ॐ RADAR ENGINEERING

## UNIT - I

## RADAR PRINCIPLES

### Introduction to Radar :-

Radar is a electromagnetic system for detection and location of the object. It operates by transmitting a particular type of waveform, a pulse modulated sine waveform for example and detects the nature of the echo signal. Radar is used to extend the capability of one's sense for observing the environment, especially the sense of vision.

An elementary form of radar consists of a transmitting antenna emitting electromagnetic radiation generated by an oscillator, a receiving antenna and an energy detecting device or receiver. A portion of the transmitted signal is intercepted by a reflecting object (target) and is reradiated in all directions. It is the energy reradiated in the back direction that is of prime interest to the radar. the receiving antenna collects the returned energy and delivers it to a receiver, where it is processed to detect the presence of the target and to extract its location and relative velocity. the distance to the target is determined by measuring the time taken for the radar signal to travel to the target and back.

The direction, or angular position, of the target may be determined from the direction of arrival of the reflected wave-front. Radar is defined technically as Radio detection and Ranging. It is generally developed for the detecting object.

The most common radar waveform is a train of narrow, rectangular-shaped pulses modulating a sinewave carrier. The distance or range to the target is determined by measuring the time $T_R$ taken by the pulse to travel to the target and return. since electromagnetic energy propagates at the speed of light $c = 3\times10^8$ m/s the Range R is

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

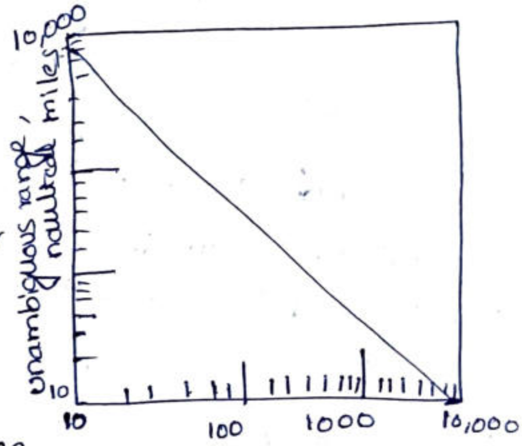
The factor 2 appears in the denominator bcoz of the two way propagation of radar. R in terms of kilometers & nautical miles the $T_R$ value will be in micro seconds.

$$R(km) = 0.15\,T_R(\mu s) \qquad R(nmi) = 0.081\,T_R(\mu s).$$

Once the transmitted pulse is emitted by the radar, a sufficient length of time must elapse to allow any echo signals to return and be detected before the next pulse may be transmitted. If the pulse repetition frequency is too high, echo signals from some targets might arrive after the transmission of the next pulse, and ambiguities in measuring range might result. Echoes that arrive after the transmission of the next pulse is called second-time-around (or) multiple time around echoes.

Such an echo would appear to be at a much shorter range than the actual and could be misleading if it were not known to be a second-time-around echo. The range beyond which targets appear as second time around echoes is called the maximum unambiguous range & is $\boxed{R_{unamb} = \frac{c}{2f_p}}$

where $f_p$ = pulse repetition frequency.




Although the typical radar transmits a simple pulse modulated waveform, there are a number of suitable modulations that might be used. The pulse carrier might be frequency or phase modulated.

The technique of using a long, modulated pulse to obtain the resolution of a short pulse, but with the energy of a long pulse is known as pulse compression. Unmodulated CW waveforms do not measure range, but a range measurement can be made by applying either frequency or phase modulation.

## The simple form of Radar Equation :-

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

If the power of the radar transmitter is denoted by $P_t$, and if an isotropic antenna is used one which radiates uniformly in all directions, the power density at a distance R from the radar is equal to the

power density from isotropic antenna $= \frac{P_t}{4\pi R^2}$ watts/unit area →(1)

Radars employ directive antennas to channel, the radiated power $P_t$ into some particular direction. the gain $G$ of an antenna is a measure of the increased power radiated in the direction of the target as compared with the power that would have been radiated from an isotropic antenna.

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

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

$$\text{power density of echo signal at radar} = \frac{P_t G}{4\pi R^2} \frac{\sigma}{4\pi R^2} \longrightarrow ③$$

The radar cross section $\sigma$ has units of area. It is a characteristic of the particular target and is a measure of its size as seen by the radar. If the radar antenna captures a portion of the echo power. If the effective area of the receiving antenna is denoted $A_e$ the power $P_r$ received by the radar is

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

The max radar range $R_{max}$ is the distance beyond which the target cannot be detected. it occurs when the received echo signal power $P_r$ just equals the min detectable signal $S_{min}$.

$$\boxed{R_{max} = \left[\frac{P_t G A_e \sigma}{(4\pi)^2 S_{min}}\right]^{1/4}} \longrightarrow ⑤$$

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

$$G = \frac{4\pi A_e}{\lambda^2} \longrightarrow ⑥ \qquad A_e = \frac{G\lambda^2}{4\pi}$$

since radars generally use the same antenna for both transmission and reception. first for $A_e$ then for $G$, to give two other forms of the radar equation

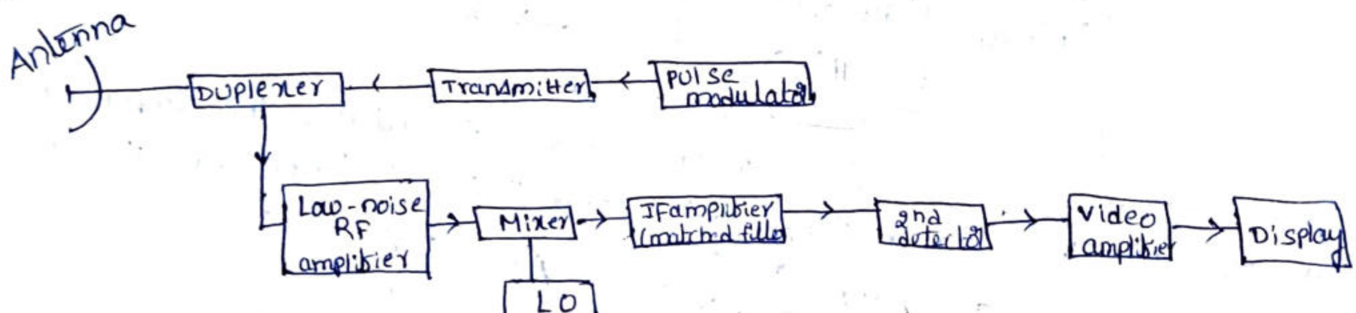
$$\boxed{\begin{aligned} R_{max} &= \left[\frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 S_{min}}\right]^{1/4} \longrightarrow ⑦ \\ R_{max} &= \left[\frac{P_t A_e^2 \sigma}{4\pi \lambda^2 S_{min}}\right]^{1/4} \longrightarrow ⑧ \end{aligned}}$$

from the equation 7 & 8 $\lambda^{1/2}$ varies from equation 7 & $\lambda^{-1/2}$ from equation 8 so the equation 5 shows range to be independent of $\lambda$. the correct relationship depends on whether it is assumed the gain is constant or the effective area is constant with wavelength.

## RADAR Block diagram and operation:-

The transmitter may be an oscillator, such as magnetron, that is pulsed (turned on and off) by the modulator to generate a repetitive train of pulses. the magnetron has probably been the most widely used of the various microwave generators for radar. The waveform generated by the transmitter travels via a transmission line to the antenna, where it is radiated into space. A single antenna is generally used for both transmitting and receiving.

The receiver must be protected from damaged caused by the high power of the transmitter. this is a function of the duplexer. the duplexer also serves to channel the returned echo signals to the receiver and not to be transmitter. the duplexer might consist of two gas-discharge devices, one known as a TR (transmit receive) and the other an ATR (anti transmit receive). the TR protects the receiver during transmission and the ATR directs the echo signal to the receiver during reception. solid state ferrite circulators and receiver protectors with gas plasma TR devices and/or diode limiters are also employed as duplexers.




Block diagram of a pulse radar.

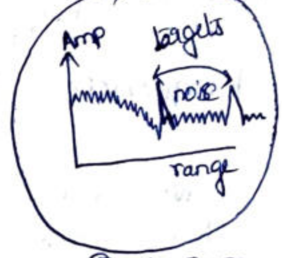
The receiver is usually of the superheterodyne type. the first stage might be a low noise RF amplifier, such as a parametric amplifier or a low-noise transistor. the receiver I/P can simply be the mixer stage, Although a mixer receiver with a low noise front-end will be more sensitive, the mixer I/P can have greater dynamic range, less susceptibility to overload, and less vulnerability to electronic interference.

The mixer and local oscillator (LO) convert the RF signal to an Intermediate frequency (IF). the IF amplifier should be designed as a 'matched filter ie, its frequency response function H (f) should maximize the peak signal-to-mean noise power ratio at the output.

After maximizing the signal-to noise ratio in the IF amplifier, the pulse modulation is extracted by the second detector and amplified by the video amplifier to a level where it can be properly displayed, usually on a cathode ray tube (CRT). timing signals are also provided to the Indicator to provide the range zero. Angle Information is obtained from the pointing direction of the antenna. the most common form of CRT's tube display is the plan position Indicator PPI.

A common form of radar antenna is a reflector with a parabolic shape, fed from a point source at its focus. the parabolic reflector ~~with a parabolic~~ focuses the energy into a narrow beam. phase array antennas have also been used for radar.


(a) PPI display

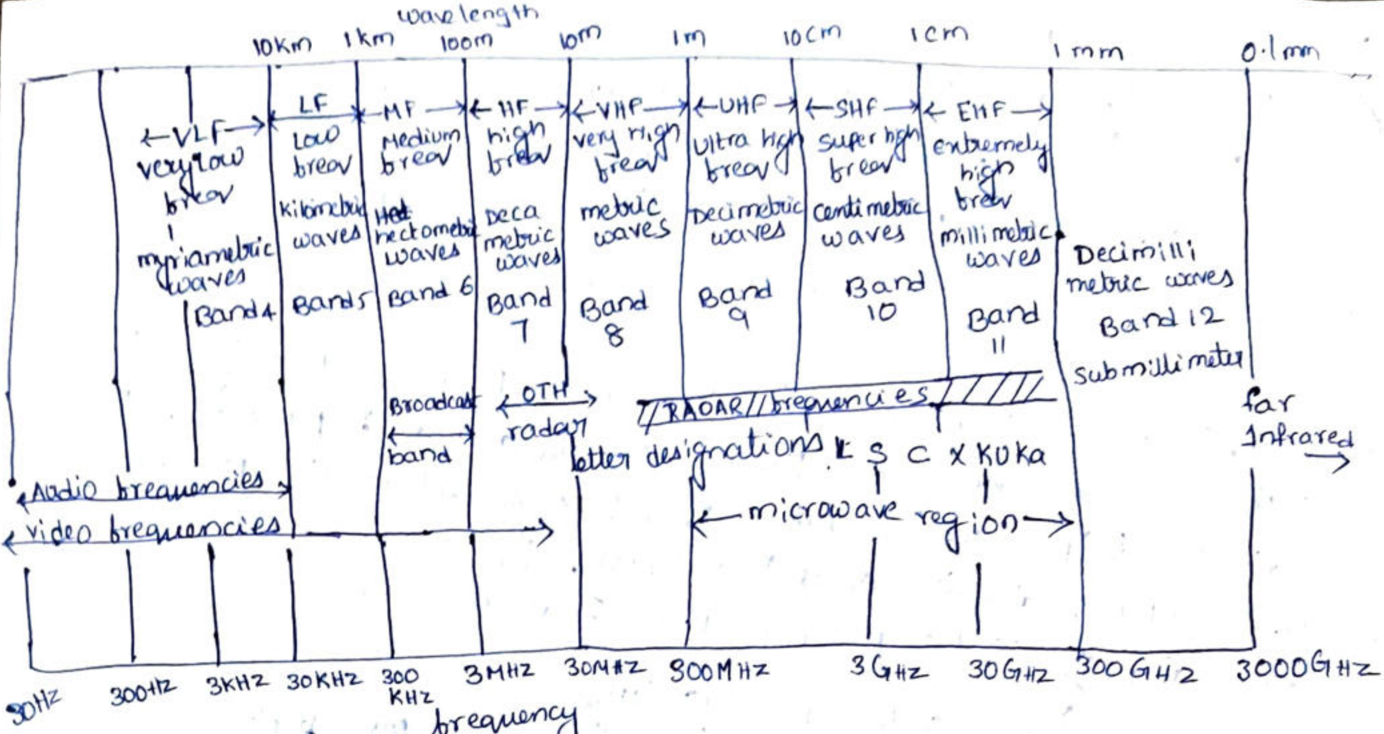



(b) A-scope

## Radar frequencies :-

conventional radars generally have been operated at frequencies extending from about 220 MHz to 35 GHz, a spread of more than seven octaves. Skywave HF over-the horizon (OTH) radar might be at frequencies as low as 4 or 5 MHz, and groundwave HF radars as low as 2 MHz. at the other end of the spectrum, millimeter radars have operated at 94 GHz. laser radars operate at even higher frequencies.

Early in the development of radar, a letter code such as S, X, L etc, was employed to designate radar frequency bands. Although its original purpose was to guard military secrecy, the designations were maintained, probably out of habits as well as the need for some convenient short nomenclature. Although the nominal frequency range for L band is 1000 to 2000 MHz, an L-band radar is thought of as being confined within the region from 1215 to 1400 MHz since that is the extent of the assigned band. letter based nomenclature is not a substitute for the actual numerical frequency limits of radars.




Radar frequencies and the electromagnetic spectrum.

| Band designation | Nominal freq range | Specific radiolocation (radar) bands based on ITU assignments for region 2 |
|---|---|---|
| HF | 3-30MHz | |
| VHF | 30-300MHz | 138 - 144MHz<br>216 - 225 |
| UHF | 300 - 1000MHz | 420 - 450 MHz<br>890 - 942 |
| L | 1000 - 2000MHz | 1215 - 1400 MHz |
| S | 2000 - 4000MHz | 2300 - 2500 MHz<br>2700 - 3700. |
| C | 4000 - 8000 MHz | 5250 - 5925MHz |
| X | 8000 MHz - 12,000MHz | 8500 - 10,680 MHz |
| Ku | 12.0 - 18 GHz | 13.4 - 14.0 GHz<br>15.7 - 17.7 |
| K | 18 - 27GHz | 24.05 - 24.25 GHz |
| Ka | 27 - 40 GHz | 33.4 - 36.0 GHz |
| mm | 40 - 300GHz | |

standard radar frequency letter band nomenclature.

## Applications of Radar:-

Radar has been employed on the ground, in the air, on the sea, and in space. ground-based radar has been applied chiefly to the detection, location, and tracking of aircraft or space targets. shipboard radar is used as a navigation aid and safety device to locate buoys, shore lines, and other ships, as well as observing aircrafts.

Airborne radar may be used to detect other aircrafts, ships or land vehicles, or it may be used for mapping of land, storm avoidance, terrain avoidance and navigation. In space, radar has assisted in the guidance of space craft and for the remote sensing of the land and sea.

① Air traffic control (ATC):- Radars are employed throughout the world for the purpose of safely controlling air crafts en route and in the vicinity of airports. Aircraft and ground vehicular traffic at large airports are monitored by means of high resolution radars. radar has been used with GCA (ground control approach) systems to guide aircrafts to a safe landing in bad weather. In addition, the the microwave landing system and the widely used ATC radar beacon system are based in large part on radar technology.

② Aircraft navigation:- The weather-avoidance radar used on aircraft to outline regions of precipitation to the pilot is a classical form of radar. radar also used for terrain avoidance and terrain following. the radio altimeter (either FM/CW or pulse) and the doppler navigators are also radars. Sometimes ground-mapping radars of moderately high resolution are used for aircrafts navigation purpose.

③ ship safety:- Radar is used for enhancing the safety of ship travel by warning of potential collision with other ships, and for detecting navigation buoys, especially in poor visibility. Automatic detection and tracking equipments (also called plot extractors) are commercially available for use with such radars for the purpose of collision avoidance.

④ space:- space vehicles have used radar for rendezvous and docking, and for landing on the moon. some of the largest ground-based radars are for the detection and tracking of satellites.

⑤ Remote sensing:- All radars are remote sensors. It implies the sensing of geophysical objects, or the "environment". radars sometime have been used as a remote sensor of the weather. remote sensing with radar is also concerned with earth resources, which includes the measurement and mapping of sea conditions, water resources, ice cover, agriculture, forestry conditions, geological formations and environmental pollution.

⑥ law enforcement:- In addition to the wide use of radars to measure the speed of automobile traffic by highway police, radar has also been employed as a means for the detection of intruders.

⑦ Military:- Many of the civilian applications of radar are also employed by the military. the traditional role of radar for military application has been for surveillance, navigation and for the control and guidance of weapons, the largest use of radar.

## Prediction of Range performance :-

→ The simple form of the radar equation expressed the maximum radar range $R_{max}$ in terms of radar and target parameters

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

where $P_t$ = transmitted power, watts,
$G$ = Antenna gain.
$A_e$ = antenna effective aperture, $m^2$
$\sigma$ = radar cross section, $m^2$.
$S_{min}$ = minimum detectable signal, watts

All the parameters are to some extent under the control of the radar
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 sensitive collected with a large antenna aperture and the receiver
must be sensitive to weak signals.

In practice, however, the simple radar equation does not predict
the range performance of actual radar equipments to a satisfactory
degree of accuracy. the predicted values of radar range are usually
optimistic. Another important factor that must be considered in the
radar equation is the statistical or unpredictable nature of several of the
parameters.

The minimum detectable signal $S_{min}$ and the target cross section
$\sigma$ are both statistical in nature and must be expressed in statistical
terms. the statistical nature of these several parameters does not allow
maximum radar range to be described by a single number. its specification
must include a statement of the probability that the radar will detect
a certain type of target at a particular range.

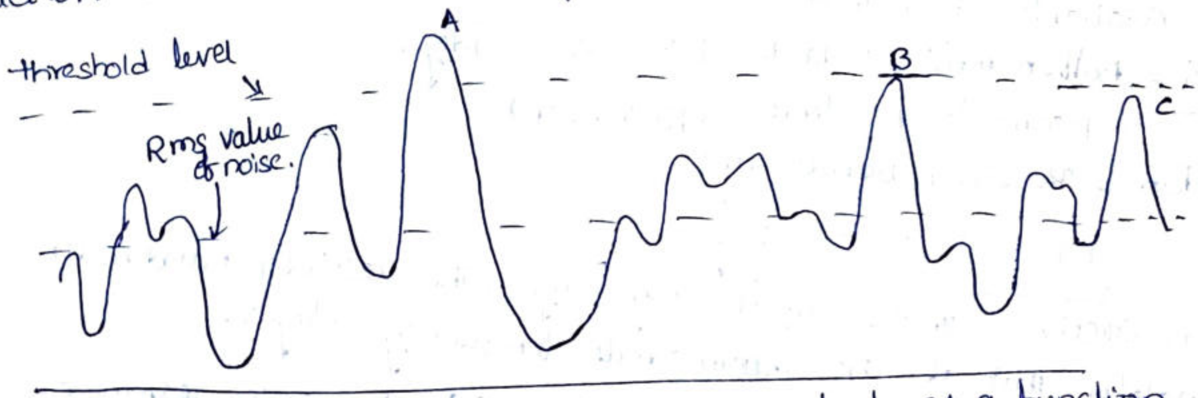
## Minimum Detectable signal :-

The ability of a radar receiver to detect a weak echo signal is
limited by 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 dectable signal. the
specifications of the minimum detectable signal is sometimes difficult
bcoz of its statistical nature and because the criterion for deciding
whether the target is present or not may not be too well defined.

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. this is called threshold detection.

→ consider the output of a typical radar receiver as a function of time as shown in the fig. this might represent one sweep of the video o/p displayed on an A-scope. the envelope has a fluctuating appearance caused by the random nature of noise. If a large signal is present such as at A, it is greater than the surrounding noise peaks and can be recognized on the basis of its amplitude. thus, if the threshold level were set sufficiently high, the envelope would not generally exceed the threshold if noise alone were present, but would exceed it if a strong signal were present.

→ The threshold level must be low if weak signals are to be detected, but it cannot be so low that noise peaks cross the threshold and give a false indiction of the presence of targets.




Typical envelope of the radar receiver output as a function of time.
A and B and C represents signal plus noise. A & B would be valid detections, but C is a missed detection.

→ A target is said to be detected if the envelope crosses the threshold. if the signal is large such as at A, it is not difficult to decide that a target is present. but consider the two signals at B and C, representing target echos of equal amplitude the noise voltage accompanying the signal at B is large enough so that the combination of signal plus noise exceeds the threshold. At C the noise is not as large and the resultant signal plus noise does not cross the threshold.

→ the selection of the proper threshold level is a compromise that depends upon how important it is if a mistake is made either by ① failing to recognize a signal that is present ② falsely indicating the presence of signal when none exists (probability of a false alarm).

The signal to Noise ratio necessary to provide adequate detection is one of the Important parameters that must be determined in order to compute the minimum detectable signal.

## Receiver Noise:-

→ 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, there would still exist an unavoidable component of noise generated by the thermal motion of the conduction electrons in the ohmic portions of the receiver I/P stages, This is called thermal noise or Johnson noise, and is directly proportional to the temperature of the ohmic portions of the ckt and the receiver bandwidth.

Available thermal Noise power $= KTB_n$.

where $K$ = Boltzmann's constant $= 1.38 \times 10^{-23}$ J/deg.

$T$ = Room temperature (290 K, 62°F)

$B_n$ = Receiver Bandwidth

$$\boxed{KT = 4 \times 10^{-21}\ \omega/Hz}$$

For radar receivers of the superhetrodyne type, the receiver bandwidth is approximately that of the Intermediate frequency stages.

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

where $H(f)$ = frequency response characteristic of IF amplifier (filter).

$f_0$ = frequency of maximum response.

When $H(f)$ is normalized to unity at midband (max response freq) $H(f_0) = 1$ the Bandwidth $B_n$ is called the noise bandwidth and is the B.W of an equivalent rectangular filter whose noise power O/P is the same as the filter with characteristic $H(f)$.

The noise power in practical receivers is often greater than can be accounted for by thermal noise alone. the addition noise components are due to mechanisms other than the thermal agitation of the conduction electrons.

The Noise figure $F_n$ of a receiver is defined by

$$F_n = \frac{N_o}{KT_oB_nG_a} = \frac{\text{Noise out of practical receiver}}{\text{noise out of Ideal receiver at std temp } T_o}$$

where $N_o$ = noise output from receiver.

$G_a$ = available gain. the standard temperature $T_o$ is taken to be 290K.

The noise No is measured over the linear portion of the receiver I/P / O/P characteristic; usually at the O/P of the IF amplifier before the nonlinear second detector.

The receiver BW $B_n$ is that of the IF amplifier in most receivers. The available gain $G_a$ is the ratio of the signal out $S_o$ to the signal in $S_i$ and $KT_oB_n$ is the I/P noise $N_i$ in an Ideal receiver.

$$F_n = \frac{S_i/N_i}{S_o/N_o}$$

the I/P signal may be expressed as

$$S_i = \frac{KT_oB_nF_nS_o}{N_o}$$

If the min detectable signal $S_{min}$ is that value of $S_i$ corresponding to the min ratio of O/P (IF) signal-to-noise ratio $(S_o/N_o)_{min}$ necessary for detection, then

$$S_{min} = KT_oB_nF_n\left(\frac{S_o}{N_o}\right)_{min}.$$

So by substituting in radar equation we get.

$$\boxed{R^4_{max} = \frac{P_t G A_e \sigma}{(4\pi)^2 KT_oB_nF_n(S_o/N_o)_{min}}}$$

~~Signal~~

## Signal-to-Noise Ratio:-

→ the signal noise theory can be applied to the radar receiver especially at higher stage to determine the signal to the noise ratio. the signal to the noise ratio can be determined at the O/P of IF amplifier based on probability detection. Consider a envelope detector which consist of second detector and video amplifier as shown in fig.

→ the B.W of IF amp can be denoted as $B_{IF}$ and the B.W of video amplifier is $B_v$.


Envelope detector.

→ Assume the B.W of video amplifier is wide in order to obtain the B.W frequency components. Generally B.W of video amplifier is greater than $\frac{B_{IF}}{2}$. The envelope detector detects or extracts the envelope of the msg signal (transmitted signal) without the carrier signal. the noise signal entering in to the receiver at IF amplifier is considered as gussian noise with probability density function given by

$$P(v) = \frac{1}{\sqrt{2\pi\psi_o}} \exp\left(\frac{-v^2}{2\psi_o}\right) \quad (1)$$

→ finding where $P(v)dv$ is the probability of the noise voltage $v$ b/w the values of $v$ and $v+dv$.

$\psi_o$ is the variance or mean square value of the noise voltage, and the mean value of $v$ is taken to be zero.

when the guassian noise is transmitted to the IF amplifier having a lesser bandwidth then the o/p of IF amplifier probability density function is given as $P(R) = \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) \rightarrow ②$

where R is the amplitude of the envelope filter O/P. It is a Rayleigh probability density function.

The probability that the envelope of the noise voltage will lie b/w the values of $V_1$ and $V_2$ is

$$\text{probability}\ (V_1 < R < V_2) = \int_{V_1}^{V_2} \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) dR \rightarrow ③$$

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

$$\text{probability}\ (V_T < R < \alpha) = \int_{V_T}^{\alpha} \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) dR \rightarrow ④$$

$$= \exp\left(-\frac{V_T^2}{2\psi_0}\right) = P_{fa} \rightarrow ⑤$$

whenever the voltage envelope exceeds the threshold, a target detection is considered to have occured. the avg time interval b/w crossings of the threshold by noise alone is defined as the false alarm time $T_{fa}$.

$$T_{fa} = \lim_{N\to\alpha} \frac{1}{N} \sum_{k=1}^{N} T_K \rightarrow ⑥$$

where $T_K$ is the time b/w crossings of the threshold $V_T$ by the noise envelope.

