MTI AND PULSE DOPPLER RADAR

PRINCIPLE OF MTI RADAR

BLOCK DIAGRAM OF MTI RADAR

POWER AMPLIFIER TRANSMITTER
Block diagram of MTI radar with power-amplifier transmitter
BUTTERFLY EFFECT THAT IS PRODUCED BY MTI
DELAY LINE CANCELLERS:
FILTER CHARACTERISTICS OF THE DELAY-LINE CANCELLER

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

Where, \( \phi_0 = \) phase shift

\[ k = \text{amplitude of video signal.} \]

The signal from the previous transmission, which is delayed by a time \( T = \) pulse repetition interval, is

\[ V_2 = k \sin (2\pi f_d (t - T) - \phi_0) \]

Everything else is assumed to remain essentially constant over the interval \( T \) so that \( k \) is the same for both pulses. The output from the subtractor is

\[ V = V_1 - V_2 = 2k \sin \pi f_d T \cos [2\pi f_d (t - T / 2) - \phi_0] \]
Frequency response of the single delay-line canceler; $T = \text{delay time} = \frac{1}{f_p}$
BLIND SPEEDS

The response of the single-delay-line canceller will be zero whenever the argument $\pi f_d T$ in the amplitude factor of $V = V_1 - V_2 = 2k \sin \pi f_d T \cos [2\pi f_d (t - T/2) - \phi_0]$ is 0, $\pi$, $2\pi$, .., etc., or when

$$f_d = n / T = n f_p$$

where, $n = 0, 1, 2, \ldots$

$$f_p = \text{pulserpetition frequency}.$$ 

$$v_n = \frac{n \lambda}{2T} = \frac{n \lambda f_p}{2} \quad n = 1, 2, 3, \ldots$$

where, $v_n$ is the nth blind speed.

If $\lambda$ is measured in meters, $f_p$ in Hz, and the relative velocity in knots, the blind speeds are
\[ v_n = \frac{n \lambda \nu_p}{1.02} \approx n \lambda \nu_p \]
DOUBLE CANCELLATION:

\[ f(t) - 2f(t + T) + f(t + 2T) \]

Fig (a) Double-delay-line canceller; (b) three-pulse canceller

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

\[ f(t) - f(t + T) - f(t + T) + f(t + 2T) \]
This configuration is commonly called the three-pulse canceller.

Relative frequency response of the single-delay-line canceller (solid curve) and the double-delay-line canceller (dashed curve).
STAGGERED PRFS

(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 $4/T_1 = 5/T_2$. 
RANGE GATED DOPPLER FILTERS

block diagram of MTI radar using range gates and filters
Frequency-response characteristic of an MTI using range gates and filters
block diagram of a non coherent MTI radar
LIMITATIONS OF MTI RADAR

Equipment instabilities

A cos \( w t \) - A cos \( (w t + \Delta \phi) \) = 2A sin \( (\Delta \phi / 2) \) sin \( (w t + \Delta \phi / 2) \).

For small phase errors, the amplitude of the resultant difference is

\[ 2A \sin \frac{\Delta \phi}{2} \approx A \Delta \phi. \]

Therefore the limitation on the improvement factor due to oscillator instability is

\[ I = \frac{1}{\Delta \phi^2} \]

Scanning modulation

As the antenna scans by a target, it observes the target for a finite time equal to

\[ t_0 = \frac{n_B}{f_p} = \frac{\theta_B}{\theta_S} \]

Where, \( n_B = \) number of hits received, \( f_p = \) pulse repetition frequency, \( \theta_B = \) antenna beamwidth and \( \theta_S = \) antenna scanning rate
Thanks
A tracking-radar system measures the coordinates of a target and provides data which may be used to determine the target path and predict its future position.
SEQUENTIAL LOBING

Lobe-switching antenna patterns and error signal (one dimension). (a) Polar representation of switched antenna patterns (b) rectangular representation (c) error signal.
CONICAL SCAN

Conical-scan tracking
Block diagram of conical-scan tracking radar
MONOPULSE TRACKING RADAR

AGC portion of tracking radar receiver

Block diagram of the AGC portion of a tracking-radar receiver
Monopulse antenna patterns and error signal
Block diagram of amplitude-comparison monopulse radar
amplitude-comparison monopulse radar for extracting error signals in both elevation and azimuth.

