for 3D radar. Another application is to use either phase- or frequency-steering in a stationary linear array to steer the beam in one angular coordinate, as for the GCA radar.
- **Phase-frequency planar array:**
  - A two-dimensional (planar) phased array can utilize frequency scanning to steer the beam in one angular coordinate and phase shifters to steer in the orthogonal coordinate. This approach is generally easier than using phase shifters to scan in both coordinates, but as with any frequency-scanned array the use of the frequency domain for other purposes is limited when frequency is employed for beam-steering.
- **Phase-phase planar array:**
  - The planar array which utilizes phase shifting to steer the beam in two orthogonal coordinates is the type of array that is of major interest for radar application because of its inherent versatility. Its application, however, has been limited by its relatively high cost. The phase-phase array is what is generally implied when the term electronically steered phased array is used.
  - The array antenna has the following desirable characteristics not generally enjoyed by other antenna types:

- **Inertia less rapid beam-steering**:
  - The beam from an array can be scanned, or switched from one position to another, in a time limited only by the switching speed of the phase shifters. Typically, the beam can be switched in several microseconds, but it can be considerably shorter if desired.
- **Multiple, independent beams**:
  - A single aperture can generate many simultaneous independent beams. Alternatively, the same effect can be obtained by rapidly switching a single beam through a sequence of positions.
- **Potential for large peak and / or average power**:
  - If necessary, each element of the array can be fed by a separate high-power transmitter with the combining of the outputs made in space to obtain a total power greater than can be obtained from a singlransmitter.
- **Control of the radiation pattern:**
  - A particular radiation pattern may be more readily obtained with the array than with other microwave antennas since the amplitude and phase of each array element may be individually controlled. Thus, radiation patterns with extremely low sidelobes or with a shaped main beam may be achieved. Separate monopulse sum and difference patterns, each with its own optimum shape, are also possible.
- **Graceful degradation:**
  - The distributed nature of the array means that it can fail gradually rather than all at once (catastrophically).
- **Convenient aperture shape:**
  - The shape of the array permits flush mounting and it can be hardened to resist blast.
- **Electronic beam stabilization:**
  - The ability to steer the beam electronically can be used to stabilize the beam direction when the radar is on a platform, such as a ship or aircraft, that is subject to roll, pitch, or yaw.

- **Limitations:**

    - The major limitation that has limited the widespread use of the conventional phased array in radar is its high cost, which is due in large part to its complexity.
    - When graceful degradation has gone too far a seperate maintenance is needed.
    - When a planar array is electronically scanned, the change of mutual coupling that accompanies a change in beam position makes the maintenance of low sidelobes more difficult.
    - Although the array has the potential for radiating large power, it is seldom that an array is required to radiate more power than can be radiated by other antenna types or to utilize a total power which cannot possibly be generated by current high-power microwave tube technology that feeds a single transmission line.