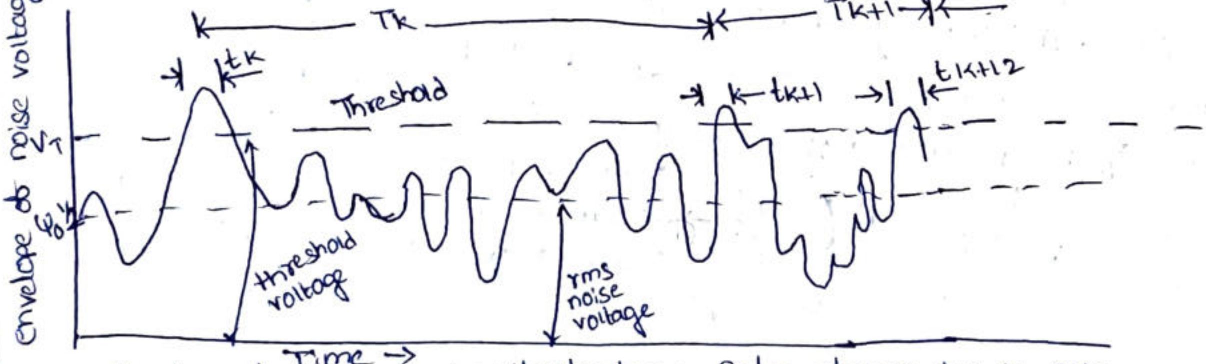
The false alarm probability may also be defined as the ratio of the duration of time the envelope is actually above the threshold to the total time it could have been the threshold &

$$P_{fa} = \frac{\sum_{K=1}^{N} t_K}{\sum_{K=1}^{N} T_K} = \frac{\langle t_K \rangle_{av}}{\langle T_K \rangle_{av}} = \frac{1}{T_{fa} B} \rightarrow ⑦$$

The average duration of a noise pulse is approximately the reciprocal of the Bandwidth B, in enevelope detector $B_{IF}$. equating eq ⑤ & ⑦ we get

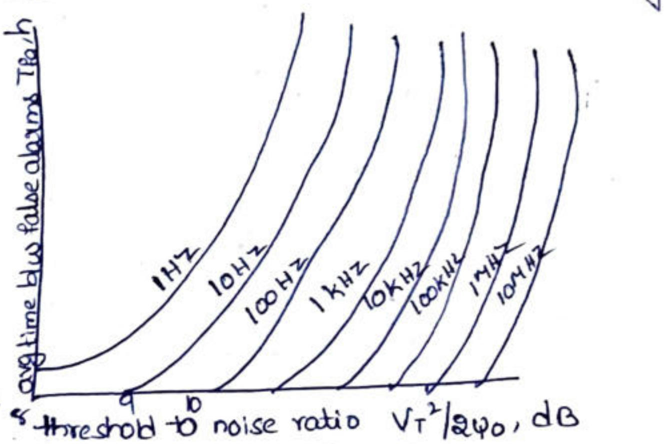
$$T_{fa} = \frac{1}{B_{IF}} \exp \frac{V_T^2}{2\psi_0} \rightarrow ⑧$$

where $V_T^2/2\psi_0$ as the abscissa.




envelope of receiver O/P illustrating false alarms due to noise.

The false alarm probabilities of practical radars are quite small. Let us consider a sine wave signal of amplitude A to be present along with noise at the I/P to the IF filter. the frequency of the signal is the same as the IF midband freq $f_{IF}$. the O/P of the envelope detector has a probability density function




$$P_s(R) = \frac{R}{\psi_0} \exp\left(-\frac{R^2+A^2}{2\psi_0}\right) I_0\left(\frac{RA}{\psi_0}\right) \rightarrow (8)$$

where $I_0(z) = \frac{e^z}{\sqrt{2\pi z}}\left(1+\frac{1}{8z}+\cdots\right)$ $\rightarrow (9)$ is a modified Bessel function of zero order and argument z.

The probability of detection $P_d$ is therefore

$$P_d = \int_{V_T}^{\infty} P_s(R)\, dR = \int_{V_T}^{\infty} \frac{R}{\psi} \exp\left(-\frac{R^2+A^2}{2\psi_0}\right) I_0\left(\frac{RA}{\psi_0}\right) dR \rightarrow (10)$$

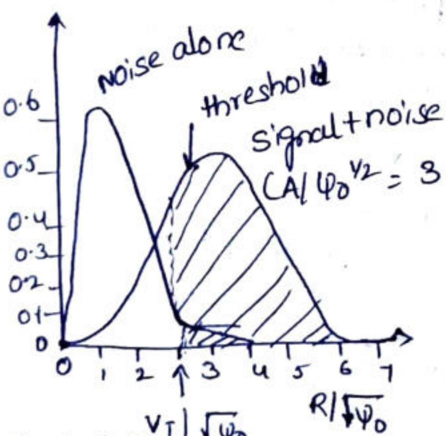
This cannot be evaluated by simple means, and numerical techniques or a series approximation must be used. A series approximation is valid when $RA/\psi_0 \gg 1$, $A \gg |R-A|$ and terms in $A^{-3}$ and beyond can be neglected.

$$P_d = \frac{1}{2}\left(1-\operatorname{erf}\frac{V_T-A}{\sqrt{2\psi_0}}\right) + \frac{\exp\left[-(V_T-A)^2/2\psi_0\right]}{2\sqrt{2\pi}\,(A/\sqrt{\psi_0})} \times \left[1-\frac{V_T-A}{4A}+\frac{1+(V_T-A)^2\psi_0}{8A^2/\psi_0}-\cdots\right] \rightarrow (11)$$

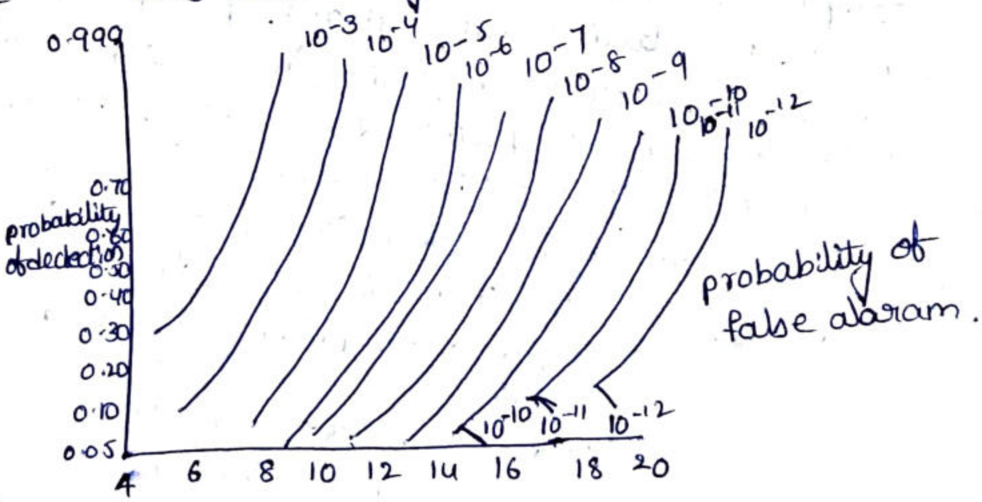
where the error function is defined as $\operatorname{erf} z = \frac{2}{\sqrt{\pi}}\int_0^z e^{-u^2}\, du$.

the eq (11) may be converted to power by replacing the signal to rms noise voltage ratio.

$$\frac{A}{\psi_0^{1/2}} = \frac{\text{signal Amplitude}}{\text{rms noise voltage}} = \frac{\sqrt{2}\,(\text{rms signal voltage})}{\text{rms noise voltage}} = \left(2\,\frac{\text{signal Power}}{\text{noise power}}\right)^{1/2} = \left(\frac{2S}{N}\right)^{1/2}.$$




Probability density function for noise alone & for signal plus noise.




Both the false alarm time and the detection probability are specified by the system requirements.

# Integration of Radar pulses :-

The relationship b/w signal to noise ratio, the probability of detection, and the probability of false alarm are applies for a single ~~plu~~ pulse only. however, 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

$$n_B = \frac{\theta_B f_P}{\theta_S} = \frac{\theta_B f_P}{6\omega m}$$

where $\theta_B$ = antenna beamwidth, deg
$f_P$ = pulse repetition freq, Hz.
$\theta_S$ = antenna scanning rate, deg/s.
$\omega_m$ = antenna scan rate, rpm.

Typical parameters for a ground based search radar might be pulse repetation frequency 300Hz, 1.5° beamwidth, and antenna scan rate 5rpm (30°/s). these parameters result in 15 hits from a point target on each scan. The process of summing all the radar echo pulses for the purpose of Improving detection is called Integration.

The most common radar Integration method is the cathode-ray tube display combined with the Integrating properties of the eye & brain of the radar operator.

Integration may be accomplished in the radar receiver either before the second detector or after the second detector. Integration before the detector is called predetection or coherent Integration, while Integration after the detector is called postdetection or noncoherent, Integration.

The comparison of predection and postdetection Integration may be briefly summarized by starting that although postdetection Integration is not as efficient as predetection Integration, it is easier to Implement in most applications. post detection Integration is preferred even though the Integrated signal-to noise ratio may not be as great. The Integration efficiency may be defined as

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

where n = no 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 per pulse required to produce same probability of detection where n pulses are Integrated.

The Improvement in the signal-to-noise ratio when n pulses are Integrated postdetection is $nE_i(n)$ and is the Integration Improvement factor.

As we know the radar range equation in terms of signal to the noise ratio.

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

similarly

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

If n pulses are integrated, the voltage out of the integrator is

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

where $V_i$ is the voltage amplitude of the $i$th pulse and $\exp(-\gamma)$ is the attenuation per pulse.

In an RC low pass filter $\gamma = T_P/RC$, where $T_p$ is the pulse repetition period and RC is the filter time constant. An efficiency will be defined which is the ratio of the average signal to noise ratio for the exponential integrator to the average signal to noise ratio for the uniform integrator.

$$\rho = \frac{\tanh(n\gamma/2)}{n\tanh(\gamma/2)}$$

An integrator that dumps is an electrostatic storage tube that is erased whenever it is read. The efficiency of an integrator that operates continuously without dumping is

$$\rho = \frac{[1-\exp(-n\gamma)]^2}{n\tanh(\gamma/2)}$$

The maximum efficiency of a dumped integrator occurs for $\gamma = 0$, but for a continuous integrator the max efficiency occurs for $n\gamma = 1.257$.

## RADAR cross section of targets:-

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.

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

where R = distance b/w radar and target.

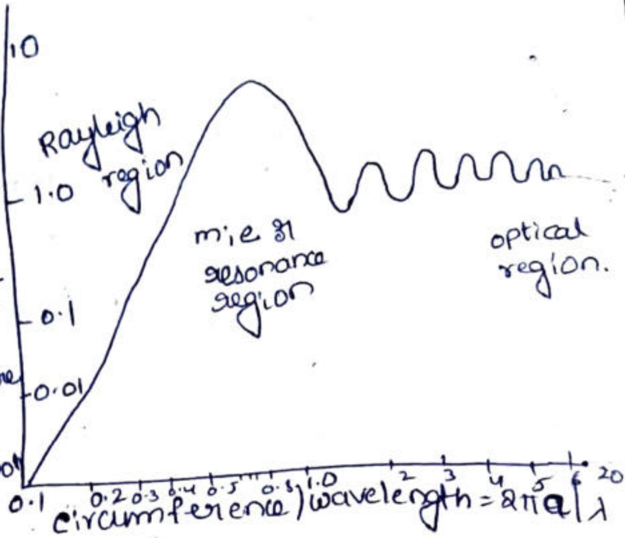
$E_r$ = reflected field strength at radar.

$E_i$ = strength of incident field at target.

Generally the real crosssection of target can be defined in terms of physical processing parameters such as scattering and diffraction. the scattered field is defined as the difference b/w the scattering area due to object to the area when the object is absent. the total scattering field due to the object is known as difraction area.

→ Basically the radar cross section can be analyzed based on the shapes of the target. the radar cross section of simple sphere is as shown in figure. the radar cross section of simple sphere contains three regions. ① Rayleigh region ② mie & resonance region ③ optical region.

Rayleigh region: In this region the size of sphere is less than wavelength i.e., $2\pi a/\lambda \Rightarrow a^2 << \lambda$ where a is the radius of the sphere and $\lambda$ is the wavelength.




→ Since the cross section of objects within the Rayleigh region varies as $\lambda^{-4}$ rain & clouds are essentially invisible to radar which operates at relatively long wavelengths (low frequencies). the usual radar targets are much larger than rain drops.

Mie region:- the cross section is oscillatory with frequency within this region. In this region the scattered area should be fluctuates b/w the two levels. the area of the sphere is equal to wavelength ($\lambda$).

optical region:- In this region dimensions of the sphere are large compared with the wavelength ($2\pi a/\lambda >> 1$). for large $2\pi a/\lambda$, the radar cross section approaches the optical cross section $\pi a^2$.

Transmitter power:-

The power $P_t$ in the radar equation is called ~~by~~ by the radar engineer the peak power. the peak pulse power as used in the radar equation is not the instantaneous peak power of a sine wave. It is defined as the power averaged over that carrier frequency cycle which occurs at the max of the pulse power.

→ peak power is usually equal to one-half the max instantaneous power. the avg radar power $P_{av}$ is also of interest in radar and is defined as the avg transmitter power over the pulse-repetition period.

→ If the transmitted ~~pow~~ waveform is a train of rectangular pulses of width $\tau$ and pulse repetition period $T_p = 1/f_p$. the avg power is related to the peak power by $P_{av} = \frac{P_t \tau}{T_p} = P_t \tau f_p$.

→ the ratio $P_{av}/P_t$, $\tau/T_p$, & $\tau f_p$ 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 interms of the avg power rather than the Peak power we get

$$R^4_{max} = \frac{P_{av} G A_e \sigma n E_i(n)}{(4\pi)^2 K T_0 F_n (B_n \tau)(S/N)_1 f_p}.$$

→ The B.W and the pulse width are grouped together since the product of the two usually of the order of unity in most pulse-radar applications. If the transmitted waveform is not a rectangular pulse, it is sometimes more convenient to express the radar equation in terms of energy $E_\tau = P_{av}/f_p$ contained in the transmitted waveform.

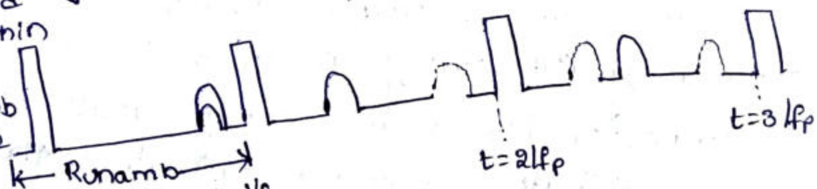
$$R^4_{max} = \frac{E_\tau G A_e \sigma n E_i(n)}{(4\pi)^2 K T_0 F_n (B_n\tau)(S/N)_1}$$

→ In this form the range does not depends explicitly on either the wavelength or the pulse repetition freq. the important parameters affecting range are the total transmitted energy $nE_\tau$, the transmitting gain $G$, the effective receiving aperture $A_e$, and the receiver noise figure $F_n$.

## Pulse Repetition frequency and range ambiguities :-

→ the pulse repetition frequency (prf) is determined primarily by the max range at which targets are ~~expressed~~ 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. target A is located within the max unambiguous range Runamb of the radar, target B is at a distance greater than Runamb but less than 2Runamb, while target C is greater than 2Runamb but less than 3Runamb.




(a)


(b) appearance of the three targets on the A-scope.


appearance of the three targets on the A-scope with a changing prf.

→ the multiple time around echoes on the A-scope cannot be distinguished from proper target echoes actually within the max unambiguous range. only the range measured for targets A is correct. those for B and C or not.

→ one method of distinguishing multiple time around echoes from unambiguous echoes is to operate with a varying pulse repetition freq. the echo signal from an unambiguous range target will appear at the same place on the A-scope on each sweep no matter whether the prf is modulated or not.

→ Instead of modulating the prf, other scheme that might be employed to identify multiple time around echoes include changing the pulse amplitude, pulse width, frequency, phase or polarization of transmission from pulse to pulse.

one of the fundamental limitations is the foldover of nearby targets ie., nearby strong ground targets can be quite large and can mask weak multiple time around targets appearing at the same place on the display. Also, more time is required to process the data when resolving ambiguities.

System losses :- The losses reduce the signal-to-noise ratio at the receiver o/p. they may be of two kinds, depending upon whether or not they can be predicted with any degree of precision. the antenna beam shape loss, collapsing loss, and losses in the microwave plumbing are examples of losses which can be calculated if the system configuration is known.

-> losses not readily subject to calculation and which are less predictable include those due to field degradation and to operator fatigue or lack of operator motivation.

① plumbing loss :- There is always some finite loss experienced in the transmission lines which connect the o/p of the transmitter to the Antenna. at the lower radar frequencies the transmission line introduces little loss, unless its length is exceptionally long. At the higher radar frequencies attenuation may not always be small and may have to be taken into account.

-> Connector losses are usually small, but if the connection is poorly made, it can contribute significant attenuation. the signal suffers attenuation as it passes through the duplexer.

② Beam-shape loss :- In reality the train of pulses returned from a target with a scanning radar is modulated in amplitude by the shape of the antenna beam. To properly take into account the pulse-train modulation caused by the beam shape, the computation of the probability of detection would have to be performed assuming a modulated train of pulses rather than constant-amplitude pulses.

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

③ limiting loss :- Limiting in the radar receiver can lower the probability of detection. Although a well designed and engineered receiver will not limit the received signal under normal circumstances, intensity modulated CRT displays such as the PPI or the B-scope have limiting dynamic range and may limit. however, by appropriately shaping the spectrum of the i/p noise it has been suggested that the degradation can be made negligibly small.

④ **Collapsing loss**:- If the radar were to integrate additional noise samples along with the wanted signal-to noise pulses, the added noise results in a degradation called the collapsing loss. It can occur in displays which collapse the range information, such as the C-scope which displays elevation ~~and~~ vs azimuth angle.

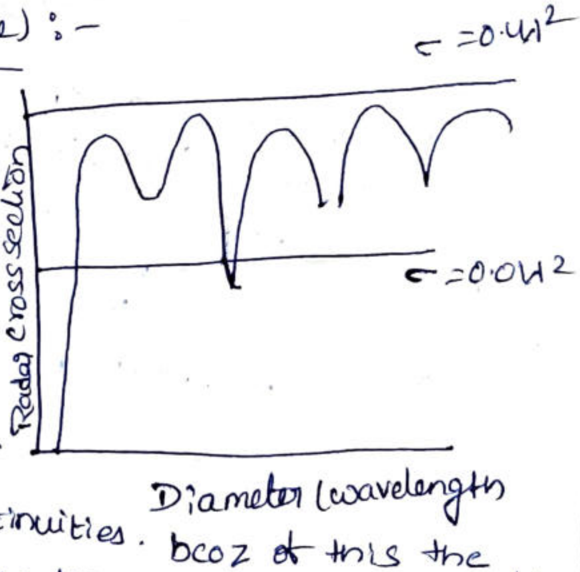
⑤ **Non ideal equipment**:- The transmitting tubes are not all uniform in quality, nor should it be expected that any individual tube will remain at the same level of performance throughout its useful life. also the power is usually not uniform over the operating band of the device. variations in the receiver noise figure over the operating band also are to be expected. If the receiver is not the exact matched filter for the transmitted waveform, a loss in signal-to-noise ratio will occur.

⑥ **operator loss**:- An alert, motivated and well-trained operator should perform as well as described by theory. however, when distracted, tired, overloaded or not properly trained operator performance will decrease.

⑦ **Field degradation**:- Factors which contributes to field degradation are poor tuning, weak tubes, water in the transmission lines, incorrect mixer-crystal current, deterioration of receiver noise figure, poor TR tube recovery, loose cable connections, tube. to minimize field degradation, radars should be designed with built in automatic performance monitoring equipment.

⑧ **other loss factors**:- A radar designed to discriminate b/w moving targets and stationary objects (MTI radar) may introduce additional loss over a radar without this facility. the MTI discrimination technique results in complete loss of sensitivity for certain values of target velocity relative to the radar. these are called blind speeds.

**Radar cross section of targets (cone-sphere)**:-




→ An interesting radar scattering object is the cone-sphere, a cone whose base is capped with a sphere such that the first derivatives of the cone and sphere contours are equal at the join b/w the two.

→ the cross section of the cone sphere from the vicinity of the nose-on direction is quite low. Scattering from any object occurs from discontinuities. bcoz of this the back scattering of the cone sphere are from the tip and from the join b/w the cone and the sphere.

problems:-

① compute the max detectable range of a radar system specified below. operating wavelength = 3.2cm, peak pulse transmitted power = 500 kW min detectable power = $10^{-13}$ w, capture area of the antenna = 5 sq.m radar cross sectional area of the target = 20 sq.cm.

sol. $R^4_{max} = \dfrac{P_t G \sigma A_e}{(4\pi)^2 S_{min}}$ where $G = \dfrac{4\pi A_e}{\lambda^2}$.

$= \dfrac{500\times10^3 \times 4\pi \times 5^2 \times 20 \times 10^{-4}}{(4\pi)^2 \times 10^{-13}} = \dfrac{5\times10^5 \times 4\pi \times 25 \times 2\times10^{-3}}{4\pi^2 \times 10^{-13} (3.2)^2}$.

$= \dfrac{50\times5\times10^2}{4\pi\times10^{-13}} = \dfrac{250}{4\pi}\times10^{15} = \dfrac{62.5}{\pi}\times10^{15} = \dfrac{62.5}{3.14}\times10^{15}$

$R_{max} = 663.90$ km

② for the specifications of a Radar listed below compute the power received at 50 km distance from the radar antenna. the specifications are operating wavelength = 3cm, pulse transmitted power = 320 kW, radar cross section = 12 sq.m Apeture area of the antenna is 5 sq m; radar cross section = 12 sq.m and transmitting gain = $9.6\times10^4$.

sol $S_{min} = \dfrac{P_t G \sigma A_e}{(4\pi)^2 R^4_{max}} = \dfrac{320\times10^3 \times 9.6\times10^4 \times 12\times5}{(4\pi)^2 \times (50\times10^3)^4}$.

$= \dfrac{32\times9.6\times60\times10^7}{16\times\pi^2\times5^4\times10^{16}} = \dfrac{32^2\times9.6\times6}{16\times\pi^2\times5^4\times10^8} = \dfrac{1152\times10^{-9}}{625\times\pi^2}$

$\therefore S_{min} = 1.86\times10^{-9}$ w.

③ If the noise figure of a receiver is 2.5 db. what reduction (measure in db) occurs in the signal noise ratio at the o/p compare to the signal noise ratio at the i/p.

sol. given F = 2.5 db

we know $F = \dfrac{(S/N)_i}{(S/N)_o} = 2.5$ db

(Q) what is the max radar cross section of an automobile license plate i.e., 12 inches wide by 6 inches height at a frear of 10.525GHz.

Sol the radar cross section of the target '$\sigma$' is taken as cone structure

$$\therefore \sigma = \frac{4\pi A_e^2}{\lambda^2} = \frac{4\pi \times (W \times H)^2}{\lambda^2} \qquad \lambda = c/f$$

$$= \frac{3\times10^8}{10.523\times10^9}$$

$$= \frac{4\pi(12\times10^{-2}\times 2.54\times6\times 2.54\times10^{-2})^2}{\left(\frac{3\times10^8}{10.525\times10^9}\right)^2}$$

1 inch = 2.5 cm $\Rightarrow \frac{3}{10.525\times10} = \frac{3}{105.25} \Rightarrow \lambda = 0.029m.$

$$= \frac{4\pi(72\times 2.54\times 2.54\times10^{-4})^2}{(0.029)^2}$$

$$= \frac{4\pi\times(72)^2\times(2.54)^2\times10^{-8}}{(0.029)^2} \Rightarrow \boxed{\sigma = 32.24m^2}$$

# UNIT - 2

## CW and Frequency Modulated Radar. 635

### Doppler Effect:-

The radar detects the presence of objects and locates their position in space by transmitting electromagnetic energy and observing ~~echo~~ 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 b/w the transmission of the pulse and the receipt of the echo is a measure of the distance to the target. seperation of the echo signal and the transmitted signal is made on the basis of differences in time.

The radar transmitter may be operated continuously rather than pulsed if the strong transmitted signal can be seperated from the weak echo. A feasible technique for seperating the received signal from the transmitted signal when there is relative motion b/w radar and target is based on recognizing the change in the echo signal frequency caused by the doppler effect.

It is well known in the fields of optics that if either the source of oscillation or the observer of the oscillation is in motion, an apparent shift in frequency will result. this is doppler effect and is the basis of CW radar. If R is the distance from the radar to target, the total number of wavelengths $\lambda$ contained in the two-way path b/w the radar and the target is $2R/\lambda$. since one wavelength corresponds to an angular excursion of $2\pi$ radians, the total angular excursion $\phi$ made by the electromagnetic wave during its transit to and from target is $4\pi R/\lambda$ radians.

→ If the target is in motion, R and the phase $\phi$ are continually changing. A change in $\phi$ with respect to time is equal to a frequency. this is the doppler angular frequency $\omega_d$ given by

$$\omega_d = 2\pi f_d = \frac{d\phi}{dt} = \frac{4\pi}{\lambda}\frac{dR}{dt} = \frac{4\pi V_r}{\lambda}$$

where $f_d$ = doppler frequency shift $V_r$ = relative velocity of target with respect to radar. the doppler frequency shift is

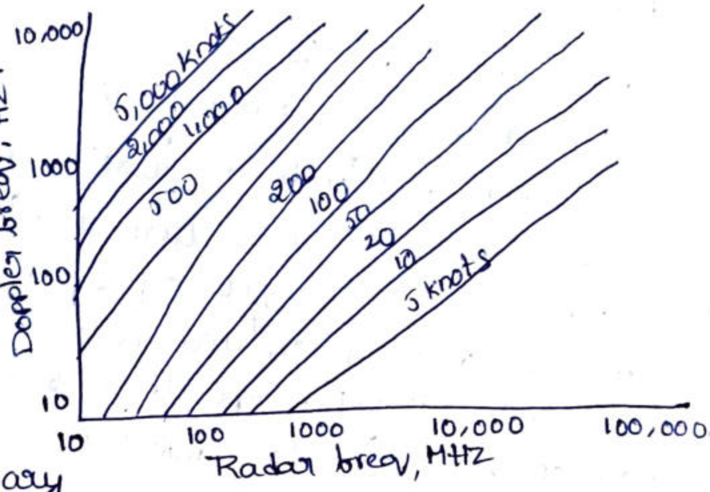
$$f_d = \frac{2V_r}{\lambda} = \frac{2V_r f_0}{c}$$

where $f_0$ = transmitted freq
$c$ = velocity of propagation = $3\times10^8$ m/s. if $f_d$ is in Hertz, $V_r$ in knots & $\lambda$ in meters

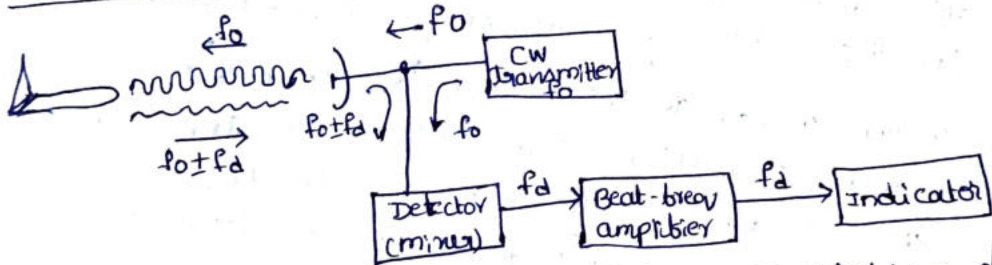
$$f_d = \frac{1.03V_r}{\lambda}$$

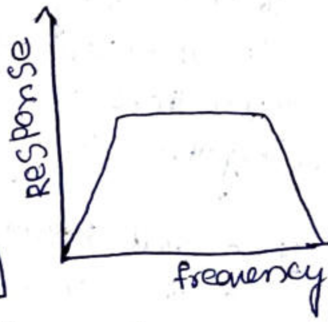
. the relative velocity may be written $V_r = V\cos\theta$, where V is the target speed and $\theta$ is the angle made by the target trajectory and the line joining radar and target. when $\theta = 0$, the doppler freq is max. the doppler is zero when the trajectory is perpendicular to the radar line of sight ($\theta = 90°$). the type of radar which employs a continuous transmission either modulated or unmodulated has had wide application.