Block diagram of two-coordinate (azimuth and elevation) amplitude-comparison monopulse tracking radar
approximately ideal feed-aperture illumination for monopulse sum and difference channels
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

\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.
Wavefront phase relationships in phase-comparison monopulse radar
Split-range-gate tracking (a) Echo pulse; (b) early-late range gates; (c) difference signal between early and late range gates.
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.
LIMITATIONS TO TRACKING ACCURACY

Main limitations to tracking accuracy of radar are,

- Amplitude fluctuations.
- Angle fluctuations.
- Receiver and servo noise
COMPARISON OF TRACKERS

Low-angle tracking
\[ \Delta R = \frac{2h_a h_t}{R} \]

Where,

\( h_a \) = radar antenna height, \( h_t \) = target height,

\( R \) = range to the target
THANK YOU
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

RESPONSE CHARACTERISTICS AND DERIVATION

\[ S(f) = \int_{-\infty}^{\infty} s(t) \exp(-j2\pi ft) \, dt \]

where \( S(f) \) is the voltage spectrum (Fourier transform) of input signal

\[ S^*(f) = \text{complex conjugate of } S(f) \]

\( t_1 = \text{fixed value of time at which signal is observed to be maximum} \)

\( G_a = \text{constant equal to maximum filter gain (generally taken to be unity)} \)
\[ |H(f)| \exp \left\{ -j\phi_m(f) \right\} = |S(f)| \exp \left\{ j[\phi_s(f) - 2\pi ft_1] \right\} \]

or

\[ |H(f)| = |S(f)| \]

and

\[ \phi_m(f) = -\phi_s(f) + 2\pi ft_1 \]

\[ h(t) = \int_{-\infty}^{\infty} H(f) \exp (j2\pi ft) \, df \]

\[ h(t) = G_a \int_{-\infty}^{\infty} S^*(f) \exp \left\{ -j2\pi f(t_1 - t) \right\} \, df \]
Since $S^*(f) = S(-f)$, we have

$$h(t) = G_a \int_{-\infty}^{\infty} S(f) \exp \left[ j2\pi f (t_1 - t) \right] df = G_a s(t_1 - t)$$
Derivation of the matched-filter characteristic

(a) Received waveform \( s(t) \); (b) impulse response \( h(t) \) of the matched filter
\[ H(f) = G_p S^*(f) \exp(-j2\pi ft_1) \]

\[ R_f = \frac{|S_o(t)|^2_{\text{max}}}{\mathcal{N}} \]

\[ |s_o(t)| = \left| \int_{-\infty}^{\infty} S(f)H(f) \exp(j2\pi ft) \, df \right| \]
\[ N = \frac{N_p}{2} \int_{-\infty}^{\infty} |H(f)|^2 \, df \]

\[ R_f = \left( \left| \int_{-\infty}^{\infty} S(f)H(f) \exp(j2\pi f t_1) \, df \right|^2 \right) \left( \frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2 \, df \right)^{-1} \]

\[ \int P^*P \, dx - \int Q^*Q \, dx \geq \left| \int P^*Q \, dx \right|^2 \]
\[ P^* = S(f) \exp(j2\pi ft_1) \quad \text{and} \quad Q = H(f) \]

and recalling that

\[ \int P^*P \, dx = \int |P|^2 \, dx \]

\[ R_f \leq \frac{\int_{-\infty}^{\infty} |H(f)|^2 \, df \int_{-\infty}^{\infty} |S(f)|^2 \, df}{\frac{N_0}{2} \int_{-\infty}^{\infty} |H(f)|^2 \, df} = \frac{\int_{-\infty}^{\infty} |S(f)|^2 \, df}{\frac{N_0}{2}} \]

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

\[ R_f \leq \frac{2E}{N_0} \]

\[ H(f) = G_a S^*(f) \exp(-j2\pi ft_1) \]
CORRELATION FUNCTION

\[ R(t) = \int_{-\infty}^{\infty} y(\lambda) s(\lambda - t) \, d\lambda \]

\[ y_0(t) = \int_{-\infty}^{\infty} y_{in}(\lambda) h(t - \lambda) \, d\lambda \]

\[ y_0(t) = \int_{-\infty}^{\infty} y_{in}(\lambda) s(t_1 - t + \lambda) \, d\lambda = R(t - t_1) \]
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 \( \tau \). \( B = \text{filter bandwidth} \).
Table Efficiency of nonmatched filters compared with the matched filter