The CW radar is of interest not only because of its many applications, but its study also serves as a means for better understanding the nature & use of the doppler information contained in the echo signal, whether in a CW or a pulse radar (MTI) application. In addition provides a measurement of relative velocity which may be used to distinguish moving targets from stationary objects or clutter.




## CW Radar:- Block diagram:-




The fig shows a simple CW Radar consisting of CW transmitter, detector, beat frequency amplifier and Indicator.

**CW transmitter:-** the transmitter generates a continuous (unmodulated) oscillation of freq fo, which is radiated by the antenna. A portion of the radiated energy is intercepted by the targets and is scattered, some of it in the direction of the radar, where it is collected by the receiving antenna. If the target is in motion with a velocity $V_r$ relative to the radar, the received signal will be in freq from the transmitted freq fo by an amount $\pm f_d$.

→ the shift in frequency is $\pm f_d$. the '+' sign indicates the distance b/w target and Radar is less i.e., the target is moving near to radar. the '-' sign indicates the distance b/w target & radar is high i.e., the target is moving away from the radar. If the target is moving towards the radar then the change in freq or shift freq is greater than fo.

Detector :- The received echo signal at a frequency $f_0 \pm f_d$ enters the radar via the antenna and is heterodyned in the detector (mixer) with a portion of the transmitter signal $f_0$ to produce a doppler beat $f_d$. the sign of $f_d$ is lost in this process.

Beat frequency amplifier :- Beat freq amplifier is generally a doppler amplifier. the purpose of the doppler amplifier is to eliminate echoes from stationary targets and to amplify the doppler echo signal to level where it can operate an Indicating device. the low freq cutoff must be high enough to reject the d-c component caused by stationary targets. the upper cutoff freq is selected to pass the highest doppler freq expected.

Indicator :- the Indicator might be a pair of earphones or a frequency meter. If exact knowledge of the doppler freq is not necessary, earphones are especially attractive provided the doppler frequencies lies within the audio freq response of the ear. earphones are not only simple devices, but the ear acts as a selective bandpass filter with a passband of the order of 50Hz centered about the signal freq.

Isolation b/w transmitter and receiver :-

→ A single Antenna serves the purpose of transmission and reception in the simple CW radar. In principle, a single antenna may be employed since the necessary Isolation b/w the transmitted and received signal is achieved via separation in freq as a result of the doppler effect. In practice, it is not possible to eliminate completely the transmitter leakage.

→ A moderate amount of leakage entering the receiver along with the echo signal supplies the reference necessary for the detection of the doppler freq shift. there are two practical effects which limit the amount of transmitter leakage power which can be tolerated at the receiver.

① the max amount of power the receiver I/P ckty can withstand before it is physically damaged or its sensitivity reduced.

② the amount of transmitter noise due to hum, microphonics, stray pick-up, and Instability which enters the receiver from the transmitter.

→ the additional noise Introduced by the transmitter reduces the receiver sensitivity. the receiver of a pulsed radar is Isolated and protected from the damaging effects of the transmitted pulse by the duplexer, which shuts circuits the receiver I/P during the transmission period. Turning off the receiver during transmission with a duplexer is not possible in a CW radar since the transmitter is operated continuously.

Isolation b/w transmitter & receiver might also be used be obtained with a single antenna by using a hybrid junction, circulator, turnstile junction, or with seperate polarization. seperate antennas for transmitting and receiving might also be used.

→ the amount of Isolation which can be readily achieved b/w the arms of practical hybrid junctions such as the magic-T, rat race or short slot coupler is ~~of~~ the order of 20 to 30dB. one limitation of the hybrid junction is the 6-dB loss in overall performance which results from the inherent waste of half the transmitted power and half the received signal power. because of this performance degradation large Isolations have limited the application of the hybrid junction to short-range radars.

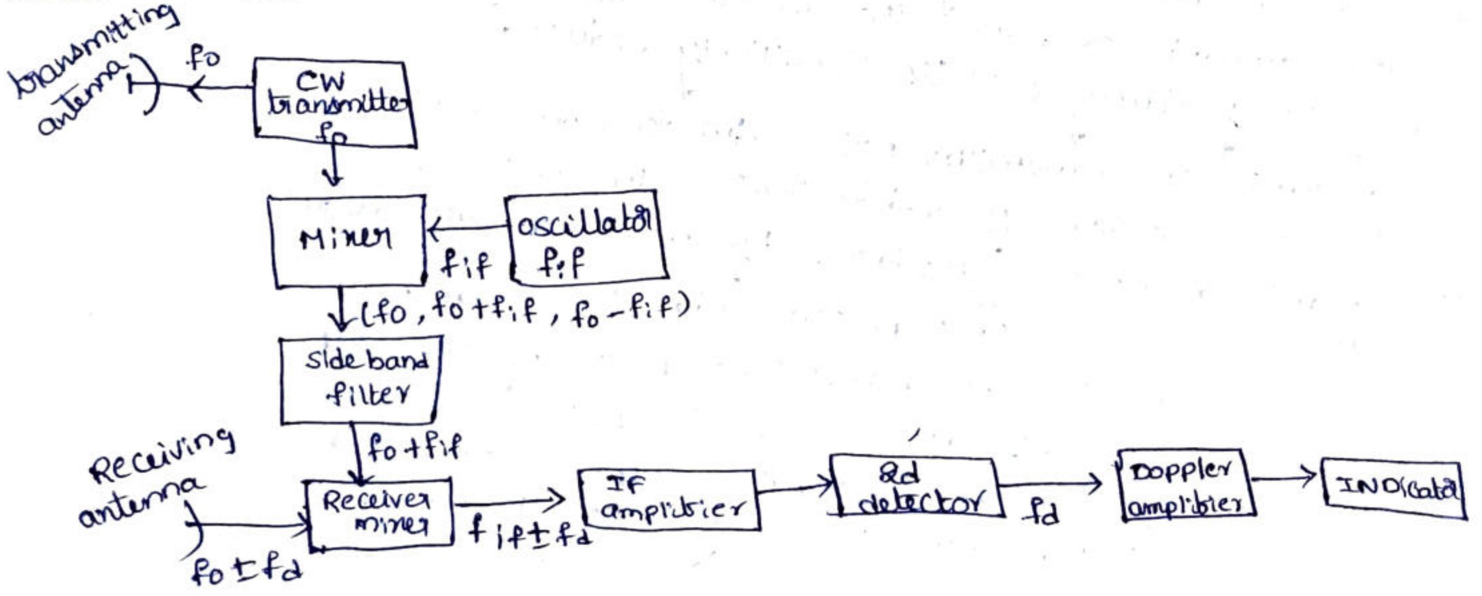
→ Ferrite Isolation devices such as the circulator dont suffer the 6-dB loss inherent in the hybrid junction. the use of orthogonal polarizations for transmitting and receiving is limited to short range radars bcoz of the relatively small amount of Isolation that can be obtained. turnstile junctions achieve Isolations as high as 40 to 60dB.

→ the reflection coefficient from a mismatched antenna with a voltage standing wave ratio $\sigma$ is $|\rho| = (\sigma-1)/(\sigma+1)$. therefore if an Isolation of 20dB is to be obtained, the VSWR must be less than 1.22. If 40dB of Isolation is required, the VSWR must be less than 1.02.

→ The largest Isolations are obtained with two antennas - one for transmission, the other reception - physically seperated from one another. Isolation of the order of 80dB or more are possible with high gain antennas. the more directive the antenna beam and the greater the spacing b/w the antennas the greater will be the Isolation.

## Intermediate - frequency receiver (or) side band superheterodyne :-

→ Block diagram of CW doppler radar with non zero IF receiver is shown below.



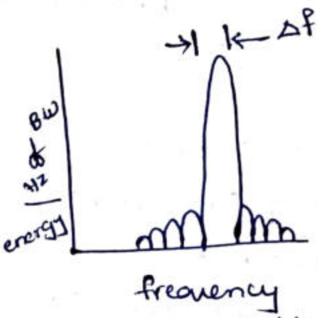
→ The receiver of the simple CW radar is in some respects analogous to a superheterodyne receiver. Receivers of this type are called homodyne receivers, or superheterodyne receivers with zero IF. the function of the local oscillator is replaced by the leakage signal from the transmitter. such a receiver is simpler than one with a more conventional Intermediate frequency since no IF amplifier or local oscillator is required.
→ The simpler receiver is not as sensitive bcoz of Increased noise at the lower Intermediate frequencies caused by flicker effect. flicker effect noise occurs in semiconductors devices such as diode detectors and cathodes of vaccum tubes. the noise power produced by ~~devices such as~~ the flicker effect varies as $1/f^{\alpha}$, where $\alpha$ is approximately unity. this is in contrast to short noise or thermal noise, which is Independent of frequency.
→ Thus at the lower range of frequencies where the doppler frequencies usually are found, the detector of the CW receiver can introduce a considerable amount of flicker noise, resulting in reduced receiver sensitivity. But for man efficiency with CW radar, the reduction in sensitivity caused by the simple doppler receiver with zero IF cannot be tolerated.
→ The effects of flicker noise are overcome in the normal superheterodyne receiver by using an Intermediate freq high enough to render the flicker noise small compared with the normal receiver noise. this results from the Inverse freq dependence of flicker noise. figure shows the block diagram of the CW radar, ~~the~~ whose receiver operates with a non zero IF. seperate antennas are shown for transmission and reception Instead of the usual local oscillator found in the conventional superheterodyne receiver, the local oscillator is dervied in this receiver from a portion of the transmitted signal mixed with a locally generated signal of freq equal to that of the receiver IF.
→ since the o/p of the mixer consists of two sidebands on either side of the carrier plus higher harmonics, a narrow band filter selects one of the sidebands as the reference signal. the Improvement in receiver sensitivity with an Intermediate freq superhetrodyne might be as much as 30dB over the simple receiver.

Receiver Bandwidth:- one of the requirements of the doppler-frequency amplifier in the simple CW radar or the IF amplifier of the sideband super heterodyne is that it be wide enough to pass the expected range of doppler frequencies.
→ In most cases of practical Interest the expected range of doppler frequencies will be much wider than the freq spectrum occupied by the signal energy.

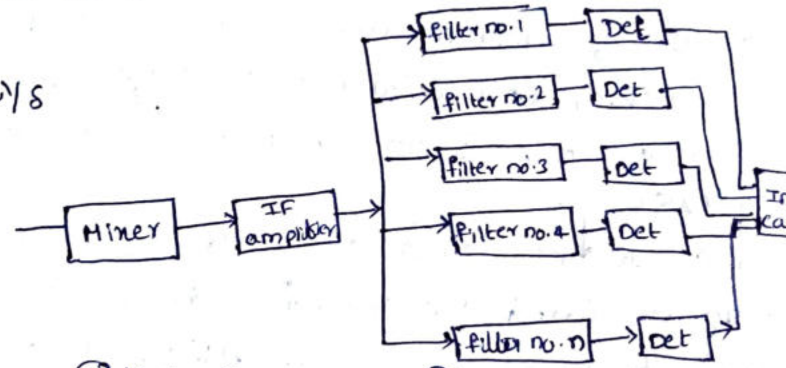
→ consequently, the use of a wideband amplifier covering the expected doppler range will result in an increase in noise and a lowering of the receiver sensitivity.

→ several factors tend to spread the CW signal energy over a finite freq band. these must be known if an approximation to the bandwidth required for the narrowband doppler filter is to be obtained. the freq spectrum of a finite duration sine wave has a shape of the form $\left[\frac{\sin \pi (f-f_0)\delta}{\pi (f-f_0)}\right]$. where $f_0$ and $\delta$ are the freq and duration of the sine wave respectively, and $f$ is the freq variable over which the spectrum is plotted.

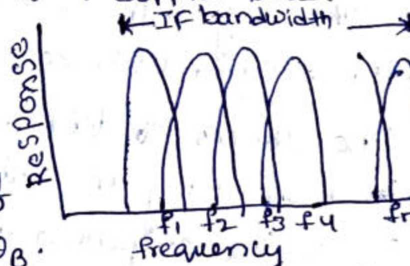

infinite duration




finite duration




(a) block diagram of IF doppler filter bank.




(b) freq response characteristics of doppler.

→ Assume a CW radar with an antenna beam width of $\theta_B$ deg scanning at the rate of $\theta_s$ deg/s. the time on target is $\delta = \theta_B/\theta_s$ s. thus the signal is of finite duration and the bandwidth of the receiver must be of the order of the reciprocal of the time on target $\theta_s/\theta_B$.

→ If the target's relative velocity is not constant, a further widening of the received signal spectrum can occur. If $a_r$ is the acceleration of the target with respect to the radar, the signal will occupy a bandwidth $\Delta f_d = \left(\frac{2a_r}{\lambda}\right)^{1/2}$.

→ when the doppler shifted echo signal is known to lie somewhere within a relatively wide band of frequencies, a bank of narrowband filters shown in a fig spaced throughout the freq range permits a measurement of freq and improves the signal-to-noise ratio. the bandwidth of each individual filter is wide enough to accept the signal energy, but not so wide as to introduce more noise than need be. the center frequencies of the filter are staggered to cover entire range of doppler frequencies. one advantage of the fold over in the video is that only half the number of filters are required than in the IF filter bank.

→ If, in any of the above techniques, moving targets are to be distinguished from stationary objects, the zero doppler freq component must be moved. the zero doppler freq component has, in practice, a finite b.w due to the finite time on target, clutter fluctuations and equipment instabilities.

## 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 rate - of climb meter for vertical - take - off aircraft.
→ In support of automobile traffic, CW radar has been suggested for the
control of traffic lights, regulation of toll booths, vehicle counting, as a replace-
ment for the "fifth wheel" speedometer in vehicle testing, as a sensor in
antilock braking system and for collision avoidance.
→ 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 is that of
large ships. It had also seen application for intruder alarms & for the measure-
ment of the velocity of missiles, ammunition and base balls.
→ the principle advantage of a CW doppler radar over other methods of measuring
speeds is that there need not be any physical contact with the object whose
speed is being measured.
→ Most. of the above applications can be satisfied with a simple, solid - state
CW source with powers in the tens of milliwatts. High power CW radars for
the detection of aircraft and other targets have been developed and have been
used in such systems at. the Hawk missile systems. the CW radar, when
used for short or moderate ranges, is characterized by simple equipment
than a pulse radar.

## Frequency - Modulated CW radar :-

→ perhaps one of the greatest shortcomings of the simple CW radar is its
inability to obtain a measurement of range. the inability of the simple
CW radar is related to be relatively narrow spectrum (bandwidth) of its
transmitted waveform. Some sort of timing mark must be applied to a
CW carrier if range is to be measured. the timing mark permits the time
of transmission and the time of return to be recognized. the sharper or
more distinct the mark, the more accurate the measurement of the transit time.
→ But the more distinct the timing mark, the broader will be the transmitted
spectrum. this follows from the properties of the fourier transform. therefore
a finite spectrum must be necessity be transmitted if transit time or
range is to be measured.

→ the spectrum of a CW transmission can be broadened by the application of modulation either amplitude, frequency or phase. The narrower the pulse, the more accurate the measurement of range and the broader the transmitted spectrum.

→ A widely used technique to broaden the spectrum of CW radar is to frequency modulate the carrier. the timing mark is the changing frequency. the transit time is proportional to the difference in freq b/w the echo signal and the transmitter signal. The greater the transmitter freq deviation in a given time interval, the more accurate the measurement of the transit time and the greater will be the transmitted spectrum.

## Range & Doppler Measurement:-

→ In the freq modulated CW radar, the transmitter freq is changed as a function of time in a known manner. Assume that the transmitter freq increases linearly with time, as shown by the solid line in the fig ⓐ.

→ 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 fig represents the echo signal.

→ If the echo signal is heterodyned with a portion of the transmitter signal in a nonlinear element such as diode, a beat note $f_b$ will be produced. If there is no doppler freq shift, the beat note $f_b = f_r$, is a measure of the target's range and where $f_r$ is the beat freq due only to the target's range.

→ If the rate of change of the carrier freq is $\dot{f_0}$, the beat freq is $f_r = \dot{f_0} T = \frac{2R}{c}\dot{f_0}$.

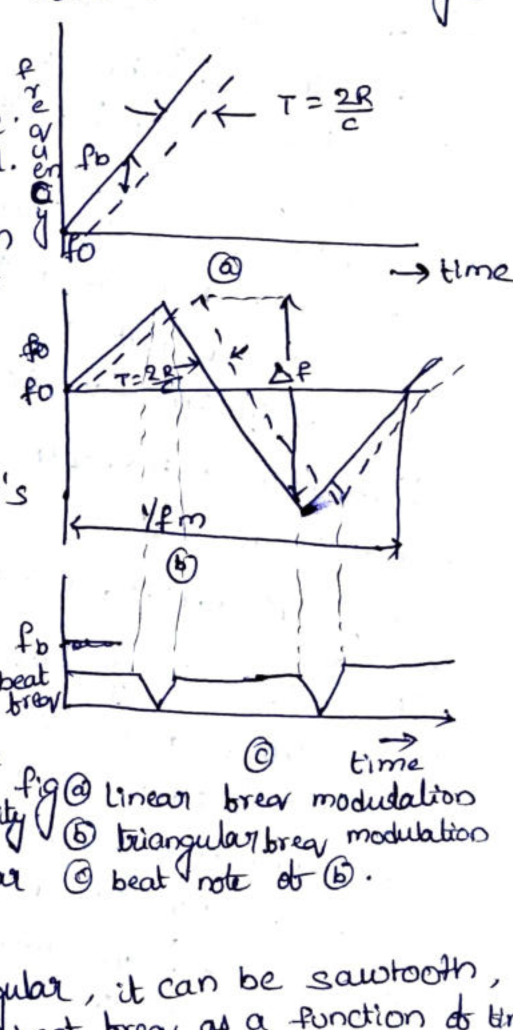



fig ⓐ Linear freq modulation
ⓑ triangular freq modulation
ⓒ beat note of ⓑ.

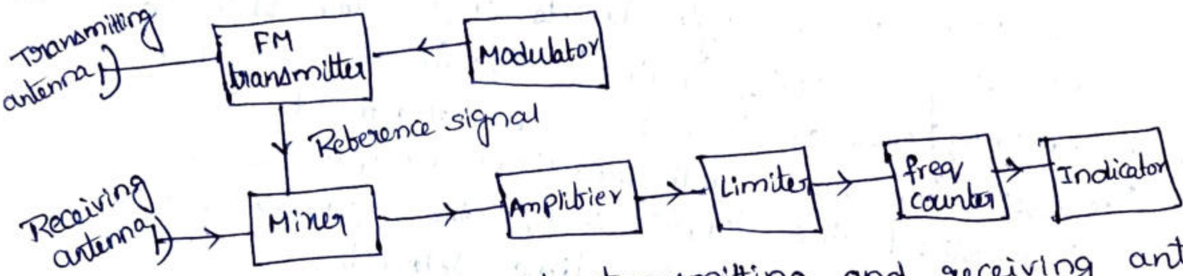
→ In any practical CW radar, the freq cannot be continually changed in one direction only. periodicity in the modulation is necessary, as in the triangular freq modulation waveform as shown in fig ⓑ.

→ The modulation need not necessarily be triangular, it can be sawtooth, sinusoidal, or some other shape. the resulting beat freq as a function of time is shown in fig ⓒ for triangular modulation. the beat note is of constant freq except at the turn around region. If the freq is modulated at a rate $f_m$ over a range $\Delta f$, the beat freq is $f_r = \frac{2R}{c} 2 f_m \Delta f \Rightarrow \boxed{f_r = \frac{4 R f_m \Delta f}{c}}$

thus the measurement of the beat freq determines the range R.

## Block diagram of FM-CW:-

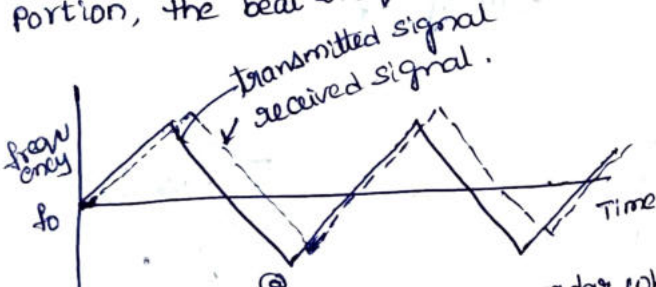
→ A Block diagram Illustrating the principle of the FM-CW radar is shown in fig. A portion of the transmitter signal acts as the reference signal required to produce the beat freq. It is Introduced directly into the receiver via a cable or other direct connection.



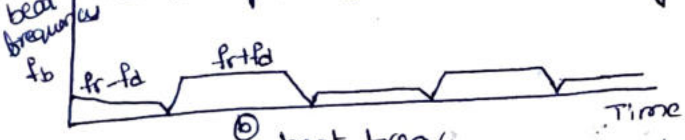

→ Ideally, the Isolation b/w transmitting and receiving antennas is made sufficiently large so as to reduce to a negligible level the transmitter leakage signal which arrives at the receiver via the coupling b/w antennas. The beat freq is amplified and limited to remove any amplitude fluctuations. the freq of the amplitude limited beat note is measured with a cycle counting freq meter calibrated in distance.

→ In the above, the target was assumed to be stationary. If this assumption is not applicable, a doppler freq shift will be superimposed on the FM range beat. the target is approaching the radar, the beat freq $f_b(UP)$ produced during the Increasing, or up portion of the FM cycle will be the difference b/w the beat freq due to the range $f_r$ and the doppler freq shift $f_d$. similarly, on the decr portion, the beat freq $f_b(down)$ is the sum of the two.

$$f_b(UP) = f_r - f_d$$
$$f_b(down) = f_r + f_d.$$




ⓐ
Freq/time relationships in FM-CW radar when the received signal is shifted in freq by the doppler effect




ⓑ beat freq

→ the range freq $f_r$ may be extracted by measuring the average beat freq. i.e., $\frac{1}{2}[f_b(UP) + f_b(down)] = f_r$. when more than one target is present within the view of radar; the mixer o/p will contain more than one difference freq. If the system is linear, there will be a freq component corresponding to each target.

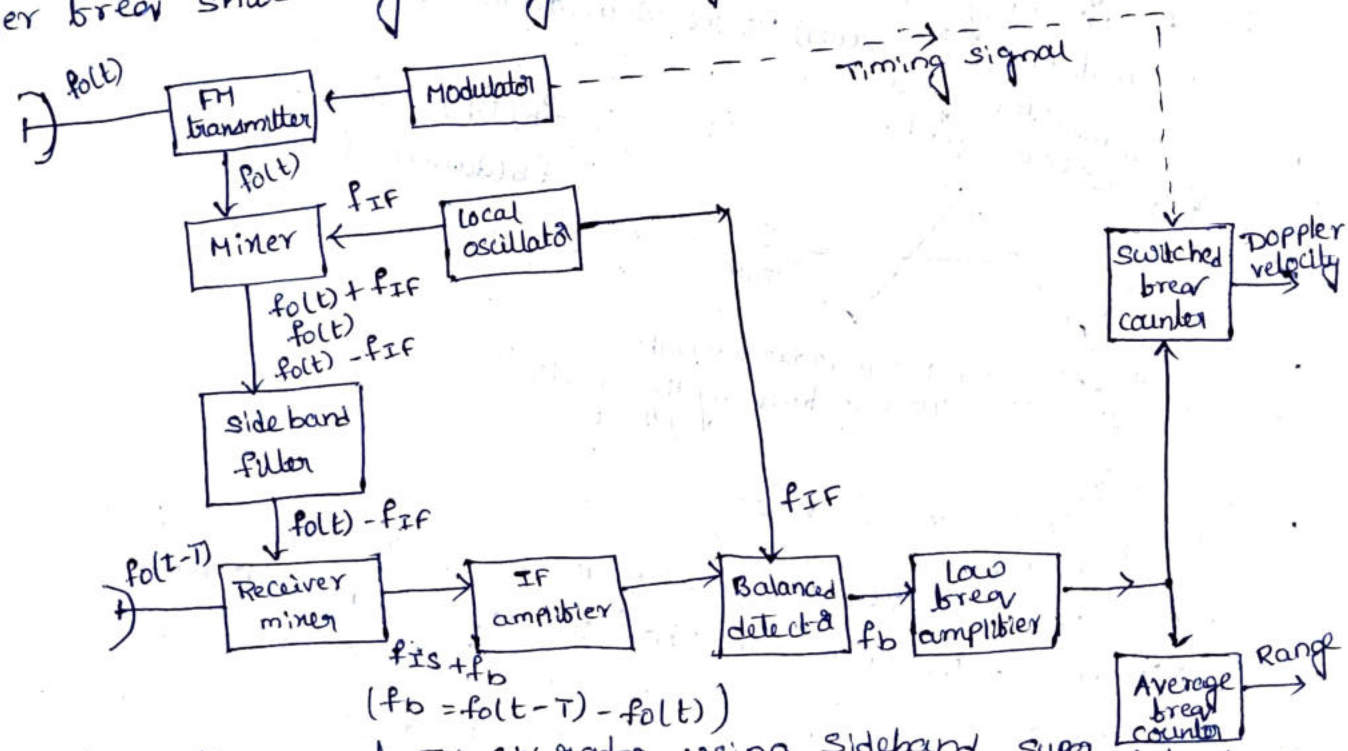
→ In principle, the range to each target may be determined by measuring the individual freq components. To measure the individual frequencies, they must be separated from one another. this might be accomplished with a bank of narrowband filters, or alternatively, a single freq corresponding to a single target may be singled out and continuously observed with a narrow band tunable filter.

→ If the FM-CW radar is used for single targets only, such as in the radio altimeter, it is not necessary to employ a linear modulation waveform. this is certainly advantageous since a sinusoidal or almost sinusoidal freq modulation is easier to obtain with practical equipments than are linear modulations. To extract the doppler freq, the modulation waveform must have equal upsweep and downsweep time intervals.