<table>
<thead>
<tr>
<th>Input signal</th>
<th>Filter</th>
<th>Optimum $B_T$</th>
<th>Loss in SNR compared with matched filter, dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rectangular pulse</td>
<td>Rectangular</td>
<td>1.37</td>
<td>0.85</td>
</tr>
<tr>
<td>Rectangular pulse</td>
<td>Gaussian</td>
<td>0.72</td>
<td>0.49</td>
</tr>
<tr>
<td>Gaussian pulse</td>
<td>Rectangular</td>
<td>0.72</td>
<td>0.49</td>
</tr>
<tr>
<td>Gaussian pulse</td>
<td>Gaussian</td>
<td>0.44</td>
<td>0 (matched)</td>
</tr>
<tr>
<td>Rectangular pulse</td>
<td>One-stage, single-tuned circuit</td>
<td>0.4</td>
<td>0.88</td>
</tr>
<tr>
<td>Rectangular pulse</td>
<td>2 cascaded single-tuned stages</td>
<td>0.613</td>
<td>0.56</td>
</tr>
<tr>
<td>Rectangular pulse</td>
<td>5 cascaded single-tuned stages</td>
<td>0.672</td>
<td>0.5</td>
</tr>
</tbody>
</table>
MATCHED FILTER WITH NON-WHITE NOISE

\[ H(f) = G_a S^*(f) \exp \left( -j2\pi ft_1 \right) \frac{1}{N_i(f)} \left( \frac{S(f)}{N_i(f)} \right)^* \exp \left( -j2\pi ft_1 \right) \]
Correlation Detection

\[ y_0(t) = \int_{-\infty}^{\infty} y_{1n}(\lambda)s(t_1 - t + \lambda) \, d\lambda = R(t - t_1) \]
Block diagram of a cross-correlation receiver
RADAR RECEIVERS
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.

- The design of the radar receiver will depend not only on the type of waveform to be detected.
NOISE FIGURE AND NOISE TEMPERATURE:

NOISE FIGURE

\[ F_n = \frac{kT_0 B_n G + \Delta N}{kT_0 B_n G} = 1 + \frac{\Delta N}{kT_0 B_n G} \]

\[ F_n = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{N_{out}}{kT_0 B_n G} \]

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
- \( N_{out} \) = available output noise power
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

$$T_e = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \cdots$$

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

$$T_s = T_a + T_e = T_0 F_s$$

$$F_n = 1 + \frac{T_e}{T_0}$$

$$T_e = (F_n - 1) T_0$$
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-band pass 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.
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 than with a broadband "mixer in which the image frequency is terminated in a matched. The reactive the image frequency to be reflected back into the mixer and reconverted to IF.

Balanced mixer

Image-recovery mixer
LOW-NOISE FRONT-ENDS

Noise figures of typical microwave receiver front-ends as a function of frequency
5 DISPLAYS – TYPES

• 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.

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.

**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.
**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.
Branch-type duplexers

Principle of branch type duplexer
Balanced duplexer

Balanced duplexer using dual TR tubes and two short-slot hybrid junctions
(a) Transmit condition; (b) receive condition
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 as in Fig. 8. The isolation between the transmitter and receiver ports of a circulator is seldom sufficient to protect the receiver from damage.
The difference in the phase of the signals in adjacent elements is $\Psi = 2\pi \left( \frac{d}{\lambda} \right) \sin \theta$, where $\theta$ 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.

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

where $\omega$ is the angular frequency of the signal. The sum can be written

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

\[
|E_a(\theta)| = \left| \frac{\sin \left[ N\pi \left( \frac{d}{\lambda} \right) \sin \theta \right]}{\sin \left[ \pi \left( \frac{d}{\lambda} \right) \sin \theta \right]} \right|
\]
The difference in the phase of the signals in adjacent elements is \( \Psi = 2\pi \frac{d}{\lambda} \sin \theta \), where \( \theta \) is the direction.

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

\[
G_a(\theta) = \frac{|E_a|^2}{N^2} = \frac{\sin^2 \left[ N\pi \frac{d}{\lambda} \sin \theta \right]}{N^2 \sin^2 \left[ \pi \frac{d}{\lambda} \sin \theta \right]}
\]
Beam steering

- 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 $\theta = 0$. 

\[
G(\theta) = \sin^2 \left[ N \pi \left( \frac{d}{\lambda} \right) \left( \sin \theta - \sin \theta_0 \right) \right] \\
\]

![Diagram of beam steering](image)
BEAM WIDTH CHANGES:

Steering of an antenna beam with variable phase shifters (parallel-fed array).
APPLICATIONS OF PHASED ARRAY ANTENNAS:

- **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.
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 separate 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 side lobes 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.
Thank you
UNIT -1

BASICS OF RADAR
**RADAR INTRODUCTION**

- What is RADAR an acronym for? Radio Detection and Ranging.
- Radio wave is generated, transmitted, reflected, and detected.
- RADAR unimpaired by night, fog, clouds, smoke
- RADAR is good for isolated targets against a relatively featureless background.
A pulse traveling to a target at range $r_{\text{max}}$ and back will cover a distance $2r_{\text{max}}$.