→ the FMCW radar principle was known and used at about the same time as pulse radar, although the early development of these two radar techniques seemed to be relatively independent of each other.

## 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 back-scatter cross section and the relatively short ranges required of altimeters permit low transmitter power & low antenna gain. since the relative motion b/w the aircraft and ground is small, the effect of the doppler freq shift may usually be neglected.




Block diagram of FM-CW radar using sideband super-heterodyne receiver.

→ The band from 4.2 to 4.4 GHz 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 portion of the freq modulated transmitted signal is applied to a mixer along with the oscillator signal. the selection of the local-oscillator freq is a bit different from that in the usual superheterodyne receiver.

→ the local-oscillator freq $f_{IF}$ should be the same as the Intermediate freq used in the receiver, whereas in the conventional superhetrodyne the LO freq is of the same order of magnitude as the RF signal. the o/p of the mixer consists of the varying transmitter freq $f_o(t)$ plus two side bands freq, one on either side of $f_o(t)$ and separated from $f_o(t)$ by the local oscillator freq $f_{IF}$. the filter selects the lower sideband $f_o(t) - f_{IF}$ and rejects the carrier and the upper sideband.

→ the sideband filter must 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 o/p of the receiver mixer is an IF signal of freq $f_{IF} + f_b$, where $f_b$ is composed of the range freq. $f_r$ and the doppler velocity freq $f_d$.

→ The IF signal is amplified and applied to the balanced detector along with the local-oscillator signal $f_{IF}$. the o/p of the detector contains the beat freq which is amplified to a level where it can actuate the frequency measuring ckts.

→ the o/p of the low-freq amplifier is divided into two channels: one feeds an avg-freq counter to determine range, the other feeds a switched freq counter to determine the doppler velocity ($f_r > f_d$). only the averaging freq counter need to be used in an altimeter application, since the rate of change of altitude is usually small.

→ A target at short range will generally result in a strong signal at low freq, while one at long range will result in a weak signal at high freq. therefore the freq characteristics of the low frequencies amplifier in the FM-CW radar may be shaped to provide attenuation at the low frequencies corresponding to short ranges and large echo signals. less attenuation is applied to the higher frequencies where the echo signals are weaker.

→ Another method of processing the range & height information from an altimeter so as to reduce the noise o/p from the receiver and improve the sensitivity uses a narrow-bandwidth low-freq amplifier with a feedback loop to maintain the beat freq constant.

Measurement errors :- The absolute accuracy of radar altimeter is usually of more important at low altitude than at high altitude. 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.

→ A common form of frequency-measuring device is the cycle counter, which measures the number of cycles & 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 number of the freq measurement gives rise to an error called the fixed error, or step error. It has also been called the quantization error. the avg no of cycles N of the beat freq $f_b$ in one period of the modulation cycle $f_m$ is $\bar{f_b}/f_m$. where the bar over $f_b$ denotes time average. $\therefore R = \frac{cN}{4\Delta f}$ where R = range (altitude), m

c = velocity of propagation, m/s.

$\Delta f$ = freq excursion, Hz.

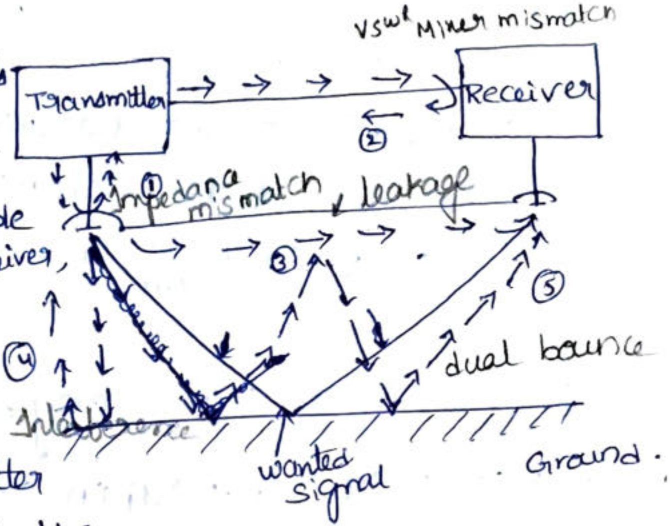
since the o/p of the freq counter N is an integer, the range will be an integral multiple of $c/(4\Delta f)$ and will give rise to a quantization error equal to $\delta R = \frac{c}{4\Delta f}$ (or) $\delta R(m) = \frac{75}{\Delta f(MHz)}$.

→ The fixed error is independent of the range and carrier freq and is a function of the freq excursion only. large frequency excursions are necessary if the fixed error is to be small.

→ other errors might be introduced in the CW radar if there are uncontrolled variations in the transmitter freq, modulation freq, & freq excursion. target motion can cause an error in range equal to $V_r T_0$, where $V_r$ is the relative velocity and $T_0$ 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. fig shows some of the unwanted signals that might occur in the FM altimeter. the wanted signals is shown by the solid line, while the 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 b/w 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.

→ Reflections from the landing gear can also cause errors.

Transmitter Leakage :-

→ The sensitivity of FM-CW radar is limited by the noise accompanying the transmitter signal which leaks into the receiver. Although advances have been made in reducing the AM and FM noise generated by high-power CW transmitters, the noise is usually of sufficient magnitude compared with the echo signal to require some means of minimizing the leakage that finds its way into the receiver.

→ the Techniques described previously for reducing leakage in the CW radar apply equally well to the FM-CW radar. Seperate antennas and direct cancellation of the leakage signal are two techniques which give considerable Isolation.

Sinusoidal Modulation :- 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 freq signal obtained by heterodyning the ..

transmitted and received signals. If the CW carrier is freq modulated by a sine wave, the difference freq obtained by heterodyning the returned signal with a portion of the transmitter signal.

→ Assume the form of the transmitted signal to be

$$\sin\left(2\pi f_0 t + \frac{\Delta f}{2 f_m} \sin 2\pi f_m t\right) \rightarrow ①$$

where $f_0$ = carrier freq
$f_m$ = modulation freq
$\Delta f$ = freq excursion (equal to twice the freq derivation).

The difference freq signal may be written

$$V_D = J_0(D)\cos(2\pi f_d t - \phi_0) + 2J_1(D)\sin(2\pi f_d t - \phi_0)\cos(2\pi f_m t - \phi_m) - 2J_2(D)\cos(2\pi f_d t - \phi_0)\cos(2\pi f_m t - \phi_m) - 2J_3(D)\sin(2\pi f_d t - \phi_0)\cos 3(2\pi f_m t - \phi_m) + 2J_4(D)\cos(2\pi f_d t - \phi_0)\cos 4(2\pi f_m t - \phi_m) + 2J_5(D) \cdots \rightarrow ②$$

where $J_0, J_1, J_2, J_3, J_4$ ... etc = Bessel function 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 $t = 0$.

$c$ = velocity of propagation.

$f_d = 2 v_r f_0 / c$ = doppler freq shift.

$v_r$ = relative velocity of target with respect to radar.

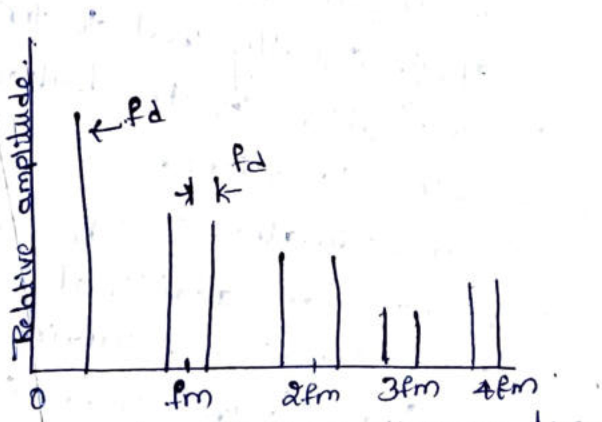
$\phi_0$ = phase shift approximately equal to angular distance $4\pi f_0 R_0 / c$.

$\phi_m$ = phase shift approximately equal to $2\pi f_m R_0 / c$.

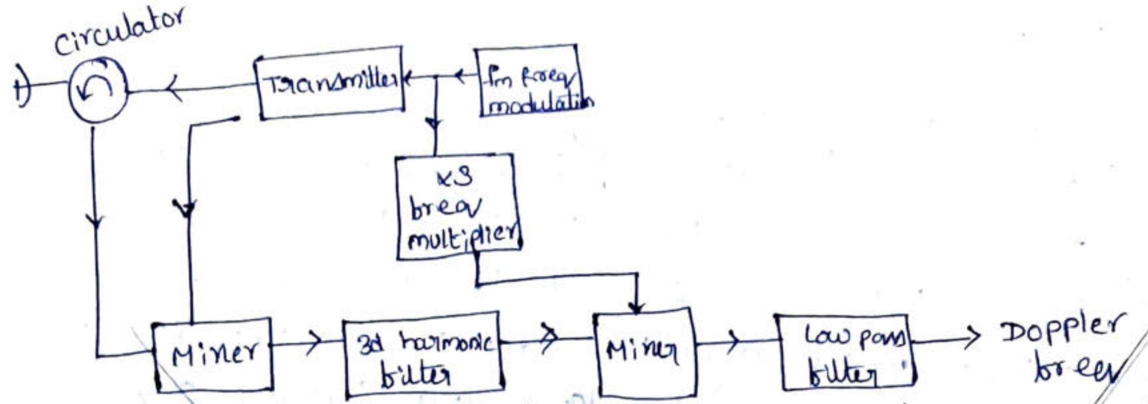
→ the difference freq signal consists of a doppler freq component of amplitude $J_0(D)$ and a series of cosine waves of freq $f_m, 2f_m, 3f_m$ etc, each of these harmonics of $f_m$ is modulated by the doppler freq component with amplitude proportional to $J_n(D)$.

→ In principle, any of the $J_n$ components of the difference freq signal can be extracted in the FM-CW radar. Consider the first the d-c term $J_0(D)\cos(2\pi f_d t - \phi_0)$. this is a cosine wave at the doppler freq with an amplitude proportional to $J_0(D)$.

→ when only a single target is involved, the freq excursion $\Delta f$ can be adjusted to obtain that value of $D$ which places the the max of the Bessel function at the target range.



spectrum of the difference freq signal.

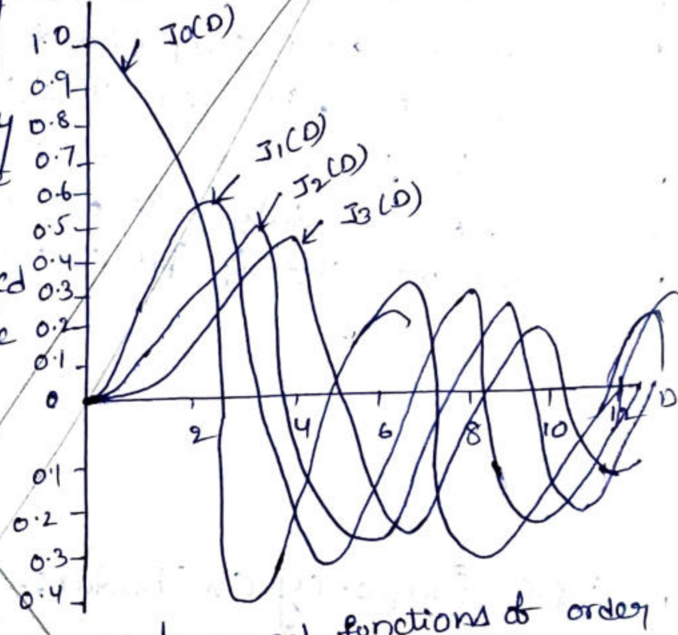



sinusoidally modulated FM - CW radar extracting the third harmonic.

→ A block diagram of a CW radar using the third harmonic ($J_3(D)$) is shown in the fig. the transmitter is sinusoidally freq modulated at a freq fm to generate waveform shown in the fig.

→ the doppler shifted echo is heterodyned with the transmitted signal to produce the beat freq signal of eq ②.

→ one of the harmonics of fm is selected by a filter centered at the harmonics. the filter bandwidth is wide enough to pass both doppler freq sidebands.




plot of Bessel functions of order 0, 1, 2 and 3 $D = \Delta f/f_m \sin 2\pi f_m R_0/c$

→ the filter o/p is mixed with the (third) harmonics of fm. the doppler freq is extracted by the low pass filter.

→ since the total energy contained in the beat freq signal is distributed among all the harmonics, extracting but one component wastes signal energy contained in the other harmonics and results in a loss of signal as compared with an ideal CW radar.

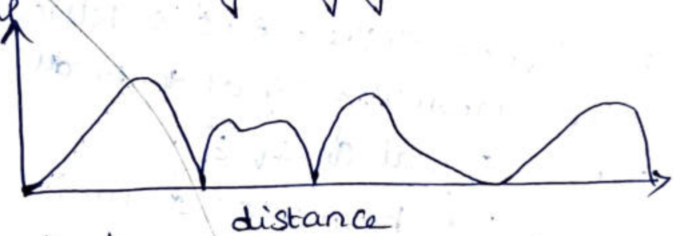
→ However, the signal-to-noise ratio is generally superior in the FM radar designed to operate with the nth harmonic as compared with a practical CW radar bcoz of the transmitter leakage noise is suppressed by the nth order bessel function.

→ Although two seperate transmitting and receiving antennas may be used, it is not necessary in many applications. A single antenna with a circulator and ~~by reflections~~ leakage introduced by the circulator and by reflections from the antenna are at close range and thus are attenuated by the $J_3$ factor.

**Matched filter detection:-** the operation of the FM-CW radar described does not employ optimum signal processing. the receiver is not designed as a matched filter for the particular transmitted waveform. therefore, the sensitivity of the FM-CW receiver described here is degraded & the ability to operate with multiple targets is usually poor.

→ for a radio altimeter whose target is earth, these limitations usually present no problem. when used for the long range detection of air targets nonoptimum detection of the classical FM-CW radar will seriously hinder its performance when compared to a properly designed pulse radar.

→ It is possible, however, to utilize matched filter processing in the FM-CW radar, thus to overcome the lower sensitivity & multiple target problems. As the duty cycle of an FM pulse compression waveform Incr- it becomes more like the FM-CW waveform of unity duty cycle.




plot of $J_3(D)$ as a function of distance.

**Multiple frequency CW Radar:-**

→ Although it has been known that CW radar does not measure range, it is possible under some circumstances to do so by measuring the phase of the echo signal relative to the phase of the transmitted signal. consider a CW radar radiating a single freq sine wave of the form $\sin 2\pi f_0 t$.

→ the signal travels to the target at a Range R and returns to the radar after a time $T = 2R/C$. where c is the velocity of propagation. the echo signal received at the radar is $\sin[2\pi f_0(t-T)]$. If the transmitted and received signals are compared in a phase detector, the O/P is proportional to the phase difference b/w the two and is $\Delta\phi = 2\pi f_0 T = 4\pi f_0 R/c$

the phase difference may therefore be used as a measure of the range &

$$R = \frac{c\Delta\phi}{4\pi f_0} = \frac{\lambda}{4\pi}\Delta\phi.$$

however, the measurement of the phase difference $\Delta\phi$ is unambiguous only if $\Delta\phi$ does not exceed $2\pi$ radians.

→ the region of unambiguous range may be extended considerably by utilizing two separate cw signals differing slightly in freq. the unambiguous range in this case corresponds to a half wavelength at the difference freq.

→ the transmitted waveform is assumed to consist of two continuous sine waves of freq $f_1$ and $f_2$ separated by an amount $\Delta f$. for convenience, the amplitudes of all signals are set equal to unity. the voltage waveforms of the two components of the transmitted signal $V_{1T}$ and $V_{2T}$ may be written as

$V_{1T} = \sin(2\pi f_1 t + \phi_1)$

$V_{2T} = \sin(2\pi f_2 t + \phi_2)$

where $\phi_1$ & $\phi_2$ are arbitrary phase angles.

. The echo signal is shifted in freq by the doppler effect. the form of the doppler-shifted signals at each of the two frequencies $f_1$ & $f_2$ 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 $R_0$ = Range to target at a particular time $t = t_0$.
$f_{d1}$ = doppler freq shift associated with freq $f_1$.
$f_{d2}$ = doppler freq shift associated with freq $f_2$.

→ Since the two RF frequencies $f_1$ and $f_2$ are approximately the same (ie, $f_2 = f_1 + \Delta f$, where $\Delta f << f_1$). the doppler freq shifts $f_{d1}$ and $f_{d2}$ are approximately equal to one another therefore we may write $f_{d1} = f_{d2} = f_d$.

→ the receiver seperates the two components of the echo signal and heterodynes each received signal component with the corresponding transmitted waveform and extracts the two doppler freq 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 b/w 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 $R_0 = \dfrac{c\Delta\phi}{4\pi\Delta f}$.

→ the two freq CW technique for measuring range was described as using the doppler freq shift. when the doppler freq is zero, as with a stationary target, it is also possible to extract the phase difference.

→ the two frequencies of the two freq radar were described as being transmitted simultaneously. they may also be transmitted sequentially in some applications by rapidly switching a single RF source.

→ A large difference in freq b/w the two transmitted signals improves the accuracy of the range measurement since large $\Delta f$ means a proportionately large change in $\Delta\phi$ for a given range. however, there is a limit to the value of $\Delta f$, since $\Delta\phi$ cannot be greater than $2\pi$ radians if the range is to remain unambiguous. the max unambiguous range $R_{unamb}$ is

$$\boxed{R_{unamb} = \frac{c}{2\Delta f}}$$

$\Delta f$ must be less than $c/2R_{unamb}$.

→ the two freq CW radar is essentially a 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 phase measurement is doubtful. It can be shown that the theoretical rms range error is $\delta R = \frac{c}{4\pi \Delta f (2E/N_0)^{1/2}}$ where E = energy contained in received signal and $N_0$ = noise power per hertz of bandwidth.

→ In the above equation the greater the seperation of b/w the two frequencies, the less will be the rms error. but, the freq difference must not be too large if unambiguous measurements are to be made. the selection of $\Delta f$ represents a compromise b/w the requirements of accuracy and ambiguity. both accurate and unambiguous range measurements can be made by transmitting three or more frequencies instead of just two.

In addition to its use in surveying, the multiple CW freq method
→ In addition to its use in surveying, the multiple CW freq method in range Instrumentation radar of measuring range has been applied for the measurement of the distance to a transponder equipped missile, the distance to satellites in satellite navigation systems based on range measurements & for detecting presence of an obstacle in the path of a moving automobile by measuring the distance, the doppler velocity and the sign of the doppler.

Limitations of CW radar :-

① In CW radar the two antennas are used so that it effectively decreases gain utilisation.

② The transmitter leakage exists in CW radar.

③ The major limitations of CW radar it doesn't calculate the range because of the lack of timing mark.

④ Impedance mismatching.

⑤ Noise added in cable.

⑥ Due to antenna present in the cable.

⑦ Due to near distance b/w TX and RX i.e. transmitter leakage.

problems:-

① With a transmit (CW) freq of 5 GHz; calculate the doppler freq seen by a stationary radar when the target radial velocity is 100 km/hr. find the doppler frequency.

$1\,km/hr = \frac{5}{18} \times 100\ km/hr.$

Sol. Given $f_0 = 5\,GHz$, $V_r = 100\,km/hr$

$V_r = \frac{500}{18} = 27.7\,m/sec.$

$$f_d = \frac{2V_r}{\lambda} = \frac{2V_r f_0}{c} = \frac{2\times100\times5\times10^9}{3\times10^8} = \frac{2\times27.77\times5\times10^9}{3\times10^8}$$

$$= \frac{2\times27.77\times50}{3} = \frac{100\times27.77}{3} = \frac{2777}{3} = 925.25\,Hz$$

② calculate the doppler freq of stationary CW Radar transmitting at 6 MHz freq when a moving target approaches the radar with a radial velocity of 100 km/hr.

Sol.

$$f_d = \frac{2V_r}{\lambda} = \frac{2\times V_r f_0}{c} = \frac{2\times27.77\times6\times10^6}{3\times10^8} = 4\times27.77\times10^{-2}$$

$$= 111.08\times10^{-2} = 1.11\,Hz.$$

③ A 8 GHz police Radar Measures a doppler freq of 1788 Hz from a car approaching the stationary police vehicle in an 80 km/hr speed limit zone. what should the police officer do?

Sol. $f_d = 1788\,Hz$, $f_0 = 8\,GHz$ Speed $= 80\,km/hr$.

$f_d = \frac{2V_r}{\lambda}$ $\quad \lambda = \frac{2V_r}{f_d} \Rightarrow f_d = \frac{2V_r f_0}{c} \Rightarrow 1788 = \frac{2V_r\times8\times10^9}{3\times10^8}$

$\frac{1788\times3\times10^8}{16\times10^9} = V_r \Rightarrow$

$= 33.525\,m/sec.$

$= 33.525\times\frac{18}{5}$

$V_r = 120.69\,km/hr.$

The car is approaching higher than the speed limit. ∴ the person who is driving the car should be punshed by the police officer.

④ Determine the acceleration of a target with the received signal bandwidth is 40 Hz and the operating wavelength is 9 cm.

Sol. $\lambda = 9\,cm$ $\quad \Delta f_d = 40\,Hz$ $\quad \Delta f_d = \left[\frac{2a_r}{\lambda}\right]^{1/2}$

$$40\times40 = \frac{2\times a_r}{9\times10^{-2}} \Rightarrow a_r = 1800\times9\times10^{-2}$$

$$a_r = 72\ m/sec^2.$$

⑥ For an unambiguous range of 81nm in a two freq CW radar. determine $f_2$ and $\Delta f$ when $f_1 = 4.2\,KHz$. Derive the expression to slove this problem?

Sol:

consider a signal which is to be transmitted at the transmitter is

$s(t) = A \sin 2\pi f_0 t$.

the second echo signal at the receiver is $s_r(t) = A_r(\sin 2\pi f_0 t - \phi)$

the second echo signal shifted from the transmitter signal.

where $\phi$ Indicates the phase angle shifted from the transmitter signal.

$\phi = 2\pi f_0 T$. where $T$ we known $R = \frac{cT}{2} \Rightarrow T = \frac{2R}{c}$. $f = \frac{\phi}{T}$

$\therefore \phi = 2\pi f_0 \times \frac{2R}{c} = 4\pi f_0 R/c \Rightarrow R = \frac{c\phi}{4\pi f_0} = \frac{\lambda\phi}{4\pi}$.

Now consider two signal to be transmitted i.e., $s_1(t) = A_1 \sin 2\pi f_1 t$.

$s_2(t) = A_2 \sin 2\pi f_2 t$.

→ the received echo signal at the receiver is $s_{r1}(t) = A_{r1} \sin 2\pi f_1 t - \phi_1$

$s_{r2}(t) = A_{r2} \sin 2\pi f_2 t - \phi_2$.

$\phi_1 = 4\pi f_1 R/c$

$||^{ly}$ $\phi_2 = 4\pi f_2 R/c$

the diff in phase angle is $\phi_2 - \phi_1 = \frac{4\pi R}{c}(f_2 - f_1)$.

$\Delta\phi = \frac{4\pi R}{c}\Delta f$.

the change in phase angle $\Delta\phi = 2\pi \Rightarrow \cancel{2\pi} = \frac{\cancel{4\pi}^{\,2} R}{c}\Delta f$

$$\boxed{R = \frac{c}{2\Delta f}}$$

Given $R = 81\,nm$

$= 81 \times 1.852 = 150.01\,Km$

$\Delta f = \frac{c}{2R} = \frac{3\times10^8}{2\times150.01\times10^3} = \frac{3\times10^5}{2\times150.01} = \frac{3\times10^5}{300.02} = \frac{3}{30002}\times10^7$

$\Delta f = 1\,KHz \simeq 999.92\,Hz$

$\Delta f = f_2 - f_1$

$1K = f_2 - 4.2\,KHz$

$f_2 = 5.19\,KHz$.

① Determine the range and doppler velocity for a FM CW radar. if the target is approaching the radar. Given the beat freq $f_b(up)$ = 20KHz & $f_b(down)$ = 30KHz. for the triangular modulation. the modulating freq is 1MHz and doppler freq shift is 1KHz.

Sol) $f_r = \frac{4 f_m \Delta f R}{c}$ Given that $f_b(up) = 20KHz \rightarrow ①$

$f_b(down) = 30K \rightarrow ②$

$f_b(up) = f_r - f_d$ $f_m = 1MHz$

$f_b(down) = f_r + f_d$ $\Delta f = 1KHz$

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

$f_r = \frac{20+30}{2} = 25KHz$.

$25 \times 10^3 = \frac{4 \times 1 \times 10^6 \times 10^3 \times R}{3 \times 10^8} \Rightarrow R = \frac{75 \times 10^3}{4 \times 10} = 1.87km$.

$f_d = \frac{1}{2}[f_b(up) - f_b(down)]$

$f_d = \frac{2V_r}{\lambda} = \frac{2V_r f_r}{c} = \frac{c f_d}{2 f_r} = V_r$.

from above ① & ② eq

$f_b(up) - f_b(down) = f_r + f_d - f_r + f_d = 2f_d = 5KHz$.

$V_r = \frac{3 \times 10^8 \times 5}{2 \times 25 \times 10^3} = 3 \times 10^{-7} m/sec$.

② estimate range of FMCW radar if its freq is modulated at rate FM over a range $\Delta f$, derive the exp used. hence calculate the range if $\Delta f = 1.6K$, $f_m = 100K$, $f_r = 30Hz$.

Sol) $f_r = f_0 T \Rightarrow f_0 \cdot \frac{2R}{c} \Rightarrow f_r = \frac{4 f_m \Delta f R}{c}$

$R = \frac{c f_r}{4 f_m \Delta f} = \frac{3 \times 10^8 \times 30}{4 \times 100 \times 10^3 \times 1.6 \times 10^3} = 14.062m$.

③ Determine the range and doppler velocity of target. If the target is moving away from a FMCW radar. the beat freq observed for triangular modulation as $f_b(up) = 50KHz$, $f_b(down) = 20KHz$. the modulating freq is 2MHz & doppler shift $\Delta f = 2KHz$.

Sol) The target is moving away from the Radar ie.,

$f_r > f_d$

$f_b(up) = f_r + f_d$ $f_r = 1/2[f_b(up) + f_b(down)]$

$f_b(down) = f_r - f_d$ $= 35KHz$

we know $f_r = \frac{4 f_m \Delta f R}{c}$ $\quad f_d = \frac{1}{2}[f_b(up) - f_b(down)]$

$= \frac{1}{2}[30KHz] = 15KHz$

$R = \frac{f_r c}{4 f_m \Delta f}$ $\quad f_r < f_d$

$= \frac{35\times10^3 \times 3\times10^8}{4\times2\times10^3\times2\times10^6} = 0.65\,m$

$f_d = \frac{2 V_r f_r}{c} = \frac{c f_d}{2 f_r} = 6.42\times10^7\,m/sec.$

④ A FM CW radar operates at a freq of 9.25 GHz, A symmetrical triangular modulating waveform is used. the magnitude of the slope being 800MHz/sec. the return form a moving target produces a beat freq of 3.5KHz over the +ve slope and 3.85KHz over the negative slope of FM. determine ① target range ② the range rate ③ whether target moving towards (or) away from Radar.

Sol. Given $f_r = 9.25GHz$, $f_0 = 800MHz$, we known $f_r = f_0 T$.

① $f_r = \frac{f_0 2R}{c} \Rightarrow R = \frac{c f_r}{2 f_0} = \frac{3\times10^8\times9.25\times10^9}{2\times500\times10^6} = 1.734\times10^9 m$

② $V_r = \frac{c f_d}{2 f_r}$ $\quad f_b(up) = 3.5K$, $f_b(down) = 3.85K$ Assume the target is moving from the radar $f_d = \frac{1}{2}[f_b(down) - f_b(up)] = \frac{1}{2}[3.85 - 3.5] = 175\,Hz.$

$f_r = \frac{1}{2}[f_b(down) + f_b(up)] = 3675\,Hz.$

$V_r = \frac{c f_d}{2 f_r} = \frac{2\times175}{2\times3675Hz} = 2.83\times10^3\,m/sec.$

③ Take $f_r = 3675\,Hz$ or $36KHz$ $\quad f_r > f_d$. so the target is approaching to radar.

③ Determine the beat freq due to range and quantization error if range =100m and freq excursion is 75Hz and modulation freq is 1KHz.

Sol. $f_r = \frac{4 f_m \Delta f R}{c}$. $\quad R = 100m$, $\Delta f = 75Hz$ $\quad f_m = 1KHz$ $\quad c f_r = 4 f_m \Delta f R$ $\quad R = \frac{c f_r}{4 f_m \Delta f}$

$\delta R = \frac{c}{4\Delta f}$ $\Rightarrow f_r = \frac{4\times10^3\times75\times100}{3\times10^8} = \frac{300\times10^5}{3\times10^8} = 0.1Hz.$

$\delta R = \frac{c}{4\Delta f} = \frac{3\times10^8}{4\times75} = 1000\,km.$

# UNIT-3

# MTI and pulse Doppler Radar

## Introduction :-

→ The doppler freq shift produced by a moving target may be used in a pulse radar, just as in the CW radar. to determine relative velocity of a target & to seperate desired moving targets from undesired stationary objects (clutter). Although there are applications of pulse radar where a determination of the target's relative velocity is made from the dopper freq shift, the use of doppler to separate small moving targets that ~~utilizes~~ ~~the doppler freq shift~~ in the presense of large clutter has probably been of far greater interest. such a pulse radar that utilizes the doppler freq shift as a means for discriminating moving from fixed targets is called an MTI (moving target Indication) or a pulse doppler radar.

→ MTI is a necessity in high-quality air-surveillance radars that operate in the presence of clutter. Its design is more challenging than that of a simple pulse radar or a simple CW radar.

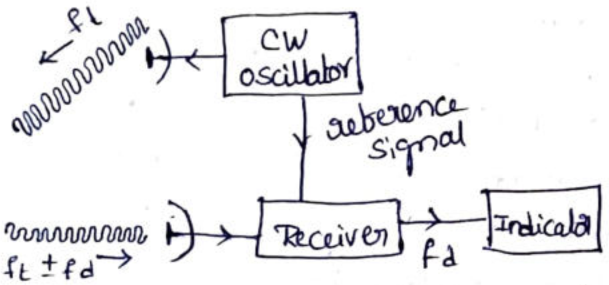
## principle of MTI Radar :-

→ A simple CW radar is shown in the fig(1) It consists of a transmitter, receiver, Indicator and the necessary antenna. In principle, the CW radar may be converted into a pulse radar as shown in fig (2) by providing a power amplifier and a modulator to turn the amplifier on and off for the purpose of generating pulses. this CW signal does more than function as a replacement for the local oscillator.

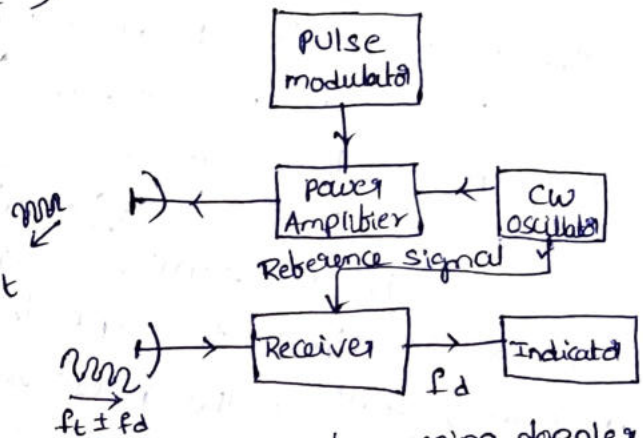
→ It acts as the coherent reference needed to detect the doppler frequency shift. By coherent it is meant that the phase of the transmitted signal is preserved in the reference signal. If the CW oscillator voltage is represented as $A_1 \sin 2\pi f_t t$, where $A_1$ is the amplitude and $f_t$ the carrier freq, the reference signal is $V_{ref} = A_2 \sin 2\pi f_t t$ → (1) and the doppler shifted echo signal 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) where $A_2$ = amplitude of reference signal, $A_3$ = amplitude of signal received from a target at a range $R_0$, $f_d$ = doppler freq shift, $t$ = time, $c$ = velocity of propagation.

→ the reference signal and the target echo signal are heterodyned in the mixer stage of the receiver. only the low freq (difference freq) component from the mixer is of Interest and is a voltage given by

$$V_{diff} = A_4 \sin\left(2\pi f_d t - \frac{4\pi f_t R_0}{c}\right) \rightarrow (3)$$



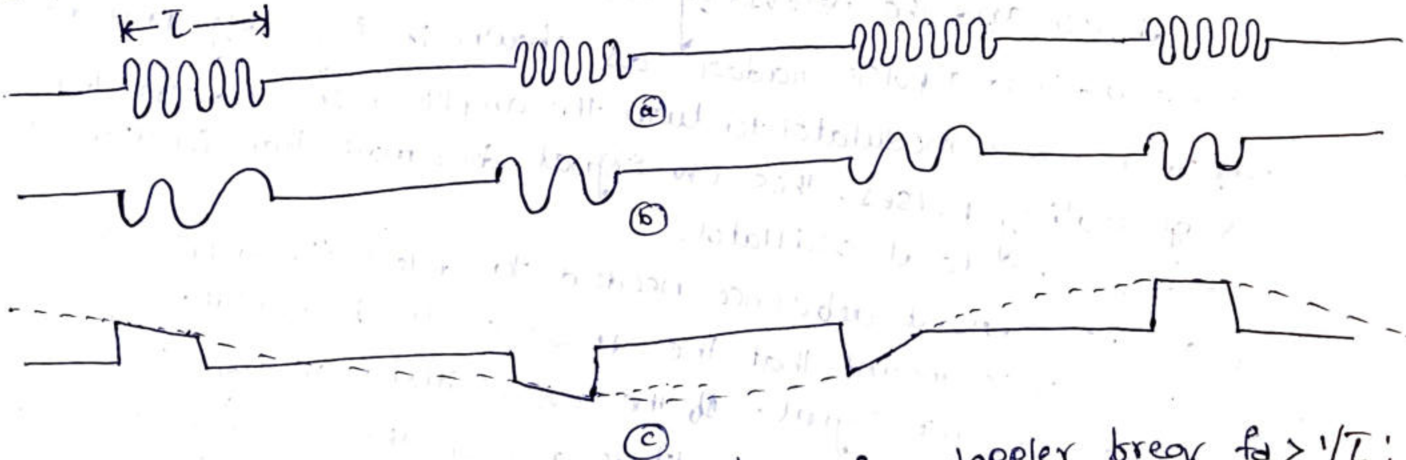

(a) Simple CW radar.




(b) Pulse radar using doppler Information.

→ equation (1) and (2) represents sine wave carriers upon which the pulse modulation is Imposed. the difference freq is equal to the doppler freq $f_d$. for stationary targets the doppler freq shift $f_d$ will be zero. hence $V_{diff}$ will not vary with time and may take on any constant value from $+A_4$ to $-A_4$ Including zero.

→ However, when the target is in motion relative to the radar, $f_d$ has a value other than zero and the voltage corresponding to the difference freq from the mixer ~~as~~ eq (3) will be a function of time.



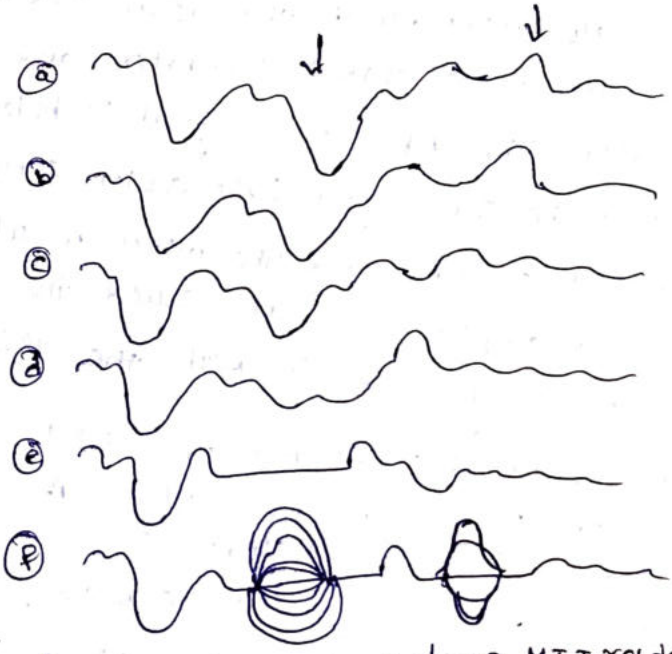

(a) RF echo pulse train (b) video pulse train for doppler freq $f_d > 1/\tau$;
(c) video pulse train for doppler freq $f_d < 1/\tau$.

→ An example of the o/p from the mixer when the doppler freq $f_d$ is large compared with the reciprocal of the pulse width shown in fig (b). the doppler signal may be readily discerned from the Information contained in a single pulse. If on the other hand, $f_d$ is small compared with the reciprocal of the pulse duration, the pulses will be modulated with an amplitude given by eav ③ and many pulses will be needed to extract the doppler Information.

→ Ambiguities in the measurement of doppler freq can occur in the case of discontinuous measurements of fig © but not when the measurement is made on the basis of a single pulse. the signals shown in the fig are called bipolar, since they contain both positive and negative amplitudes.

→ Moving targets may be distinguished from stationary targets by observing the video o/p on an A-scope (Amp vs range). A single sweep ~~show~~ on an A-scope might appear as shown in fig ③. this sweep shows several fixed targets and two moving targets Indicated by the two arrows.

→ Echoes from fixed targets remain constant throughout, but echoes from moving targets vary in amplitude from sweep to sweep at a rate corresponding to the doppler freq. the superposition of the successive A-scope sweeps is shown in the fig ③ Ⓕ. the moving targets produce, with time, a "butterfly" effect on the A-scope.



(a-e) successive sweeps of an MTI radar A-scope display. Ⓕ superposition of many sweeps. arrows indicates position of moving targets.

→ Although the butterfly effect is suitable for recognizing moving targets on an A-scope, it is not appropriate for display on the PPI.

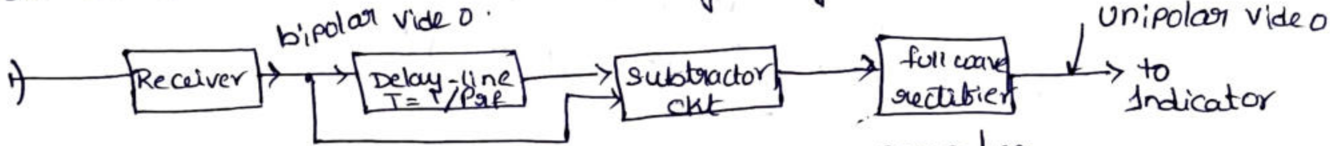
→ one method commonly employed to extract doppler Information in a form suitable for display on the PPI scope is with a delay-line canceler

→ the delay line canceler acts as a filter to eliminate the d-c component of fixed targets and to pass the a-c components of moving targets.

→ the video portion of the receiver is divided into two channels. one is normal video channel. In the other, the video signal experiences a time delay equal to one pulse repetition period

→ the o/p's from the two channels are subtracted from one another. the fixed targets with unchanging amplitudes from pulse to pulse are canceled on subtraction.

→ However, the amplitudes of the moving targets echoes are not constant from pulse to pulse, and subtraction results in an uncanceled residue. the O/P of the subtraction ckt is bipolar video, just as was the I/P. before bipolar video can intensity modulate a PPI display, it must be converted to unipotential voltages by a full wave rectifier.



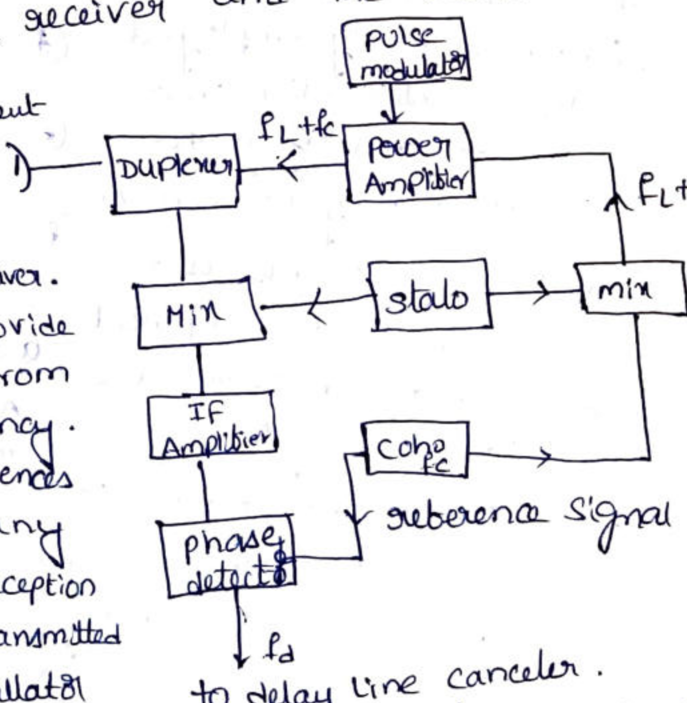

MTI receiver with delay-line canceler.

MTI Radar with power Amplifier transmitter:-

→ The block diagram of a more common MTI radar employing a power amplifier is shown in the fig. The coherent reference is supplied by an oscillator called the coho, which stands for coherent oscillator. the coho is a stable oscillator whose freq is the same as the intermediate freq used in the receiver.

→ In addition to providing the reference signal, the O/P of the coho $f_c$ is also mixed with the local oscillator freq $f_L$. the local oscillator must also be a stable oscillator and is called stalo, for stable local oscillator. The RF echo signal is heterodyned with the stalo signal to produce the IF signal just as in conventional superheterodyne receiver.

→ The stalo, coho and the mixer in which they are combined plus any low-level amplification are called the receiver-exciter because of the dual role they serve in both the receiver and the transmitter.




→ The characteristic feature of coherent MTI radar is that the transmitted signal must be coherent (inphase) with the reference signal in the receiver.

→ the function of the stalo is to provide the necessary frequency translation from the IF to the transmitted (RF) frequency. Although the phase of the stalo influences the phase of the transmitted signal, any stalo phase shift is canceled on reception bcoz the stalo that generates the transmitted signal also acts as the local oscillator in the receiver.

Block diagram of MTI radar with power amplifier transmitter.

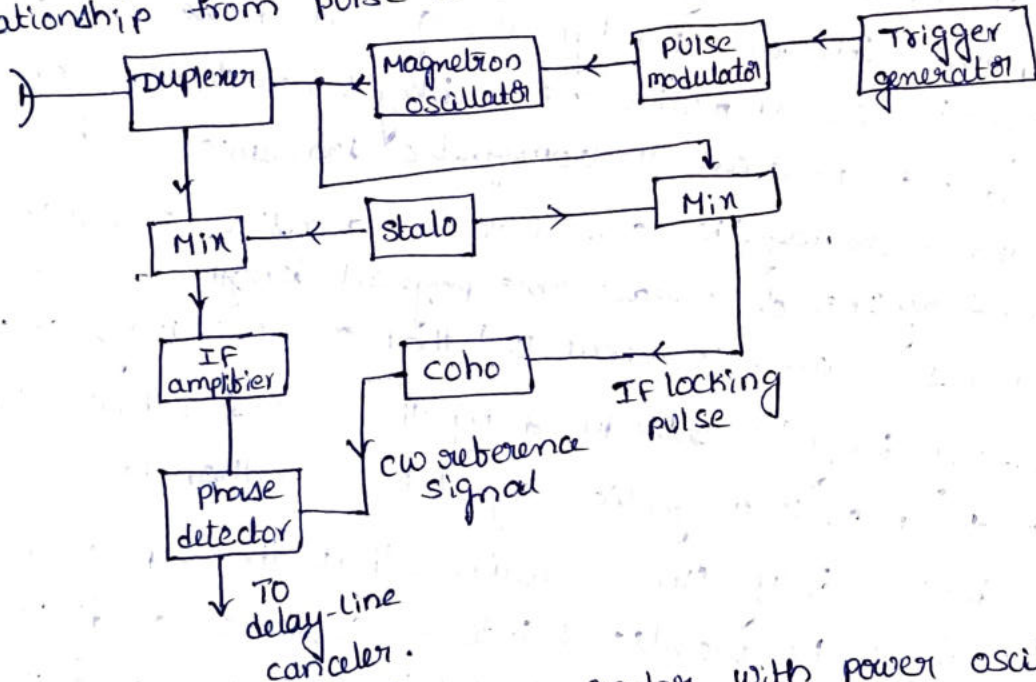
→ The reference signal from the coho and the IF echo signal are both fed into a mixer called the phase detector. the phase detector differs from the normal amplitude detector since its o/p is proportional to the phase difference b/w the two i/p signals.

→ Any one of a number of transmitting-tube types might be used as the power amplifier. these include the triode, tetrode, klystron, travelling-wave tube, and the crossed-field amplifier.

MTI Radar with power oscillator transmitter :-

→ A transmitter which consists of a stable low-power oscillator followed by a power amplifier is sometimes called MOPA, which stands for master oscillator power amplifier.

→ Before the development of the klystron amplifier, the only high power transmitter available at microwave frequencies for radar application was the magnetron oscillator. In an oscillator the phase of the RF bears no relationship from pulse to pulse.




Block diagram of MTI radar with power oscillator transmitter.

→ For this reason the reference signal cannot be generated by a continuously running oscillator. however, a coherent reference signal may be readily obtained with the power oscillator by readjusting the phase of the coho at the beginning of each sweep according to the phase of the transmitted pulse. the phase of the coho is locked to the phase of the transmitted pulse each time a pulse is generated.

A block diagram of an MTI radar is shown in the fig. A portion of the transmitted signal is mixed with the stalo O/P to produce an IF beat signal whose phase is directly related to the phase of the transmitter.

→ This IF pulse is applied to the coho and causes the phase of the coho CW oscillation to "lock" in step with the phase of the IF reference pulse. the phase of the coho is then related to the phase of the transmitted pulse and may be used as the reference signal for echoes received from that particular transmitted pulse.

→ upon the next transmission another IF locking pulse is generated to relock the phase of the CW coho until the next locking pulse comes along. this type of MTI radar has had wide applications.

Delay-Line cancelers :-

The capability of this device depends on the quality of the medium used as the delay line. the delay line must introduce a time delay equal to the pulse repetition interval. for typical ground-base air surveillance radars this might be several milliseconds. delay times of this magnitude cannot be achieved with practical electromagnetic transmission lines.

→ By converting the electromagnetic signal to an acoustic signal it is possible to utilize delay lines of a reasonable physical length since the velocity of propagation of acoustic waves is about $10^{-5}$ that of electromagnetic waves.

→ After the necessary delay is introduced by the acoustic line, the signal is converted back to an electromagnetic signal for further processing.

→ The use of digital delay lines requires that the O/P of the MTI receiver phase detector be quantized into a sequence of digital words. the compactness and convenience of digital processing allows the implementation of more complex delay-line cancelers with filter characteristics not practical with analog methods.

→ one of the advantages of a time-domain delay-line canceler as compared to the more conventional freq domain filter is that a single n/w operates at all ranges and does not require a separate filter for each range resolution cell. freq domain doppler filter banks are of interest in some forms of MTI and pulse-doppler radar.

## Filter characteristics of the delay-line canceler :-

→ the delay line canceler acts as a filter which rejects the d-c component of clutter. because of its periodic nature, the filter also rejects energy in the vicinity of the pulse repetation freq and its harmonics.

→ the video signal received from a particular target at a range $R_0$ is $V_1 = K\sin(2\pi f_d t - \phi_0)$ →① where $\phi_0$ = phase shift and K = amplitude of video signal. the signal from the previous transmission, which is delayed by a time T = pulse repetition interval is

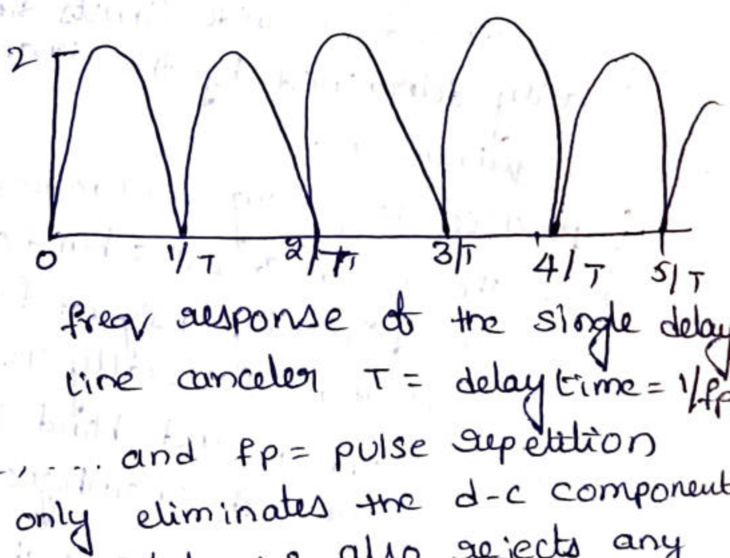
$$V_2 = k\sin[2\pi f_d(t-T) - \phi_0] \rightarrow ②$$

→ Everything else is assumed to remain essentially constant over the interval T so that k is the same for both pulses. the o/p from the subtractor is $V = V_1 - V_2 = 2k\sin\pi f_d T\cos[2\pi f_d(t - T/2) - \phi_0] \rightarrow ③$

→ It is assumed that the gain through the delay-line canceler is unity. the o/p from the canceler consists of a cosine wave at the doppler freq $f_d$ with an amplitude $2k\sin\pi f_d T$.

→ thus the amplitude of the canceled video o/p is a function of the doppler freq shift and the pulse repetation interval or prf.

→ the magnitude of the relative freq response of the delay-line canceler is shown in the fig.




freq response of the single delay line canceler T = delay time = $1/f_p$

## Blind speeds :-

→ the response of the single delay line canceler will be zero whenever the argument $\pi f_d T$ in the eq ③ the amplitude factor is $0, \pi, 2\pi \ldots$, etc., or when

$$f_d = \frac{n}{T} = n f_p. \rightarrow ①$$ where $n = 0, 1, 2, \ldots$ and $f_p$ = pulse repetition freq. the delay-line canceler not only eliminates the d-c component caused by clutter $(n=0)$ but unfortunately is also rejects any moving targets whose doppler freq happens to be the same as the prf or a multiple thereof.

→ Those relative target velocities which result in zero MTI response are called blind speeds and are given by

$$V_n = \frac{n\lambda}{2T} = \frac{n\lambda f_p}{2} \qquad n = 1, 2, 3 \ldots \rightarrow ②$$

where $V_n$ is the $n^{th}$ blind speed. If $\lambda$ is measured in meters, $f_p$ in Hz, and the relative velocity in knots the blind speeds are

$$v_n = \frac{n\lambda f_p}{1.02} \simeq n\lambda f_p \rightarrow ③$$

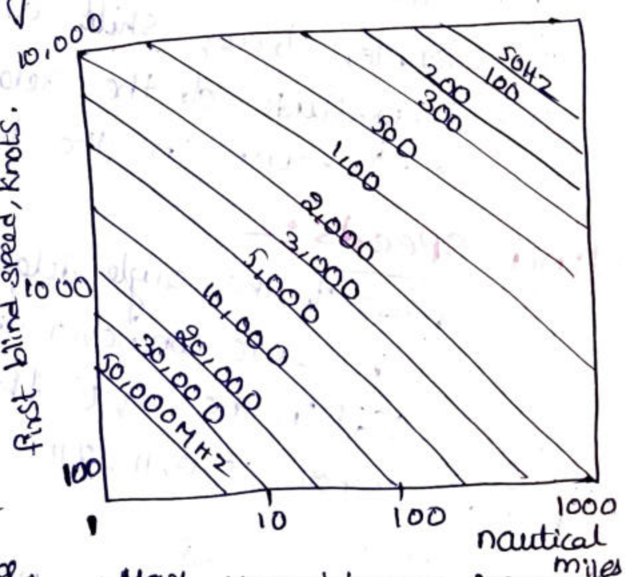
→ the blind speeds are one of the limitations of pulse MTI radar which do not occur with CW radar. they are present in pulse radar bcoz doppler is measured by discrete samples at the prf rather than continuously.

→ If the first blind speed is to be greater than the max radial velocity expected from the target, the product $\lambda f_p$ must be large. thus the MTI radar must operate at long wavelengths (low frequencies) or with high pulse repetition frequencies, or both.

→ Unfortunately, there are usually constraints other than blind speeds which determine the wavelengths and the pulse repetition freq. therefore blind speeds might not be easy to avoid. low radar frequencies have the disadvantage that antenna beam widths, for a given size antenna, are wider than at the higher frequencies and would not be satisfactory in applications where angular accuracy or angular resolution is important.

→ the pulse repetition frequency cannot always be varied over wide limits since it is primarily determined by the unambiguous range requirement.

→ In practice, long range MTI radars that operate in the region of L or S band or higher and are primarily designed for the detection of aircraft must usually operate with ambiguous doppler and blind speeds if they are to operate with unambiguous range.