The pulse will make it back to the radar before the next pulse is emitted if:

$$2R_{\text{max}} = \frac{c}{PRF} \quad \text{or} \quad R_{\text{max}} = \frac{c}{2 \cdot PRF}$$

maximum unambiguous range
Radar range equation

- Radar range equation relates the range of the radar to the characteristics of the Tx, Rx, antenna, target and the environment. It is used for radar system design.

- $P_t = \text{power radiated by an isotropic antenna}$

- Power density at a range $R$ from an isotropic antenna

\[
\frac{P_t}{4\pi R^2} \quad \text{W/m}^2
\]

- Power density from directive antenna

\[
\frac{P_t G}{4\pi R^2} \quad \text{W/m}^2
\]

- where $G$ is the directive gain
Isotropic Radiation Pattern

\[ \frac{P_t}{4\pi R^2} \text{ watts/m}^2 \]

Radar
Directional Radiation Pattern

\[ \frac{P_t G}{4\pi R^2} \text{ watts/m}^2 \]
Amount of power intercepted by the target = $\sigma$

It is also called the “radar cross section of the target”

It depend on the target’s shape, size and composition

Total power intercepted by the target $= \frac{P_t G \sigma}{4\pi R^2}$ W

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

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

Where $A_e$ is the effective area of the antenna.
The maximum range of the radar $R_{\text{max}}$ is the distance beyond which the target cannot be detected. This happens when $P_r = S_{\text{min}}$, minimum detectable signal.

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

This is called radar range equation.
## Radar frequency band designations

<table>
<thead>
<tr>
<th>Band designation</th>
<th>Nominal frequency range</th>
<th>Specific radar bands based on ITU assignment</th>
</tr>
</thead>
<tbody>
<tr>
<td>HF</td>
<td>3 – 30 MHz</td>
<td></td>
</tr>
<tr>
<td>VHF</td>
<td>30 – 300 MHz</td>
<td>138-144, 216-225 MHz</td>
</tr>
<tr>
<td>UHF</td>
<td>300 – 1000 MHz</td>
<td>420-450, 590-942 MHz</td>
</tr>
<tr>
<td>L</td>
<td>1 – 2 GHz</td>
<td>1215-1400 MHz</td>
</tr>
<tr>
<td>S</td>
<td>2 – 4 GHz</td>
<td>2300-2500, 2700-3700 MHz</td>
</tr>
<tr>
<td>C</td>
<td>4 – 8 GHz</td>
<td>5250-5925 MHz</td>
</tr>
<tr>
<td>X</td>
<td>8 – 12 GHz</td>
<td>8500-10680 MHz</td>
</tr>
<tr>
<td>Ku</td>
<td>12–18 GHz</td>
<td>13.4-14, 15.7-17.7 GHz</td>
</tr>
<tr>
<td>K</td>
<td>18 – 27 GHz</td>
<td>24.05-24.25 GHz</td>
</tr>
<tr>
<td>Ka</td>
<td>27 – 40 GHz</td>
<td>33.4-36 GHz</td>
</tr>
</tbody>
</table>
Applications

- Navigational aid on ground and sea
- Radar altimeters (height measurement)
- Radar blind lander (aircraft landing during poor visibility)
- Airborne radar for satellite surveillance
- Space applications like planetary observations
- Police radars (Law enforcement and Highway safety)
- Radars for determining speed of moving targets
- Remote sensing (weather monitoring)
- Air traffic control (ATC) and Aircraft safety
- Ship safety
- Non-contact method of speed and distance in industry
Military Applications:

- Detection and ranging of enemy targets even at night
- Aiming guns at aircrafts and ships
- Bombing ships, aircrafts, or cities even during night
- Early warning regarding approaching aircrafts or ships
- Directing guided missiles
- Searching for submarines, land masses and buoys
The simple form of the radar equation expressed the maximum radar range $R_{\text{max}}$, in terms of radar and target parameters.