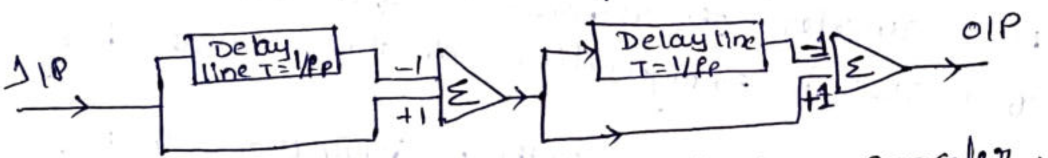



→ the presence of blind speeds within the doppler freq band reduces the detection capabilities of radar. the effect of blind speeds can be significantly reduced, without incurring range ambiguities, by operating with more than one pulse repetition frequency.

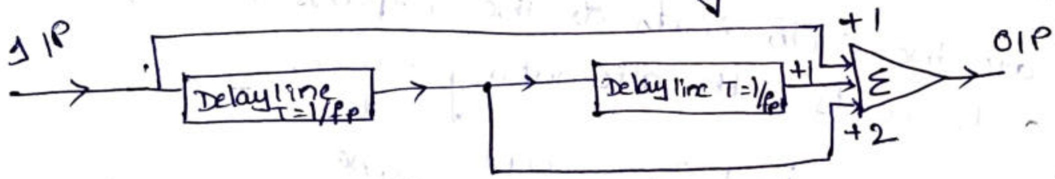
→ This is called a staggered - prf MTI. Operating at more than one RF freq can also reduce the effect of blind speeds.

<u>Double cancellation:-</u>

→ The frequency response of a single delay line used previously does not always have as broad a clutter rejection null as might be desired in the vicinity of d-c. The clutter rejection notches may be widened by passing the O/P of the delay-line canceler through a second delay line canceler as shown in fig ⓐ.




ⓐ Double delay-line canceler.




three pulse canceler.

→ The O/P of the two single-delay line cancelers in cascade is the square of that from a single canceler. Thus the freq response is $4\sin^2 \pi f_d T$. The relative response of the double canceler compared with that of a single-delay canceler is shown in fig. The finite width of the clutter spectrum is also shown in the fig to illustrate the additional cancellation of clutter offered by the double canceler.

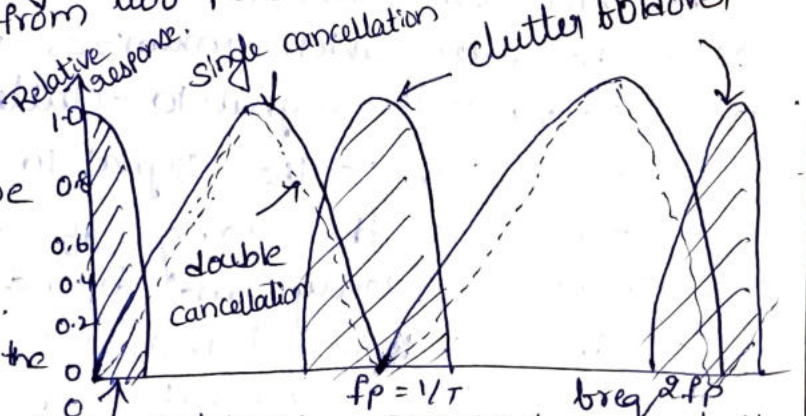
→ The two delay line configuration in fig ⓑ has the same freq response characteristics as the double-delay-line canceler. The operation of the device is as follows. A signal $f(t)$ is inserted into the adder along with the signal from the preceding pulse period, with its amplitude weighted by the factor -2, plus the signal from two pulse periods previous. The O/P of the adder is

$\therefore f(t) - 2f(t+T) + f(t+2T)$,

which is the same as the O/P from the double delay-line canceler

$f(t) - 2f(t+T) - f(t+T) + f(t+2T)$.

This configuration is commonly called the three phase canceler.




0

clutter spectrum

Relative freq response of the single delay line canceler and the double delay line canceler.

Transversal filter :-

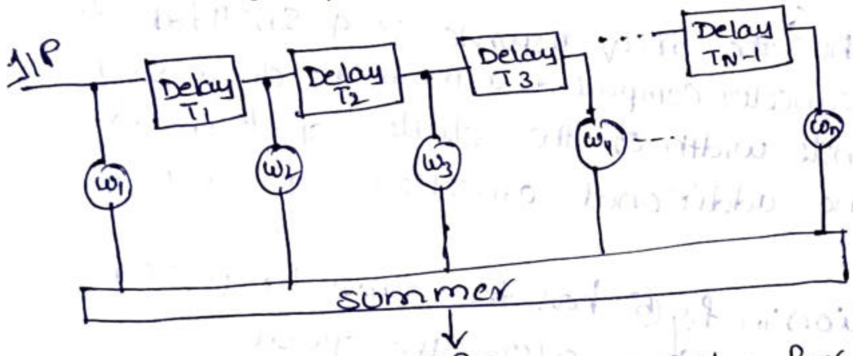
→ The three pulse canceler shown in the fig previously is an example of a transversal filter. Its general form with N pulses and N-1 delay lines is shown in the fig. It is also sometimes known as a feed forward filter, a nonrecursive filter, a finite memory filter & a tapped delay line filter.

→ The weights $w_i$ for a three pulse canceler utilizing two delay lines arranged as a transversal filter are 1, -2, 1. the freq response function is proportional to $\sin^2 \pi f_d T$. A transversal filter with three delay lines whose weights are 1, -3, 3 -1 gives a $\sin^3 \pi f_d T$ response.

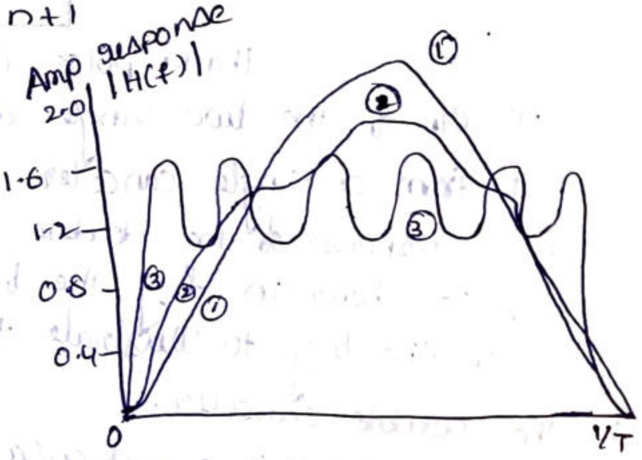
→ This is a four-pulse canceler. Its response is equivalent to a triple canceler consisting of a cascade configuration of three single-delay-line canceler.

→ The weight for a transversal filter with n delay lines that gives a response $\sin^n \pi f_d T$ are the coefficients of the expansion of $(1-x)^n$ which are the binomial coefficients with alternating signs.

$$w_i = (-1)^{i-1} \frac{n!}{(n-i+1)!\,(i-1)!} \quad , \quad i = 1, 2 \ldots n+1$$




General form of a transversal filter for MTI signal processing.




① classical three pulse canceler. ② five pulse delay-line canceler with optimum weights and ③ 15-pulse chebyshev design.

→ The transversal filter with alternating binomial weights is closely related to the filter which maximizes the avg of the ratio. $I_C = \frac{(S/C)_{out}}{(S/C)_{in}}$. where $(S/C)_{out}$ is the signal-to-clutter ratio at the O/P of the filter and $(S/C)_{in}$ is the signal to clutter ratio at the I/P. the avg is taken over the range of doppler frequencies. It is independent of target velocity and depends only on the weights $w_i$.

→ the difference b/w a transversal filter with optimal weights and one with binomial weights for a three pulse canceler is less than 2dB.

AN N-1 delay line canceler requires N pulses, which sets a restriction on the radar's pulse repetition freq, beamwidth and antenna rotation rate or dwell time. the N-pulse nonrecursive delay-line canceler allows the designer 'N-zeros' for synthesizing the freq response. the result is that many delay lines are required for highly-shaped filter responses.

→ There are limits to the number of delay lines that can be employed therefore other approaches to MTI filter implementation are sometimes desired.

## Multiple, or staggered, pulse repetition frequencies :-

→ The use of more than one pulse repetition frequency offers additional flexibility in the design of MTI doppler filters. It not only reduces the effect of the blind speeds, but it also allows a sharper low-freq cutoff in the freq response than might be obtained with a cascade of single delay line cancelers with $\sin^n \pi f_d T$ response.

→ The blind speeds of two independent radars operating at the same frequency will be different in their pulse repetition frequencies are different. therefore if one radar were "blind" to moving targets, it would be unlikely that the other radar would be "blind" also. Instead of using two seperate radars, the same result can be obtained with one radar which time-shares its pulse repetition freq b/w two or more different values.

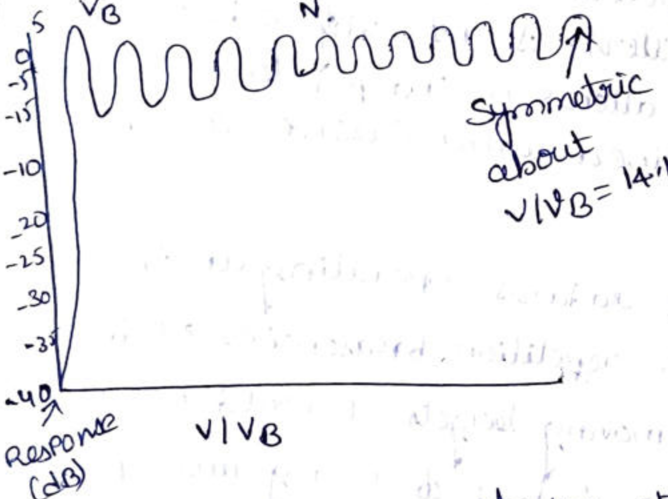
→ the pulse repetition freq might be switched every other scan or every time the antenna is scanned a half beamwidth, & the period might be alternated on every other pulse. when the switching is pulse to pulse, it is known as a staggered prf.

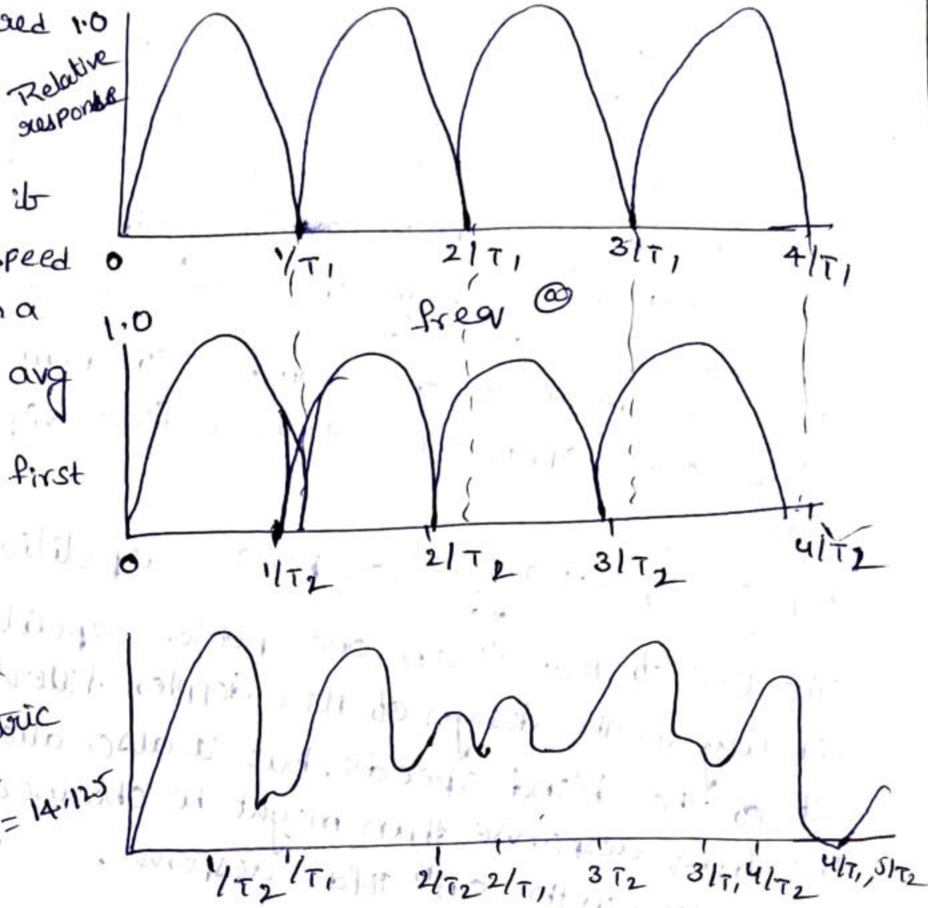
→ The composite (avg) response of an MTI radar operating with two separate pulse repetition frequencies on a time-shared basis as shown in the fig. the pulse repetition frequencies are in the ratio of 5:4. zero response occurs only when the blind speeds of each prf coincide. In the fig the blind speeds are coincident for $4/T_1 = 5/T_2$. Although the first blind speed may be extended by using more than one prf, regions of low sensitivity might appear within the composite pass band. the closer the ratio $T_1 : T_2$ approaches unity, the greater will be the value of the first blind speed.

→ If the periods of the staggered waveforms have the relationship $n_1/T_1 = n_2/T_2 = \dots n_N/T_N$ where $n_1, n_2 \dots n_N$ are integers, and if $V_B$ is equal to the first blind speed of a nonstaggered waveform with a constant period equal to the avg period

$T_{av} = (T_1 + T_2 + \dots T_N)/N$ then the first blind speed $V_1$ is

$$\frac{V_1}{V_B} = \frac{n_1 + n_2 \dots + n_N}{N}$$




freq response of a single-delay line canceler for $f_p = 1/T_1$; ⓑ same for $f_p = 1/T_2$ ⓒ composite response with $T_1/T_2 = 4/5$.

→ The disadvantage of the 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 thus produces uncanceled residue. Second time around clutter echoes can be removed by use of a constant prf, providing there is pulse-to-pulse coherence as in the power amplifier form of MTI.

## Range - Gated Doppler filters :-

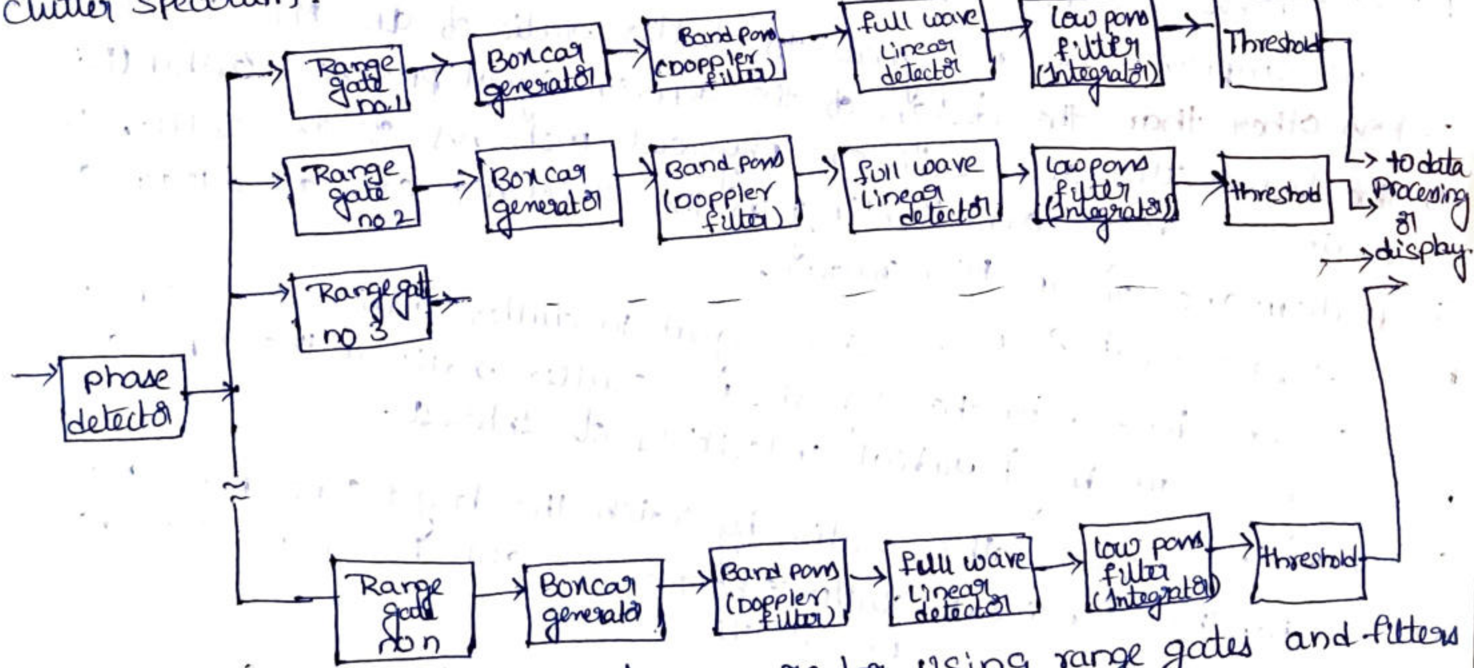
→ The delay-line canceler, which can be considered as a time-domain filter, has been widely used in MTI radar as the means for separating moving targets from stationary clutter. It is also possible to employ the more usual freq domain bandpass filters of conventional design in MTI radar to sort the doppler freq shifted targets.

→ The filter configuration must be more complex, however, than the single narrowband pass filter. A narrow band filter with a pass band designed to pass the doppler freq components of moving targets will "ring" when excited by the usual short radar pulse.

→ the loss of the range Information and the collapsing loss may be eliminated by first quantizing the range into small Intervals. this process is called range gating. the width of the range gates depends upon the range accuracy desired and the complexity which can be tolerated, but they are usually of the order of the pulse width. Range resolution is established by gating.

→ The block diagram of the video of an MTI radar with multiple range gates followed by clutter- rejection filters as shown in the fig. the O/P of the phase detector is sampled sequentially by the range gates. the range gate acts as a switch or a gate which opens and closes at the proper time.

→ the range gates are activated once each pulse repetition Interval. the O/P for a stationary target is a series of pulses of constant amplitude. An echo from a moving target produces a series of pulses which vary in amplitude according to the doppler frequency. the O/P of the range gates is stretched in a ckt called the boxcar generator, or sample-and-hold ckt, whose purpose is to aid in the filtering and detection process by emphasizing the fundamental of the modulation freq & eliminating harmonics of the pulse repetition freq. the clutter rejection filter is a bandpass filter whose bandwidth depends upon the extent of the Expected clutter spectrum.




Block diagram of MTI radar using range gates and filters and an Integrator

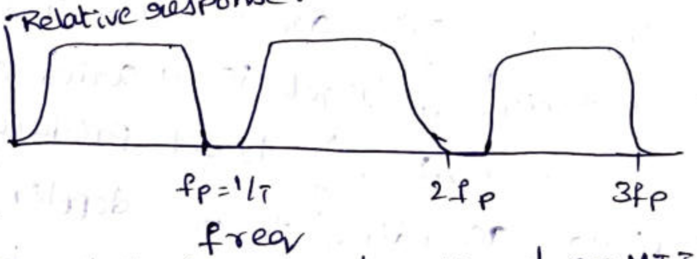
→ following the doppler filter is a full wave linear detector and an Integrator. the purpose of the detector is to convert the bipolar video to unipolar video.

→ the O/P of the Integrator is applied to a threshold detection ckt. only those signals which cross the threshold are reported as targets. following the threshold detector, the O/P's from each of the range channels must be properly combined for display on the PPI or A-scope or for any other appropriate Indicating or data processing device.

→ The bandpass filter can be designed with a variable low-frequency cutoff that can be selected to conform to the prevailing clutter condition. the selection of the lower cutoff might be at the option of the operator or it can be done adaptively. A variable lower cutoff might be advantageous when the width of the clutter spectrum changes with time as when the radar receives unwanted echoes from birds.

→ MTI radar using gates and filters is usually more complex than an MTI with a single delay line canceler.

→ the additional complexity is justified in those applications where good MTI performance and the flexibility of the range gets and filter MTI are desired. the better MTI performance results from the better MTI performance results from the better match b/w the clutter filter characteristics and the clutter spectrum.




freq response characteristics of an MTI using range gates & filters.

## Limitations To MTI performance:-

→ The Improvement in signal-to clutter ratio of an MTI is affected by factors other than the design of the doppler signal processor: Instabilities of the transmitter and receiver, physical motions of the clutter, the finite time on target and limiting in the receiver can all detract from the performance of an MTI radar.

**MTI Improvement factor:-** The signal to clutter ratio at the O/P of the MTI system divided by the signal-to-clutter ratio at the I/P, averaged uniformly over all target radial velocities of Interest.

**Subclutter Visiblity:-** the ratio by which the target echo power may be weaker than the coincident clutter echo power & still be detected with specified detection and false alaram probabilities.

**clutter visibility factor:-** the signal to clutter ratio, after cancellation or doppler filtering, that provides stated probabilities of detection and false alaram.

clutter attenuation :- The ratio of clutter power at the canceler I/P to the clutter residue at the O/P, normalized to the attenuation of a single pulse passing through the unprocessed channel of the canceler.

cancellation ratio :- The ratio of canceler voltage amplification for the fixed target echoes received with a fixed antenna, to the gain for a single pulse passing through the un-processed channel of the canceler.

Interclutter visibility :- this describes the ability of an MTI radar to detect moving targets which occur in the relatively clear resolutions cells b/w patches of strong clutter.

① Equipment Instabilities :-

→ pulse-to-pulse changes in the amplitude, freq or phase of the transmitter signal, changes in the stalo or coho oscillators in the receiver, jitter in the timing of the pulse transmission, variations in the time delay through the delay lines and changes in the pulse width can cause the apparent freq spectrum from perfectly stationary clutter to broaden and thereby lower the Improvement factor of an MTI radar.

→ The stability of the equipment in an MTI radar must be considerably better than that of an ordinary radar. It can limit the performance of an MTI radar if sufficient care is not taken in design, construction and maintenance.

→ consider the effect of phase variations in an oscillator. If the echo from stationary clutter on the first pulse is represented by A cosωt & from the second pulse is A cos (ωt + Δφ), where Δφ is the change in oscillator phase b/w the two, then the difference b/w the two after subtraction is A cos ωt − A cos (ωt+Δφ) = 2A sin(Δφ/2) sin(ωt+Δφ/2). For small phase errors, the amplitude of the resultant difference is 2A sin Δφ/2 ≃ AΔφ. therefore the limitation on the Improvement factor due to oscillator Instability is

$I = \frac{1}{(\Delta\phi)^2}$. this would apply to the coho locking to the phase change introduced by a power amplifier.

Internal fluctuation of clutter :- Although clutter targets such as buildings water towers, bare hills or mountains produce echo signals that are constant in both phase and amplitude as a function of time. there are many types of clutter that cannot be considered as absolutely stationary

8 8

→ Echoes from trees, vegetation, sea, rain and chaff fluctuate with time and these fluctuations can limit the performance of MTI radar. the echo at the radar receiver is the vector sum of the echo signals received from each of the Individual scatters.

→ the experimentally measured power spectra of clutter signals may 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 as a function of freq

$g(f)$ = fourier transform of I/P waveform (clutter echo).

$f_0$ = radar carrier freq

$a$ = a parameter dependent upon clutter.

The Improvement factor can be written as.

$$I = \left(\frac{S_o/C_o}{S_i/C_i}\right)_{ave} = \left(\frac{S_o}{S_i}\right)_{ave} \times \frac{C_i}{C_o} = \left(\frac{S_o}{S_i}\right)_{ave} \times CA.$$

where $S_o/C_o$ = O/P signal-to-clutter ratio, $S_i/C_i$ = I/P signal to clutter ratio, CA = clutter attenuation. for single ~~line~~ delay line canceler the clutter attenuation is $CA = \frac{\int_0^\infty W(f)\,df}{\int_0^\infty W(f)|H(f)|^2\,df}$.

$$CA = \frac{f_p^2}{4\pi^2\sigma_c^2} = \frac{f_p^2\lambda^2}{16\pi^2\sigma_v^2} = \frac{af_p^2}{2\pi^2 f_0^2}$$ where $f_p$ is the pulse repetition freq $f_p = 1/T$.

→ the average gain $(S_o/S_i)_{ave}$ of the single delay line canceler can be shown equal to 2. therefore, the Improvement factor is

$$I_{1C} = \frac{f_p^2}{2\pi^2\sigma_c^2} = \frac{f_p^2\lambda^2}{8\pi^2\sigma_v^2} = \frac{af_p^2}{\pi^2 f_0^2}$$

similarly for a double canceler, whose average gain is 6, the Improvement factor is

$f_p$ = pulse repetition freq

$\sigma_c$ = rms freq response

$\sigma_c = \frac{2\sigma_v}{\lambda}$ → rms velocity.

$$I_{2C} = \frac{f_p^2}{8\pi^4\sigma_c^4} = \frac{f_p^4\lambda^4}{128\pi^4\sigma_v^4} = \frac{a^2 f_p^4}{2\pi^4 f_0^4}.$$

The general expression for Improvement factor for an N-pulse canceler with $N_i = N-1$ delay line is

$$I_{NC} = \frac{2^{N_i}}{N_i!}\left(\frac{f_p}{2\pi\sigma_c}\right)^{2N_i}.$$

Antenna scanning Modulation :- As the antenna scans by a target, its observes the target for a finite time equal to $t_0 = n_B/f_P = \theta_B/\theta_S$ where $n_B$ = no of hits received, $f_P$ = Pulse repetition freq $\theta_B$ = antenna beamwidth & $\theta_S$ = antenna scanning rate. the received pulse train of finite duration $t_0$ has a freq spectrum whose width is proportional to $1/t_0$.

→ If the clutter spectrum is too wide bcoz the observation time is too short, it will affect the Improvement factor. this limitation has some times been called scanning fluctuations & scanning modulation.

→ the clutter attenuation is $CA = \dfrac{\int_0^{\alpha} W_S(f)\,df}{\int_0^{\alpha} W_S(f)\,|H(f)|^2\,df}$.

where $H(f)$ is the freq response function of the MTI signal processor. If the antenna main beam pattern is approximated by the gaussian shape, the spectrum will also be gaussian.

→ the voltage waveform of the received signal is modulated by the square of the antenna electric field strength - pattern, which is equal to the antenna power pattern $G(\theta)$, described by the gaussian function as

$G(\theta) = G_0 \exp\left(-\dfrac{2.776\theta^2}{\theta_B^2}\right)$. the modulation of the received signal due to the antenna pattern is $S_a(t) = k \exp\left(-\dfrac{2.776\,t^2}{t_0^2}\right)$.

the standard deviation $\sigma_f = \dfrac{1.178}{\pi t_0}$. this applies to the voltage spectrum since the standard deviation of the power spectrum is less than that of the voltage spectrum by $\sqrt{2}$, the power spectrum due to antenna scanning can be described by a standard deviation

$$\sigma_S = \frac{1.178}{\sqrt{2}\pi t_0} = \frac{1}{3.77 t_0}.$$

→ to obtain the limitation ~~due~~ due to the Improvement factor caused by antenna scanning. these are

$$I_{1S} = \frac{n_B^2}{1.388} \text{ (single canceler).}$$

$$I_{2S} = \frac{n_B^4}{3.853} \text{ (double canceler).}$$

**Limiting in MTI radar:-** A limiter is usually employed in the IF amplifier before the MTI processor to prevent the residue from large clutter echoes from saturating the display. Ideally an MTI radar should reduce the clutter to a level comparable to receiver noise. Unfortunately, nonlinear devices such as limiters have side-effects that can degrade performance. Limiters cause the spectrum of strong clutter to spread into the cancelers passband and result in the generation of additional residue that can significantly degrade MTI performance as compared with a perfect linear system.

**problems:-**

① calculate the second blind speed of MTI radar whose operating wavelength is 5cm and PRF is 2000 Hz.

Sol: given $\lambda = 5cm$ $f_p = 2000\ Hz$

$$V_2 = \frac{n\lambda f_p}{2} = \frac{2 \times 5 \times 10^{-2} \times 2000}{2} = 100\ m/s.$$

② calculate the lowest blind speed of an MTI system operating at 3.6cm wavelength and transmitting at a pulse repetition time of 330 μsec.

Sol: ① $T = 330\mu sec \Rightarrow f_p = \frac{10^6}{330}$

$\lambda = 3.6cm$

$$V_1 = \frac{n\lambda f_p}{2} = \frac{3.6 \times 10^{-2} \times 10^6}{330 \times 2} = \frac{3.6 \times 10^4}{660} = \frac{1.2 \times 10^3}{22}$$

$$= \frac{0.6 \times 10^3}{11} = \frac{600}{11} = 54.54$$

$\therefore V_1 = 54.54\ m/sec.$

② for $\lambda = 4.2cm$ $\tau = 286\mu sec.$

$$\tau = 286\ \mu sec. \quad V_1 = \frac{n\lambda f_p}{2} = \frac{4.2 \times 10^{-2} \times 10^6}{2 \times 286}$$

$V_1 = 73.4\ m/sec.$

# UNIT - 4

## Tracking Radar.

### Tracking with Radar:-

→ A tracking radar system measures the coordinates of a target and provides data which may be used to determine the target path & to predict its future position. available radar data - range, elevation angle, azimuth angle and doppler ~~shift~~ frequency shift may be used in predicting future position. ie, a radar might track in range, in angle in doppler or with any combination.

→ But, In general, it is the method by which angle tracking is accomplished that distinguish b/w a continuous tracking radar and a track-while scan (TWS) radar. 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 mono pulse.

→ The tracking radar must first find its target before it can track. Some radars operate in a search, or acquistion mode in order to find the target before switching to a tracking mode.

→ when the radar is used in its tracking mode, it has no knowledge of other potential targets. Also, if the antenna pattern is a narrow pencil beam & if the search volume is large, a relatively long time might be required to find the target. therefore many radar tracking system employ a seperate search radar to provide the Information necessary to position the tracker on the target.

→ A search radar, when used for this purpose, it called an acquistion radar. the acquisition radar designates targets to the tracking radar by providing the coordinates where the targets are to be found.

→ A surveillance radar that provides target tracks is sometimes called a track-while-scan radar. landing radars used for GCA (ground control of appraach) and some missile control radars are of this type.

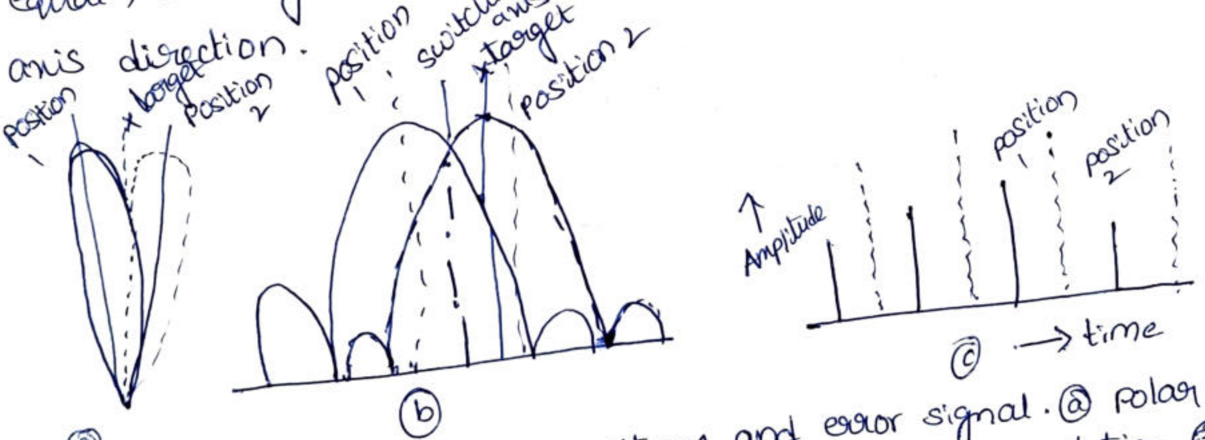
## sequential lobing:-

→ The antenna pattern commonly employed with tracking radars is the symmetrical pencil beam in which the elevation and azimuth beam widths are approximately equal. However, a simple pencil beam antenna is not suitable for tracking radars unless means are provided for determining the magnitude and the direction of the target's angular position with respect to some reference direction, usually the axis of the antenna.

→ the difference b/w 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 is one coordinate is by alternately switching the antenna beam b/w two positions. this is called lobe switching, squential switching or sequential lobing.

→ the difference in amplitude b/w 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 target. when the voltages in the two switched positions are equal, the target is on axis and its position may be determine from the axis direction.




Lobe switching antenna patterns and error signal. (a) polar representation of switching antenna patterns (b) rectangular representation (c) error signal.

→ 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.

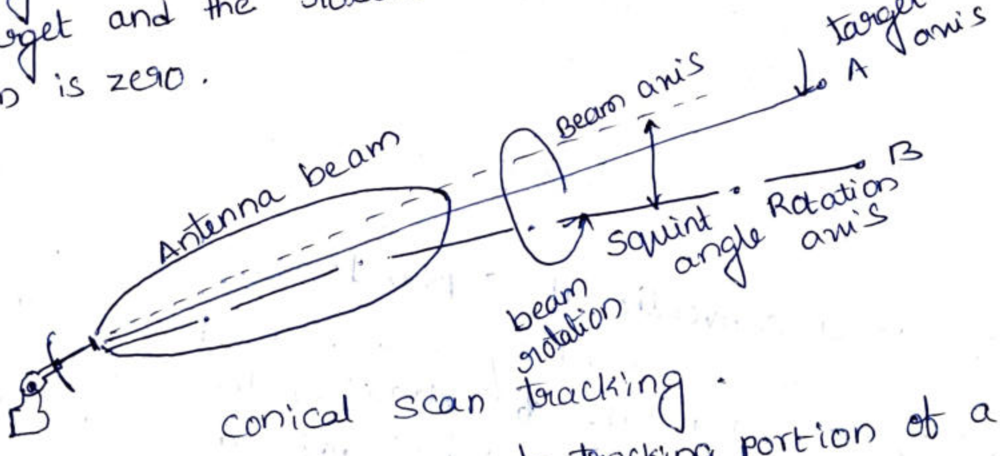
→ An important feature of sequential lobing is that the target position accuracy can be far better than that given by the antenna beamwidth. the accuracy depends on how well equality of the signals in the switched positions can be determined. the fundamental limitation to accuracy is system noise caused either by mechanical or electrical fluctuations. sequential lobing or lobe switching was one of the first tracking radar techniques to be employed.

Conical scan:-

→ A logical extension of the simultaneous lobing technique described in the previous section is to rotate continuously an offset antenna beam rather than discontinuously step the beam b/w four discrete positions. this is known as conical scanning. the angle b/w the axis of rotation and the axis of 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 angle b/w the target line of sight and the rotation axis.

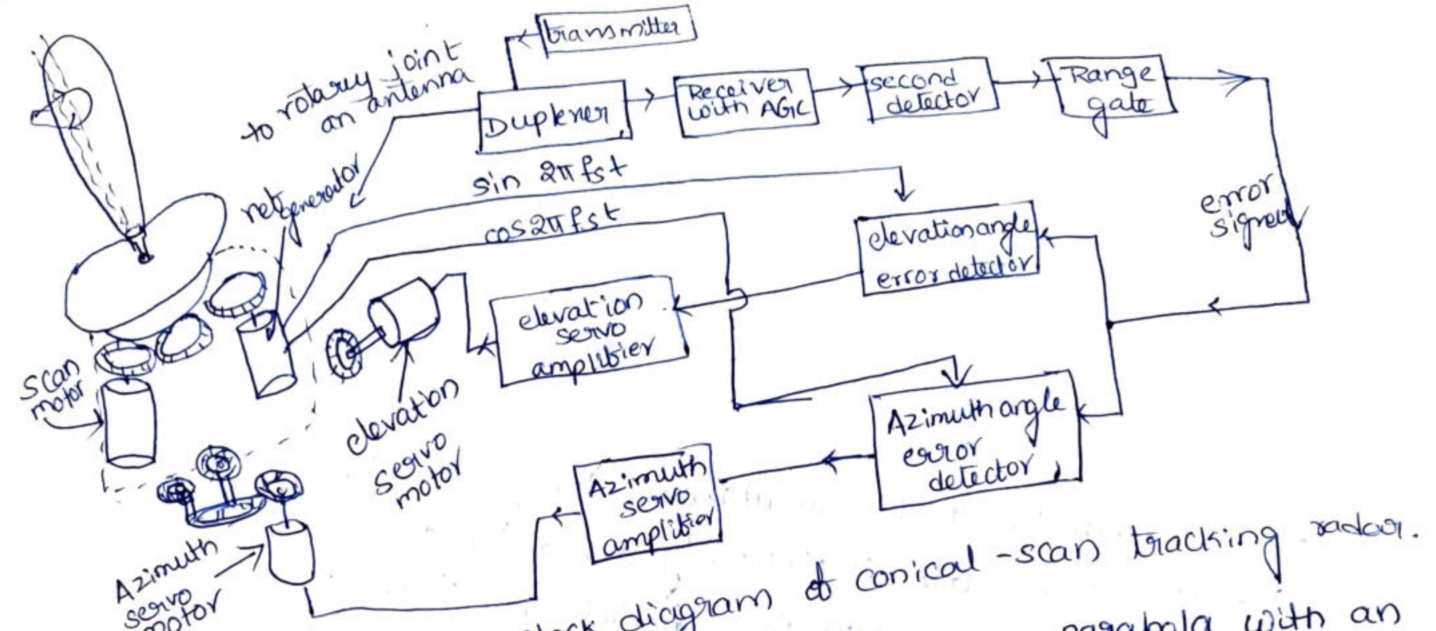
→ the conical scan modulation is extracted from the echo signal and applied to a servo-control system which continually positions the antenna on the target. when the antenna is on target, as in B the line of sight to the Target and the rotation axis coincide and the conical scan modulation is zero.




conical scan tracking.

→ The block diagram of the angle tracking portion of a typical conical scan tracking radar is shown in the fig. the antenna is mounted so that it can be positioned in both azimuth and elevation by seperate 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.

②




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. 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 rotatary 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 o/p's 90° apart in phase. these two o/p's 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. one rotary joint permits motion in azimuth; the other in elevation.

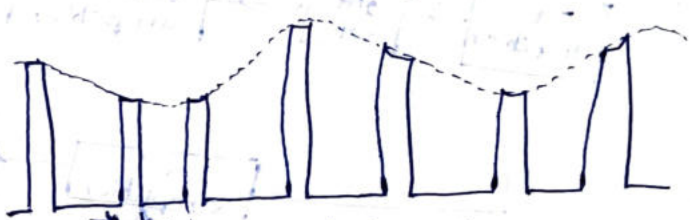
→ the receiver is a conventional superheterodyne except for features peculiar to the conical scan tracking radar. one feature not found in other radar receivers is a means of extracting the conical scan modulation or error signal. this is accomplished after the second detector in the video portion of 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 phase sensitive detector is a non-linear device in which the I/P signal is mixed with the reference signal.

→ The angular position of the target may be determined from the elevation and azimuth of the antenna axis.

Boxcar generator :- when extracting the modulation imposed on a repetitive train of narrow pulses, it is usually convenient to stretch the pulses before low-pass filtering. this is called boxcaring or sample and hold. here the device is called the boxcar generator. the boxcar generator consists of an electric cht that clamps the potential of a storage element, such as a capacitor, to the video-pulse amplitude each time the pulse is received.

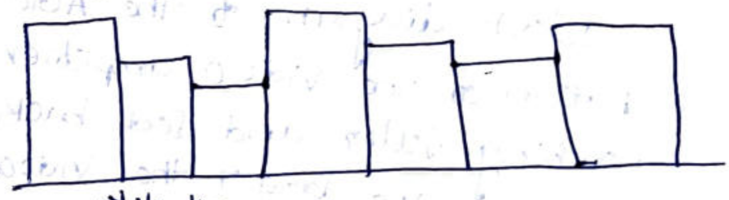
→ the capacitor maintains the potential of the pulse during the entire repetition period and is altered only when a new video pulse appears whose amplitude differs from the previous one. the boxcar generator eliminates the pulse repetition freq and reduces its harmonics.

→ It also has the practical advantage that the magnitude of the conical scan modulation is amplified because pulse stretching puts more of the available energy at the modulation freq.



pulse train with conical scan modulation

→ the pulse repetition freq must be sufficiently large compared with the conical scan freq for proper boxcar filtering. If not, it may be necessary to provide additional filtering to attenuate undesired cross modulation freq components.



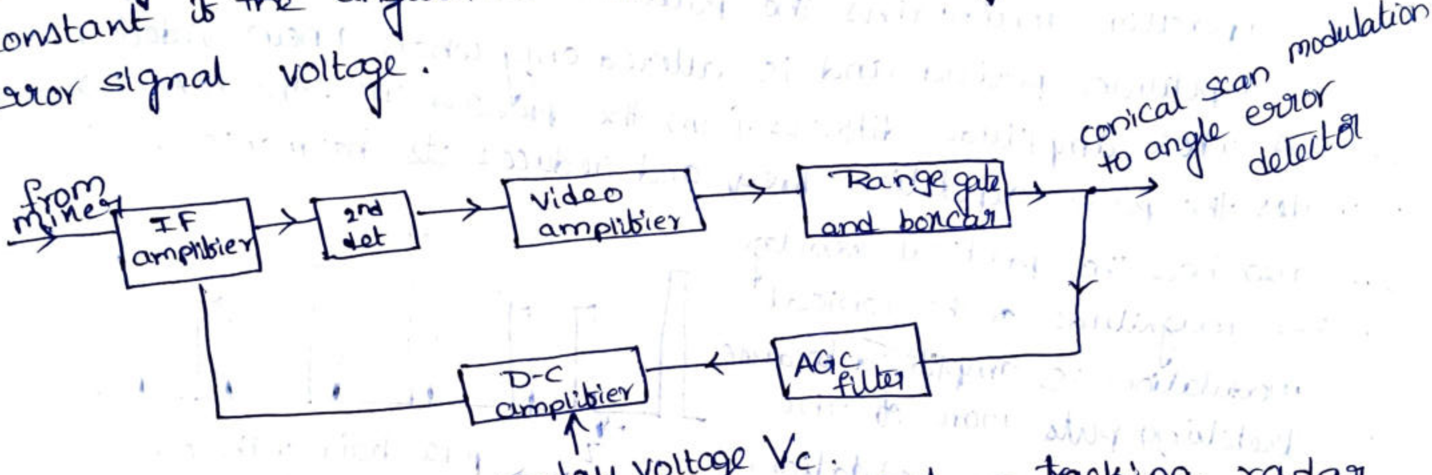
same pulse train after passing through boxcar generator.

Automatic gain control (AGC) :-

→ 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 b/w the echo signal and range. ② the conical-scan modulation and ③ Amplitude fluctuations in the target cross section.

→ the function of the automatic gain control (AGC) is to maintain the dc-level of the receiver o/p 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 freq.

→ one of the purposes of AGC in any receiver is to prevent saturation by large signals. the scanning modulation and the error signal would be lost if the receiver were to saturate. In the conical scan tracking radar an AGC that maintains the dc 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.




Block diagram of the AGC portion of a tracking radar receiver.

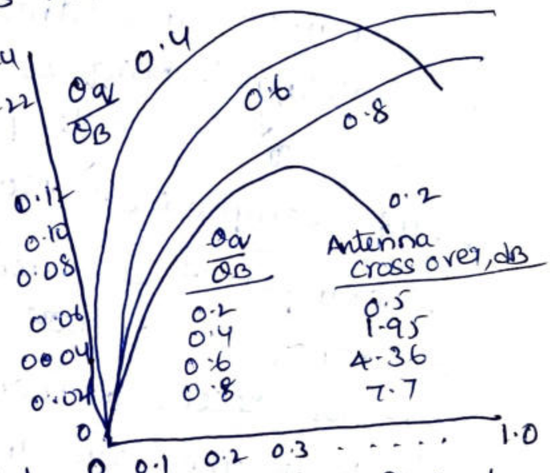
→ A portion of the video amplifier o/p is passed through a low pass or smoothing filter and fed back to control the gain of the IF amplifier. the larger the video o/p, the larger will be the feedback signal and the greater will be the gain reduction.

→ the filter in the AGC loop should pass all frequencies from direct current to just below the conical-scan modulation freq. the loop gain of the AGC filter measured at the conical-scan freq should be low so that the error signal will not be affected by AGC action.

→ A phase change of the error signal is equivalent to a rotation of the reference axes and introduces cross coupling or "cross talk", b/w the elevation and azimuth angle tracking loops. cross talk affects the stability of the tracking and might result in an unwanted nutating motion of the antenna.

→ 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 performance will not be as good.
→ An alternative AGC filter design would maintain the AGC loop gain up to frequencies much higher than the conical-scan freq.

<u>squint angle</u>:- the angle error signal voltage as a function of $\theta_T$, the angle b/w the axis of rotation and the direction to the target. the squint angle $\theta_q$ is the angle b/w the antenna beam axis and the axis of rotation. and $\theta_B$ is the half power beamwidth.




<u>other considerations</u> :-
→ In both the sequential-lobing and conical scan techniques, the measurement of the angle error in two orthogonal coordinates (azimuth and elevation) requires that a <u>min</u> of three pulses be processed.
→ A conical-scan-on-receive only (COSRO) tracking radar radiates a nonscanning transmit beam, but receives with a conical scanning beam to extract the angle error. the analogous operation with sequential lobing is called lobe-on-receive-only (LORO).

## <u>Monopulse Tracking</u> Radar:-

→ the conical-scan and sequential lobing tracking radars require a <u>min</u> number of pulses in order to extract the angle error signal. In the time interval during which a measurement is made with either sequential lobing or conical scan, the train of echo pulses must contain no amplitude modulation components other than the modulation produced by scanning.
→ If the echo pulse train did contain additional modulation components caused by a fluctuating target cross section, the tracking accuracy might be degraded, especially if the freq components of the fluctuations were at or near the conical scan freq or the sequential lobing rate.

→ pulse-to-pulse amplitude fluctuations of the echo signal have no effect on tracking accuracy if the angular measurement is made on the basis of one pulse rather than many. there are several methods by which angle error Information might be obtained with only a single pulse.

→ More than one antenna beam is used simultaneously in these methods, in contrast to the conical-scan & lobe switching tracker, which utilizes one antenna beam on a time shared basis.

→ the angle of arrival of the echo signal may be determined in a single pulse system by Measuring the relative phase & the relative amplitude of the echo pulse received in each beam.

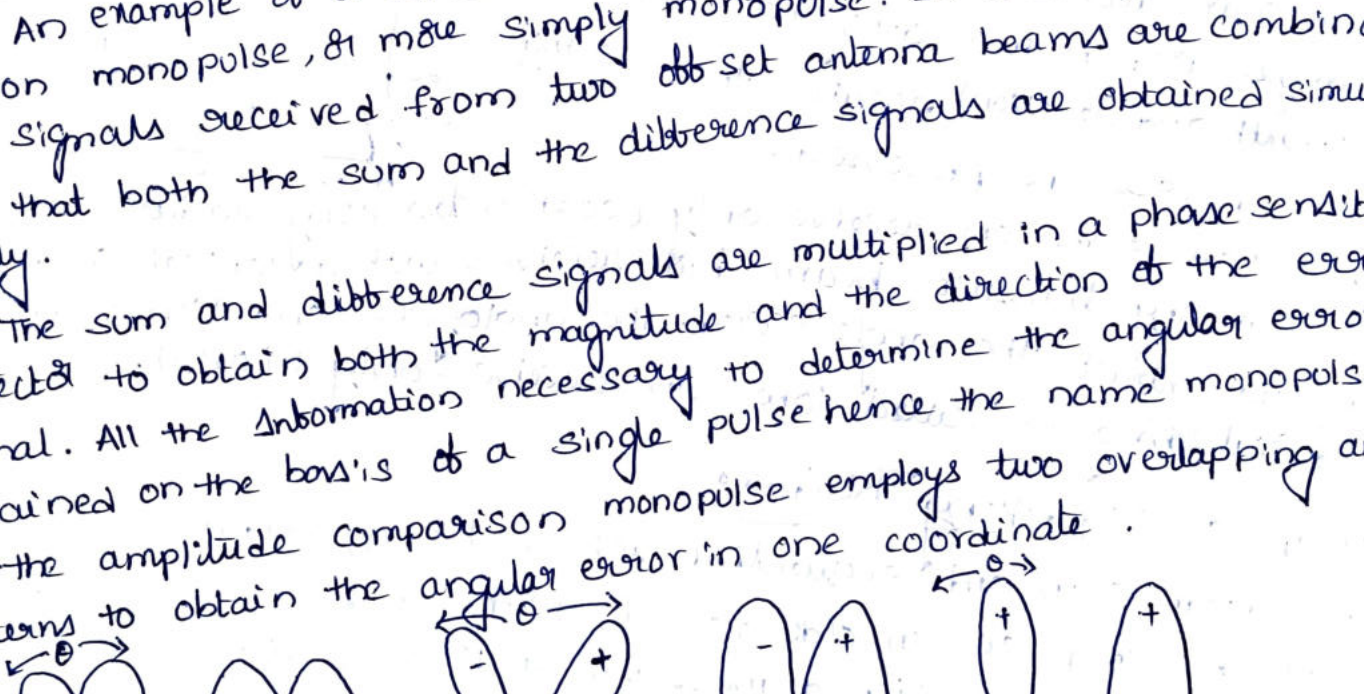
→ the names simultaneous lobing and monopulse are used to describe those tracking techniques which derive angle error Information on the basis of a single pulse.

## Amplitude comparison Monopulse :-

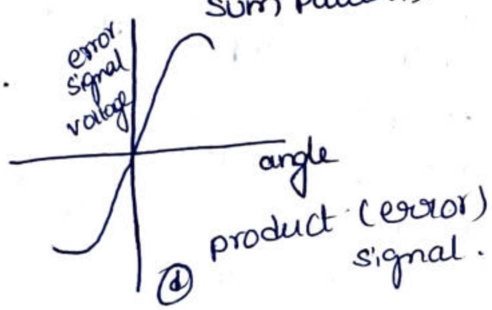
→ An example of a simultaneous-lobing technique is amplitude-comparison monopulse, & more simply monopulse. In this technique the RF signals received from two off set antenna beams are combined so that both the sum and the difference signals are obtained simultaneously.

→ The sum and difference signals are multiplied in a phase sensitive detector to obtain both the magnitude and the direction of the error signal. All the Information necessary to determine the angular error is obtained on the basis of a single pulse hence the name monopulse.

→ the amplitude comparison monopulse employs two overlapping antenna patterns to obtain the angular error in one coordinate.




Mono pulse antenna patterns and error signals. (a-c) are polar coordinates (b-d) are rectangular coordinates. (a) overlapping antenna patterns



→ the sum of the two antenna patterns are used for transmission, while both the sum pattern and the difference pattern are used on reception. the sum signal provides the range measurement and is also as a reference to extract 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. the signal received with the difference pattern provides the magnitude of the angle error.

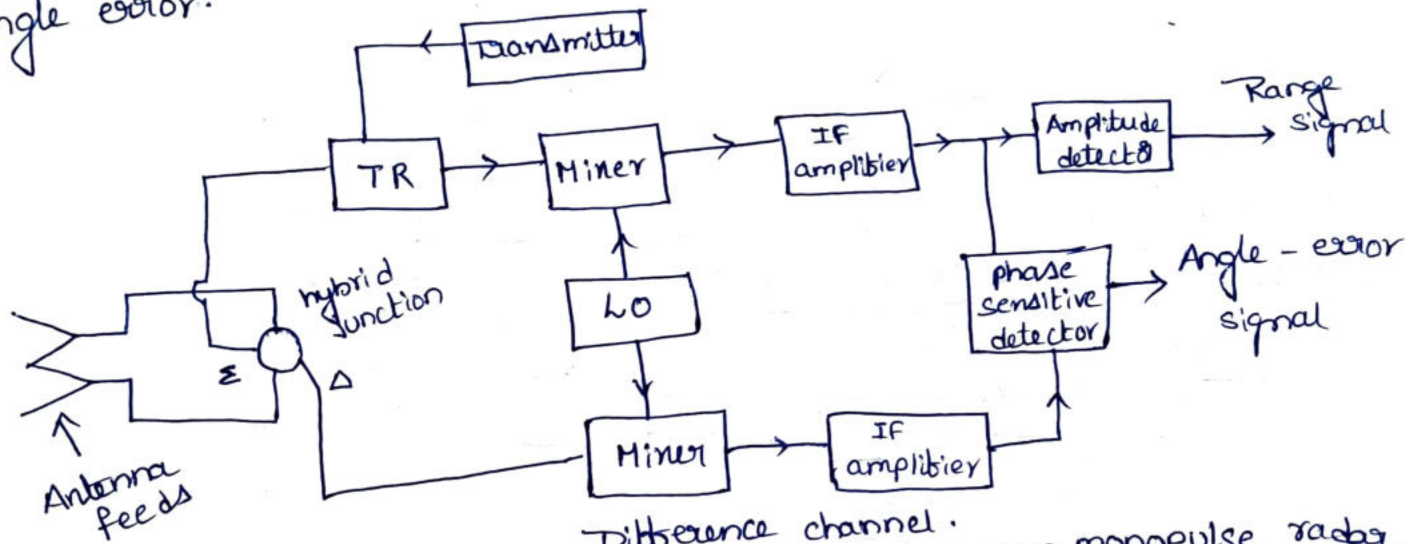



Fig: Block diagram of amplitude comparison monopulse radar.