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

where $P_t =$ transmitted power, watts
$G =$ antenna gain
$A_e =$ antenna effective aperture, m$^2$
$\sigma =$ radar cross section, m$^2$
$S_{\text{min}} =$ minimum detectable signal, watts
All the parameters are to some extent under the control of the radar designer, except for the target cross section $\sigma$.

- The radar equation states that if long ranges are desired,
- the transmitted power must be large, the radiated energy must be concentrated into a narrow beam (high transmitting antenna gain),
the received echo energy must be collected with a large antenna aperture (also synonymous with high gain),

- and the receiver must be sensitive to weak signals
the noise energy that occupies the same portion of the frequency spectrum as does the signal energy. The weakest signal the receiver can detect is called the minimum detectable signal.

Fig. 2.1 Typical envelope of tile radar receiver output as a function of time. A, B and C represent signal plus noise. A and B would be valid detections, but C is a missed detection.
Since noise is the chief factor limiting receiver sensitivity, it is necessary to obtain some means of describing it quantitatively.

Noise is unwanted electromagnetic energy which interferes with the ability of the receiver to detect the wanted signal.

It may originate within the receiver itself, or it may enter via the receiving antenna along with the desired signal.
If the radar were to operate in a perfectly noise-free environment so that no external sources of noise accompanied the desired signal, and if the receiver itself were so perfect that it did not generate any excess noise,

- The available thermal-noise power generated by a receiver of bandwidth \( B_n \) (in hertz) at a temperature \( T \) (degrees Kelvin) is equal to

\[
\text{Available thermal-noise power} = kTB_n
\]

An integrated bandwidth

\[
B_n = \frac{\int_{-\infty}^{\infty} |H(f)|^2 \, df}{|H(f_0)|^2}
\]
If the minimum detectable signal $S_{\text{min}}$ is that value of $S_i$ corresponding to the minimum ratio of output (IF) signal-to-noise ratio $(S_o/N_o)_{\text{min}}$ necessary for detection, then

$$S_{\text{min}} = k T_0 B_n F_n \left( \frac{S_o}{N_o} \right)_{\text{min}}$$

Substituting above Eq. discussed above into RADAR Eq. earlier results in the following form of the radar equation

$$R_{\text{max}}^4 = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_n F_n (S_o/N_o)_{\text{min}}}$$
UNIT -II
RADAR EQUATION
SIGNAL-TO-NOISE RATIO

- It is a measure of degradation of the signal to noise ratio (SNR), caused by components in the RF signal chain, for a given bandwidth.

- Consider an IF amplifier with bandwidth $B_{IF}$ followed by a second detector and a video amplifier with bandwidth $B_v$

![Fig. Envelope detector](image-url)
The noise entering the IF filter (the terms filter and amplifier are used interchangeably) is assumed to be gaussian, with probability-density function given by

\[ p(v) = \frac{1}{\sqrt{2\pi \psi_0}} \exp \left( -\frac{v^2}{2\psi_0} \right) \]

If gaussian noise were passed through a narrowband IF filter-onewhose bandwidth is small compared with the mid frequency-the probability density of the envelope of the noise voltage output is shown by Rice to be

\[ p(R) = \frac{R}{\psi_0} \exp \left( -\frac{R^2}{2\psi_0} \right) \]
where $R$ is the amplitude of the envelope of the filter output. Equation above is a form of the Rayleigh probability-density function. The probability that the envelope of the noise voltage will lie between the values of $V_1$ and $V_2$ is

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

The probability that the noise voltage envelope will exceed the voltage threshold $V_T$ is

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

$$= \exp \left( - \frac{V_T^2}{2\psi_0} \right) = P_f$$
The average time interval between crossings of the threshold by noise alone time is defined as the false-alarm time $T_{fa}$:

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

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

$$P_{fa} = \frac{\sum_{k=1}^{N} t_k}{\sum_{k=1}^{N} T_k} = \frac{\langle t_k \rangle_{av}}{\langle T_k \rangle_{av}} = \frac{1}{T_{fa} B}$$
Fig. Envelope of receiver output illustrating false alarms due to noise.
INTEGRATION OF RADAR PULSES

Many pulses are usually returned from any particular target on each radar scan and can be used to improve detection. The number of pulses $n_B$ returned from a point target as the radar antenna scans through its beamwidth is

$$n_B = \frac{\theta_B f_p}{\dot{\theta}_s} = \frac{\theta_B f_p}{6\omega_m}$$

where $\theta_n =$ antenna beamwidth, deg
$f_p =$ pulse repetition frequency, Hz
$\dot{\theta}_s =$ antenna scanning rate, deg/s
$\omega_m =$ antenna scan rate, rpm
The efficiency of postdetection integration relative to ideal predetection integration has been computed by Marcum when all pulses are of equal amplitude. The integration efficiency may be defined as follows:

\[ E_i(n) = \frac{(S/N)_1}{n(S/N)_n} \]

where:
- \( n \) = number of pulses integrated
- \( (S/N)_1 \) = value of signal-to-noise ratio of a single pulse required to produce given probability of detection (for \( n = 1 \))
- \( (S/N)_n \) = value of signal-to-noise ratio per pulse required to produce same probability of detection when \( n \) pulses are integrated
The radar cross section of a target is the (fictional) area intercepting that amount of power which when scattered equally in all directions, produces an echo at the radar equal to that from the target; or in other terms

\[
\sigma = \frac{\text{power reflected toward source/unit solid angle}}{\text{incident power density}/4\pi} = \lim_{R \to \infty} 4\pi R^2 \left| \frac{E_r}{E_i} \right|^2
\]

where \( R = \text{distance between radar and target} \)
\( E_r = \text{reflected field strength at radar} \)
\( E_i = \text{strength of incident field at target} \)
Fig. Radar cross section of the sphere. $a = \text{radius}; \lambda = \text{wavelength}$
The power $P_t$ in the radar equation is called by the radar engineer the peak power.

\[ P_{av} = \frac{P_t \tau}{T_p} = P_t \tau f_p \]

The ratio $P_{av}/P_t$, $\tau/T_p$, or $\tau f_p$ is called the duty cycle of the radar. A Pulse radar for detection of aircraft might have typically a duty cycle of 0.001, while a CW radar which transmits continuously has a duty cycle of unity. Writing the radar equation in terms of the average power rather than the peak power, we get

\[ R_{max}^4 = \frac{P_{av} GA_e \sigma n E_i(n)}{(4\pi)^2 k T_0 F_n(B_n \tau)(S/N)_{1,f_p}} \]
PRF AND RANGE AMBIGUITIES

- The pulse repetition frequency (prf) is determined primarily by the maximum range at which targets are expected.
- If the prf is made too high the likelihood of obtaining target echoes from the wrong pulse transmission is increased.
- Echo signals received after an interval exceeding the pulse-repetition period are called multiple-time-around echoes.
- They can result in erroneous or confusing range measurements. Consider the three targets labeled A, B, and C in Fig. 9a.
Target A is located within the maximum unambiguous range $R_{unamb}$ of the radar, target B is at a distance greater than $R_{unamb}$ but less than $2R_{unamb}$ while target C is greater than $2R_{unamb}$ but less than $3R_{unamb}$.

The appearance of the three targets on an A-scope is sketched in Fig. 9 b. The multiple-time-around echoes on the A-scope cannot be distinguished from proper target echoes actually within the maximum unambiguous range. Only the range measured for target A is correct; those for Band C are not.
Fig. 9 Multiple-time-around echoes that give rise to ambiguities in range (a) Three targets A, B and C, where A is within $R_{\text{unamb}}$, and B and C are multiple-time-around targets;
SYSTEM LOSSES

One of the important factors omitted from the simple radar equation was the losses that occur throughout the radar system.

The losses reduce the signal-to-noise ratio at the receiver output. They may be of two kinds, depending upon whether or not they can be predicted with any degree of precision beforehand.

100 ft of RG-113/U A1 waveguide transmission line (two-way)  
Loss due to poor connections (estimate)  
Rotary-joint loss  
Duplexer loss  
Total plumbing loss  

1.0 dB  
0.5 dB  
0.4 dB  
1.5 dB  
3.4 dB
The mathematical derivation of the collapsing loss, assuming a square-law detector may be carried out as suggested by Marcum who has shown that the integration of $m$ noise pulses, along with $n$ signal-plus-noise pulses with signal-to-noise ratio per pulse $(S/N)_n$, is equivalent to the integration of $m + n$ signal-to-noise pulses each with signal-to-noise ratio $n(S/N)_n/(m + n)$. The collapsing loss in this case is equal to the ratio of the integration loss $L_i$ for $m + n$ pulses to the integration loss for $n$ pulses, or

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

CW AND FREQUENCY MODULATED RADAR
DOPPLER EFFECT:

- A radar detects the presence of objects and locates their position in space by transmitting electromagnetic energy and observing the returned echo. A pulse radar transmits a relatively short burst of electromagnetic energy, after which the receiver is turned on to listen for the echo. The echo not only indicates that a target is present, but the time that elapses between the transmission of the pulse and the receipt of the echo is a measure of the distance to the target. Separation of the echo signal and the transmitted signal is made on the basis of differences in time.
The Doppler frequency shift is

\[ f_d = 2* \nu_r / \lambda = 2* \nu_r * f_o / c \]

where,
\[ f_o = \text{transmitted frequency} \]
\[ c = \text{velocity of propagation} = 3 \times 10^8 \text{m/s}. \]

If \( f_d \) is in hertz \( \nu_r \) in knots, and \( \lambda \) in meters,

\[ f_d = 1.03 * \nu_r / \lambda \]
\[ \omega_d = 2\pi f_d = \frac{d\phi}{dt} = \]

Where,
- \( f_d \) = Doppler frequency shift
- \( v_r \) = relative (or radial) velocity of target with respect to radar.
Fig. 3.2 (a) Simple CW radar block diagram; (b) response characteristic of beat-frequency amplifier
ISOLATION BETWEEN TRANSMITTER AND RECEIVER

- A single antenna serves the purpose of transmission and reception in the simple CW radar.
- In practice, it is not possible to eliminate completely the transmitter leakage. However, transmitter leakage is not always undesirable.
- There are two practical effects which limit the amount of transmitter leakage power which can be tolerated at the receiver.
- These are (1) the maximum amount of power the receiver input circuitry can withstand before it is physically damaged or its sensitivity reduced (burnout) and
- (2) the amount of transmitter noise due to hum, microphonics, stray pick-up, and instability which enters the receiver from the transmitter.
NON-ZERO IF RECEIVER
A bank of narrowband 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.

Because of foldover, a frequency which lies to one side of the IF carrier appears, after detection, at the same video frequency as one which lies an equal amount on the other side of the IF.

Therefore the sign of the doppler shift is lost
Fig. (a) Block diagram of IF doppler filter bank; (b) frequency-response characteristic of doppler filter bank.
The sign of the doppler frequency, and therefore the direction of target motion, may be found by splitting the received signal into two channels as shown in Fig. 3.4.

In channel A the signal is processed as in the simple CW radar. The received signal and a portion of the transmitter heterodyne in the detector (mixer) to yield a difference signal

\[ E_A = K_2 \ E_0 \cos (\pm w_d t + \phi) \]

\[ E_A = \text{amplitude of transmitter signal} \]

\[ K_2 = \text{a constant determined from the radar equation} \]
\( w_d \) = doppler angular frequency shift

\( \phi \) = a constant phase shift, which depends upon range of initial detection

The other channel is similar, except for a 90° phase delay introduced in the reference signal. The output of the channel B mixer is

\[
E_B = K_2 \ E_0 \cos(\pm w_d t + \phi + \pi / 2)
\]

If the target is approaching (positive Doppler), the outputs from the two channels are

\[
E_A(+) = K_2 \ E_0 \cos(w_d t + \phi)
\]

\[
E_B(+) = K_2 \ E_0 \cos(w_d t + \phi + \pi / 2)
\]

If the targets are receding (negative doppler), the outputs from the two channels are

\[
E_A(-) = K_2 \ E_0 \cos(w_d t - \phi)
\]

\[
E_B(-) = K_2 \ E_0 \cos(w_d t - \phi - \pi / 2)
\]
Fig. Measurement of Doppler direction using synchronous, two-phase motor
APPLICATIONS OF CW RADAR

- The chief use of the simple, unmodulated CW radar is for the measurement of the relative velocity of a moving target, as in the police speed monitor or in the previously mentioned rate-of-climb meter for vertical-take-off aircraft.

- In support of auto-mobile traffic, CW radar has been suggested for the control of traffic lights, regulation of tollbooths, vehicle counting, as a replacement for the fifth-wheel speedometer in vehicle testing as a sensor in antilock braking systems, and for collision avoidance.

- For railways, CW radar can be used as a speedometer to replace the conventional axle-driven tachometer. In such an application it would be unaffected by errors caused by wheelslip on accelerating or wheelslide when braking.
It has been used for the measurement of railroad-freight-car velocity during humping operations in marshalling yards, and as a detection device to give track maintenance personnel advance warning of approaching trains.

CW radar is also employed for monitoring the docking speed of large ships.

It has also seen application for intruder alarms and for the measurement of the velocity of missiles, ammunition, and baseballs.

The principal advantage of a CW doppler radar over other (nonradar) methods of measuring speed is that there need not be any physical contact with the object whose speed is been measured.

In industry this has been applied to the measurement of turbine-blade vibration, the peripheral speed of grinding wheels, and the monitoring of vibrations in the cables of suspension bridges.

High-power CW radars for the detection of aircraft and other targets have been developed and have been used in such systems as the Hawk missile systems.
UNIT - IV
FM-CW RADAR
RANGE AND DOPPLER MEASUREMENT

- 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. 4.2(a). If there is a reflecting object at a distance R, an echo signal will return after a time $T = \frac{2R}{c}$.

- 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

$$f_r = f_0 T = 2 \frac{R}{c} f_0 / c$$
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. 4.2(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. 4.2 (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 \( \Delta f \), the beat frequency is

\[
 f_r = 2 * 2 R f_m / c = 4 R f_m \Delta f / c
\]

Thus the measurement of the beat frequency determines the range \( R \).

\[
 R = c f_r / 4 f_m \Delta f
\]
Fig. 4.2 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).
DIAGRAM AND CHARACTERISTICS
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.
On decreasing portion, the beat frequency, $f_b(\text{down})$ is the sum of the two.

$$f_b(\text{up}) = f_r - f_d$$

$$f_b(\text{down}) = f_r + f_d$$

The range frequency $f_r$, may be extracted by measuring the average beat frequency; that is,

$$f_r = \frac{1}{2}[f_b(\text{up}) + f_b(\text{down})].$$

If $f_b(\text{up})$ and $f_b(\text{down})$ 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 $f_r > f_d$. 
Fig 4.1.2 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
FM–CW 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 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 receivers shown in Fig. 4.3. A portion of the frequency–modulated transmitted signal is applied to a mixer along with the oscillator signal.

The selection of the local–oscillator frequency 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.
SINUSOIDALLY MODULATED FM–CW RADAR
EXTRACTING THE THIRD HARMONIC
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 $f_m$. 
Assume the form of the transmitted signal to be

\[ \sin \left( 2\pi f_0 t + \frac{\Delta f}{2f_m} \sin 2\pi f_m t \right) \]

where,

\( f_0 = \) carrier frequency

\( f_m = \) modulation frequency

\( \Delta f = \) frequency excursion (equal to twice the frequency derivation)

\[ v_D = J_0(D) \cos (2\pi f_d t - \phi_0) + 2J_1(D) \sin (2\pi f_d t - \phi_0) \cos (2\pi f_m t - \phi_m) \]

\[ -2J_2(D) \cos (2\pi f_d t - \phi_0) \cos 2(2\pi f_m t - \phi_m) \]

\[ -2J_3(D) \sin (2\pi f_d t - \phi_0) \cos 3(2\pi f_m t - \phi_m) \]

\[ + 2J_4(D) \cos (2\pi f_d t - \phi_0) \cos 4(2\pi f_m t - \phi_m) + 2J_5(D) \cdots \]
where \( J_0, J_1, J_2, \) etc = Bessel functions of first kind and order 0, 1, 2, etc., respectively

\[
D = \left( \frac{\Delta f}{f_m} \right) \sin 2\pi f_m R_0 / c
\]

\( R_0 = \) 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

\( \phi_0 = \) phase shift approximately equal to angular distance \( 2\pi f_0 R_0 / c \)

\( \phi_m = \) phase shift approximately equal to \( 2\pi f_m R_0 / c \)
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 $\Delta f$. 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 $\phi_1$ and $\phi_2$ are arbitrary (constant) phase angles.
\[ v_{1R} = \sin \left[ 2\pi (f_1 \pm f_{d1})t - \frac{4\pi f_1 R_0}{c} + \phi_1 \right] \]

\[ v_{2R} = \sin \left[ 2\pi (f_2 \pm f_{d2})t - \frac{4\pi f_2 R_0}{c} + \phi_2 \right] \]

Where,

\[ R_0 = \text{range to target at a particular time } t = t_0 \quad (\text{range target were not moving}) \]

\[ f_{d1} = \text{doppler frequency shift associated with frequency } f_1 \]

\[ f_{d2} = \text{doppler frequency shift associated with frequency } f_2 \]
The phase difference between these two components is

\[ \Delta \phi = \frac{4\pi (f_2 - f_1) R_0}{c} = \frac{4\pi \Delta f R_0}{c} \]

The maximum unambiguous range \( R_{\text{unamb}} \) is

\[ R_{\text{unamb}} = \frac{c}{2\Delta f} \]
UNWANTED SIGNALS IN FM ALTIMETER