→ 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 o/p's of the sum arm and the difference arm are each heterodyned to an Intermediate freq and amplified as in any superheterodyne receiver.

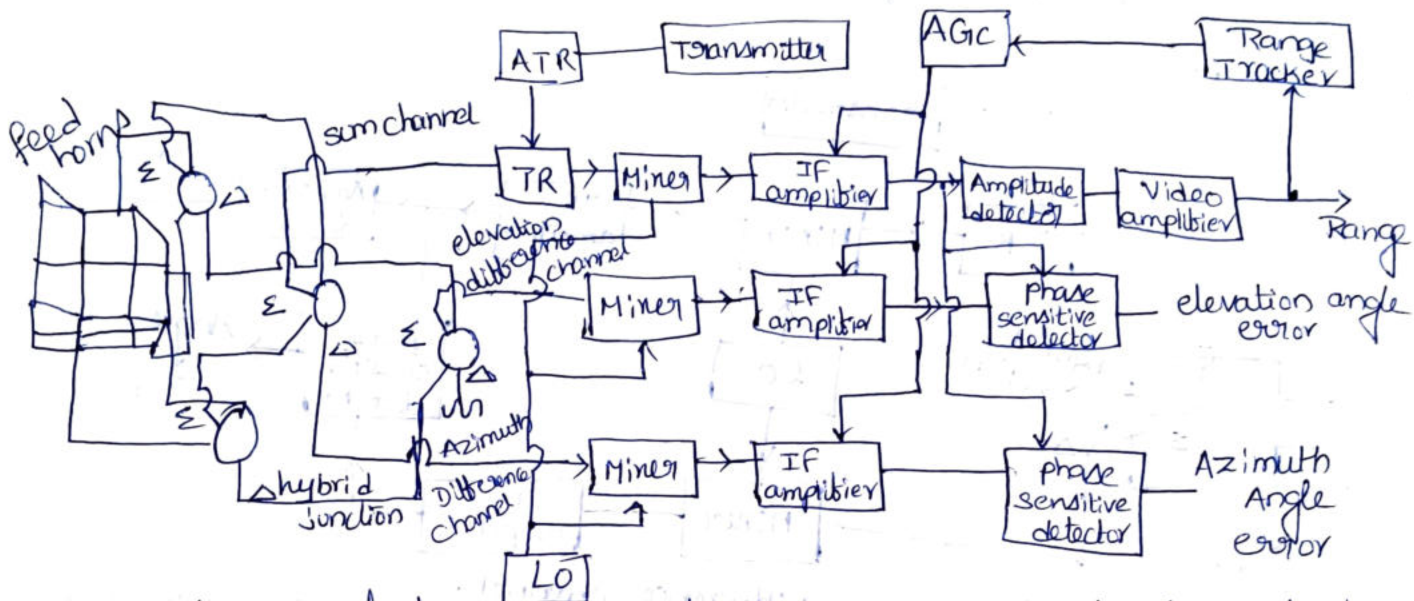
→ The transmitter is connected to the sum arm. Range Information is also extracted from the sum channel. A duplexer is included in the sum arm for the protection of the receiver. the o/p of the phase sensitive detector is an error signal whose magnitude is proportional to the angular error & whose sign is proportional to the direction.

→ the o/p of the monopulse radar is used to perform automatic tracking. the sign of the difference signal 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 $A_s \cos w_{IF} t$, the difference signal would be either $A_d \cos w_{IF}$ or $- A_d \cos w_{IF} t$ ($A_s > 0, A_d > 0$) 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 180° out of phase.

→ A block diagram of a monopulse radar with provision for extracting error signals in both elevation and azimuth is shown in the fig.




Block diagram of two - coordinates (azimuth and elevation) amplitude comparison monopulse tracking radar.

→ 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 seperate 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 b/w 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 o/p of the sum channel after amplitude detection.

→ Since a phase comparison is made b/w the o/p 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 b/w channels must be maintained to within 25° or better for resonably proper performance. the gains of the channels also must not differ by more than specified amounts.

→ the Monopulse antenna must generate a sum pattern with high efficiency and a difference pattern with a large value of slope at the cross over of the offset beams. the sidelobes of both the sum and the difference pattern must be low, the antenna must be capable of the desired bandwidth and the desired polarization characteristics. Antenna design patterns must have the successful realization of a good monopulse is an important part of the radar.

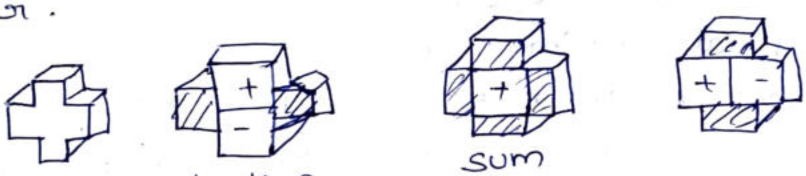



Fig: Approximately "Ideal" Difference feed aperture illumination for monopulse and difference channels.

## Phase comparison Monopulse :-

→ the sequential-lobing and conical-scan techniques used a single, time-shared antenna beam, while the monopulse technique used two or more simultaneous beams. the difference in amplitudes in the several antenna positions was proportional to the angular error. the angle of arrival (in one coordinate) may also be determined by comparing the phase difference b/w the signals from two separate antennas.

→ The amplitudes of the target echo signals are essentially the same from each antenna beam, but the phases are different. A tracking radar which operates with phase information is similar to an active interferometer and might be called an interferometer radar. It has also been called simultaneous-phase-comparison radar, or phase comparison monopulse.

→ the two antennas are shown seperated by a distance d. the distance to the target is R and is assumed large compared with the antenna seperation d. the line of sight to the target makes an angle θ to the perpendicular bisector of the line joining the two antenna. the distance from antenna 1 to the target is

$$R_1 = R + \frac{d}{2}\sin\theta.$$

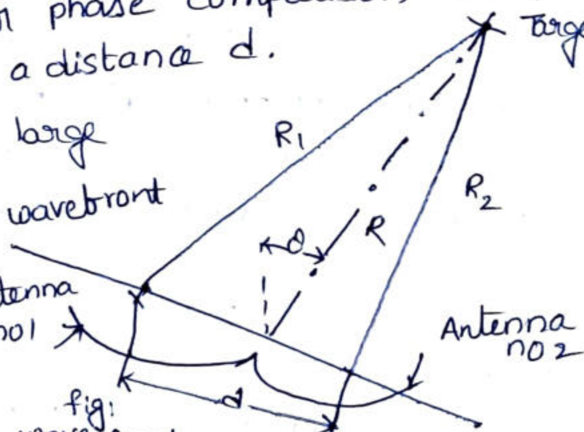



fig: wave front phase relationships in phase comparison monopulse radar.

and the distance from antenna 2 to the target is

$$R_2 = R - \frac{d}{2}\sin\theta .$$

The phase difference b/w the echo signals in the two antenna is approximately $\Delta\phi = \frac{2\pi}{\lambda} d \sin\theta$.

→ for small angles where $\sin\theta \simeq \theta$, the phase difference is a linear function of the angular error & may be used to position the antenna via a servo-control loop. the phase comparison monopulse radar, the angular error was determined by measuring the phase difference b/w the o/p's of receivers connected to each antenna. the o/p from one of the antennas was used for transmission and for providing the range information.

→ the phase and amplitude comparison principles can be combined in a single radar to produce two dimensional angle tracking with only two, rather than four, antenna beams. the angle information in one plane (the azimuth) is obtained by two seperate antennas placed side by side as in a phase comparison monopulse. one of the beam is tilted slightly upward, while the other is tilted slightly downward, to achieve the squint needed for amplitude comparison monopulse in elevation.

→ the horizontal projection of the antenna patterns is that of a phase comparison system, while the vertical projection is that of an amplitude comparison system. both of them employ two antenna beams. the meas-urements made by the two systems are not the same. consequently, the characteristics of the antenna beams will also be different.

→ In the amplitude-comparison monopulse the two beams are offset, i.e. the point in slightly different directions this type of pattern may be generated by using one reflector dish with two feed horns side by side. (four feed horns for two coordinate data). Since the feeds may be placed side by side, they could be as close as one-half wavelength.

→ With such close spacing the phase difference b/w the signals received in the two feeds is negligibly small. Any difference in the amplitudes b/w the two antenna o/p's in the amplitude comparison monopulse system is a result of differences in Amplitude and not phase.

→ the phase-comparison monopulse, on the other hand, measures the phase differences only and is not concerned with Amplitude difference. therefore the antenna beams are not offset, but are directed to illuminate a common volume in space. seperate antennas are needed since it is difficult to illuminate a single reflector with more than one feed and produce independent antenna patterns which illuminate the same volume in space.

## Tracking In Range :-

→ In most tracking - radar applications the target is continuously tracked
in range as well as in angle. Range tracking might be accomplished by an
operator who watches an A-scope & J-scope presentation and manually
positions a handwheel ~~in a measure~~ order to maintain a marker over the
desired target pip. the setting of the handwheel is a measure of the target
range and may be converted to voltage that is supplied to a data processor
→ As target speed Increase, it is Increasingly difficult for an operator
to perform at the necessary levels of efficiency over a sustained Period
of time, & automatic tracking becomes a necessity.
→ the technique for automatically tracking in range is based on the
split range gate. two range gates are generated as shown in the fig.
one is the early gate, the other is the late gate. the portion of the
signal energy contained in the early gate is less than that in the late
gate. If the O/P's 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 error signal is a measure of the difference b/w the
center of the pulse & the center of gates. the sign of the error signal
determines the direction in which the gates must be repositioned by a
feed-back 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 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.

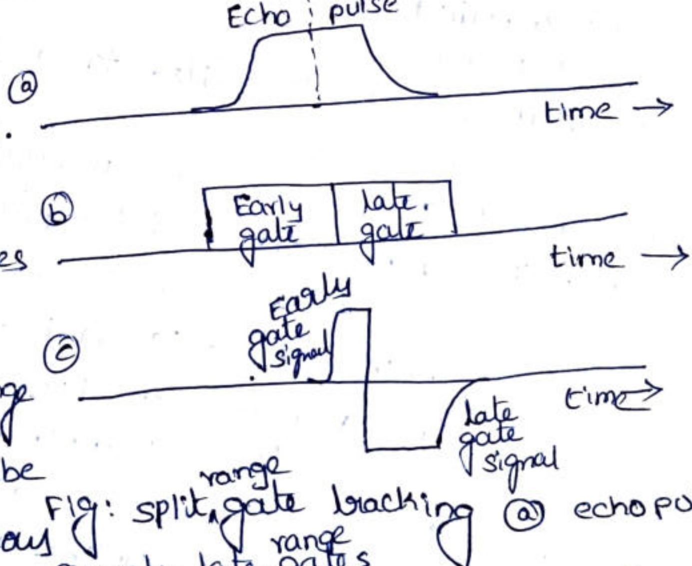
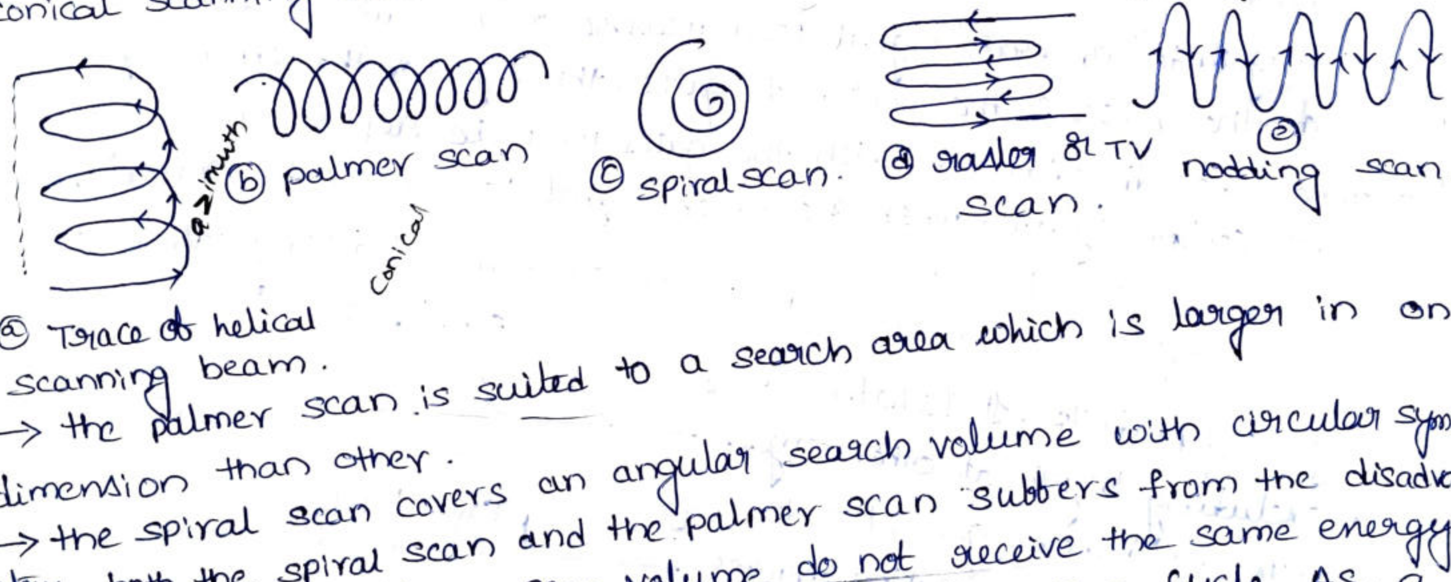



Fig: split range gate tracking ⓐ echo pulse.
ⓑ early late range gates
ⓒ difference signal b/w early & late range gates.

## Acquisition :-

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
→ In the helical scan, the antenna is continuously rotated in azimuth
while it is simultaneously raised & lowered in elevation. It traces a helix
in space. the SCR-584 antenna was rotated at the rate of 6 rpm and
covered a 20° elevation angle in 1min. the palmer scan 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. bcoz of this property, the palmer scan is sometimes
used with conical-scan tracking radars which must operate with a
search as well as track mode since the same mechanism used to produce
conical scanning can also be used for palmer scanning.




(a) Trace of helical scanning beam.
(b) palmer scan
(c) spiral scan.
(d) raster & TV scan.
(e) nodding scan.

→ the palmer scan is suited to a search area which is larger in one
dimension than other.
→ the spiral scan covers an angular search volume with circular symm-
etry. both the spiral scan and the palmer scan suffers from the disadvan-
tage 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
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 & 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, rectangle in shape.
→ similar to the raster scan is the nodding scan produced by
oscillating the antenna beam rapidly in elevation and slowly in
azimuth.

→ the raster scan is sometimes called a n-bar scan, where n is the no of horizontal rown. Although it may be used employed to cover a limited sector - as does the raster scan nodding scan may also be used to obtain hemispherical coverage, ie, elevation angle extending to 90° and azimuth scan angle to 360°.

→ 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.

## Comparison of Trackers :-

→ of the four continuous tracking radar techniques conical scan and Amplitude comparison monopulse have been seen more application than the other two. the phase comparison monopulse has not been too popular bcoz of the relative awkwardness of its antenna. and bcoz the sidelobe levels might be higher than desired.

→ Although sequential lobing is similar to conical scan, the latter is preferred in most applications, since it suffers less loss and the antenna and feed systems are usually less complex.

→ when the target is being tracked, the signal-to-noise ratio available from the monopulse radar is greater than that of a conical-scan radar, all other things being equal, since the monopulse radar views the target at the peak of its sum pattern while the conical scan radar views the target at an angle off the peak of the antenna beam. for the same size aperture, the beamwidth of a conical-scan radar will be slightly greater than that of the monopulse bcoz its feed is offset from the focus.

→ the tracking accuracy of a monopulse radar is superior to that of the conical-scan radar bcoz of the absence of target amplitude fluctuations & bcoz of its greater signal to noise ratio. It is the preferred technique for precision tracking.

→ the monopulse radar is the more complex of the two. three separate receivers are necessary to derive the error signal in two orthogonal angular coordinates. only one receiver is needed in the conical scan radar.

→ since the monopulse radar compares the amplitudes of signals received in three separate channels, it is Important that the gain & phase shift through these channels be Identical.

→ A popular form of antenna for monopulse is the cassegrain. with the monopulse tracker it is possible to obtain a measure of the angular error in two coordinates on the basis of a single pulse. A minimum of four pulses are usually necessary with the conical scan radar.

→ The monopulse radar first makes its angle measurement and then integrates a no of pulses to obtain the required signal-to noise ratio & to smooth the error. the conical-scan radar, on the other hand, integrates the no of pulses first and then extracts the angle measurement.

→ bcoz a monopulse radar is not degraded by amplitude fluctuations it is less susceptible to hostile electronic countermeasures than in conical scan.

→ In brief, the monopulse radar is the better tracking technique but in Many applications where the ultimate in performance is not needed, the conical-scan radar is used bcoz it is less costly & less complex.

# UNIT - V

## Detection of Radar signals in Noise

### Introduction:-

→ the two basic operations performed by radar are ① detection of the presence of reflecting objects & ② extraction of information from the received waveform to obtain such target data as position, velocity & perhaps size.

→ the operation of detection & extraction may be performed separately & in either order, although a radar that is a good detection device is usually a good radar for extracting information, & vice versa. Now we are considering the problem of detecting radar signals in the presence of Noise.

→ Noise ultimately limits the capability of any radar. the problem of extracting information from the received waveform.

### Matched - filter Receiver:-

→ A n/w whose freq - response function maximizes the O/P 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 freq response function, denoted H(f), expresses the relative amplitude & phase of the O/P of a n/w with respect to the I/P when the I/P is a pure sinusoid. the Magnitude |H(f)| of the freq response function is the receiver amplitude passband characteristic.

→ If the B.W of the rxer passband is wide compared with that occupied by the signal energy, extraneous noise is introduced by the excess B.W which lowers the O/P signal to noise ratio. on the other hand, if the receiver bandwidth is narrower than the B.W 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 B.W at which the signal to noise ratio is a max. this is well known to the radar receivers 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.

→ the receiver freq response function is assumed to apply from the antenna terminals to the O/P of the IF amplifier. the second detector & video portion of the well designed radar superhetrodyne receiver will have negligible effect on the O/P signal to noise ratio if the receiver is designed as a matched filter. Narrowbanding is most conveniently accomplished in the IF.

→ the bandwidths of the RF & mixer stages of the normal superhetrodyne receiver are usually large compared with the IF b.w. therefore the freq response function of the portion of the rxer Included b/w the antenna terminals to the O/P of the IF amplifier is taken to be that of the IF amplifier alone. the IF Amplifier may be considered as a filter with gain.

→ for a received waveform s(t) with a given ratio of signal energy E to noise energy No showed that the freq- response function of the linear, time-invariant filter which maximizes the O/P peak signal to mean noise (power) ratio for a fixed I/P signal to noise (energy) ratio is

$$H(f) = G_a S^*(f) \exp(-j2\pi f t_1) \rightarrow (1)$$

where $S(f) = \int_{-\alpha}^{\alpha} s(t) \exp(-j2\pi f t)\, dt$ = voltage spectrum (fourier transform) of I/P signal.

$S^*(f)$ = complex conjugate of $S(f)$.

$t_1$ = fixed value of time at which signal is observed to be max

$G_a$ = constant equal to max filter gain (generally taken to be unity).

→ the Noise that accompanies the signal is assumed to be stationary & to have a uniform spectrum (white noise). It need not be gaussian. the filter whose freq response function is given by eq (1) have been called the North filter, the conjugate filter, & more usually the matched filter. It has also been called the Fourier transform criterion.

→ The freq spectrum of the received signal may be written as an amplitude spectrum $|S(f)|$ & a phase spectrum $\exp[-j\phi_s(f)]$. the matched filter freq response function may similarly be written in terms of its amplitude and phase spectra $|H(f)|$ & $\exp[-j\phi_m(f)]$. Ignoring the constant $G_a$.

$$|H(f)| \exp[-j\phi_m(f)] = |S(f)| \exp\{j[\phi_s(f) - 2\pi f_1]\} \rightarrow (2)$$

$$|H(f)| = |S(f)| \quad \text{and} \quad \phi_m(f) = -\phi_s(f) + 2\pi f_1 \rightarrow (3)$$

(or)

→ 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 -ve of the phase spectrum of the signal plus a phase shift proportional to freq.

→ the matched filter may also be specified by its Impulse response h(t), which is the Inverse fourier transform of the freq response function.

$$h(t) = \int_{-\alpha}^{\alpha} H(f) \exp(j2\pi f t)\, df \rightarrow ④$$

→ physically, the Impulse response is the o/p of the filter as a function of time when the i/p is an Impulse (delta function). substituting eq ① in eq ④ gives

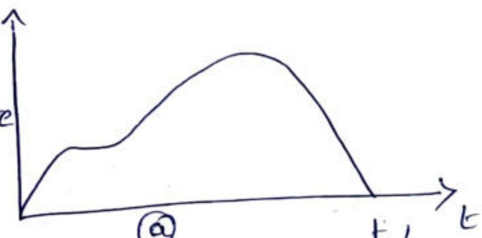
$$h(t) = G_a \int_{-\alpha}^{\alpha} S^*(f) \exp[-j2\pi f(t_1 - t)]\, df \rightarrow ⑤$$

since $S^*(f) = S(-f)$ we have

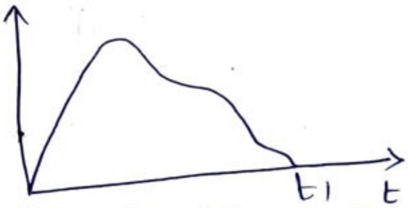
$$h(t) = G_a \int_{-\alpha}^{\alpha} S(f) \exp[j2\pi f(t_1 - t)]\, df = G_a\, s(t_1 - t) \rightarrow ⑥$$

→ A rather Interesting result is that the Impulse response of the matched filter is the Image of the received waveform. ie., it is same as the received signal run backward in time starting from the fixed time $t_1$.

→ the Impulse response of the filter, if it is to be realizable, is not defined for $t<0$. 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 $\exp(-j2\pi f t_1)$.




ⓐ Received waveform




ⓑ Impulse response h(t) of the mated filter.

## Derivation of the Matched filter characteristics:-

→ we shall derive the matched filter freq response function using the schwartz inequality. the freq response function of the linear, time invariant filter which maximizes the o/p peak-signal to mean noise ratio is

$$H(f) = G_a\, S^*(f) \exp(-j2\pi f t_1) \rightarrow ①$$

→ when the i/p noise is stationary & white (uniform spectral density). the ratio to maximize is $R_f = \frac{|s_o(t)|^2_{max}}{N} \rightarrow ②$

where $|s_o(t)|_{max}$ = max value of o/p signal voltage &

N = mean noise power at receiver o/p.

→ the o/p voltage of a filter with freq response function H(f) is

$$|s_o(t)| = \left| \int_{-\alpha}^{\alpha} S(f) H(f) \exp(j2\pi f t)\, df \right| \rightarrow ③$$

where S(f) is the fourier transform of the i/p (received) signal. the mean o/p noise power is

$$N = \frac{N_0}{2} \int_{-\alpha}^{\alpha} |H(f)|^2\, df \rightarrow ④$$

where No is the i/p noise power per unit bandwidth. the factor $\frac{1}{2}$ appears before the integral bcoz the limits extend from $-\alpha$ to $+\alpha$. whereas No is defined as the noise power per cycle of bandwidth over positive values only. sub eq ③ & ④ in eq ② & assuming that the max value of $|s_0(t)|^2$ occurs at time $t = t_1$ & ratio $R_f$ becomes

$$R_f = \frac{\left| \int_{-\alpha}^{\alpha} S(f)\, H(f)\, \exp(j2\pi f t_1)\, df \right|^2}{\frac{N_0}{2} \int_{-\alpha}^{\alpha} |H(f)|^2\, df} \rightarrow ⑤$$

Schwartz's inequality states that if P & Q are two complex functions, then

$$\int P^* P\, dx = \int Q^* Q\, dx \geq \left| \int P^* Q\, dx \right|^2 \rightarrow ⑥$$ the equality sign applies

when $P = KQ$ where K is a constant. letting

$$P^* = S(f) \exp(j2\pi f t_1) \text{ \& } Q = H(f)$$ & recalling that

$\int P^* P\, dx = \int |P|^2\, dx$. we get on applying the schwartz inequality to the numerator of eq ⑤

$$R_f \leq \frac{\int_{-\alpha}^{\alpha} |H(f)|^2\, df \int_{-\alpha}^{\alpha} |S(f)|^2\, df}{\frac{N_0}{2} \int_{-\alpha}^{\alpha} |H(f)|^2\, df} = \frac{\int_{-\alpha}^{\alpha} |S(f)|^2\, df}{\frac{N_0}{2}} \rightarrow ⑦$$

from parseval's theorem,

$$\int_{-\alpha}^{\alpha} |S(f)|^2\, df = \int_{-\alpha}^{\alpha} s^2(t)\, dt = \text{signal energy} = E. \rightarrow ⑧$$

$\therefore$ we have $R_f \leq \frac{2E}{N_0} \rightarrow ⑨$ which maximizes the peak signal-to noise ratio

→ the freq response function which maximizes the peak signal-to noise ratio $R_f$ may be obtained by noting that the equality sign in eq ⑥ applies when $P = KQ$ or $H(f) = G_a S^*(f) \exp(-j2\pi f t_1) \longrightarrow ⑩$ where the constant K has been set equal to $1/G_a$.

→ The interesting property of the matched filter is that no matter what the shape of the i/p signal waveform, the max ratio of the peak signal power to the mean noise power is simply twice the energy E contained in the signal divided by the noise power per hertz of the bandwidth No. No is equal to $KT_0F$ where K is the boltzmann constant $T_0$ is the standard temperature (290 K) and F is the Noise figure.

**The Matched Filter & the correlation function :-** The O/P of the Matched filter is not a replica of the I/P 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 I/P pulse rather than maximize the O/P signal to noise ratio some other criterion must be employed.

→ the O/P of the matched filter may be shown to be proportional to the I/P signal cross correlated with a replica of the transmitted signal, except for the time delay $t_1$. the cross correlation function $R(t)$ of two signals $y(\lambda)$ & $s(\lambda)$ each of finite duration is defined as

$$R(t) = \int_{-\alpha}^{\alpha} y(\lambda)\, s(\lambda - t)\, d\lambda \rightarrow ①$$

→ the O/P $y_o(t)$ of a filter with Impulse response $h(t)$ when the I/P is $y_{in}(t) = s(t) + n(t)$ is $y_o(t) = \int_{-\alpha}^{\alpha} y_{in}(\lambda)\, h(t-\lambda)\, d\lambda \rightarrow ②$

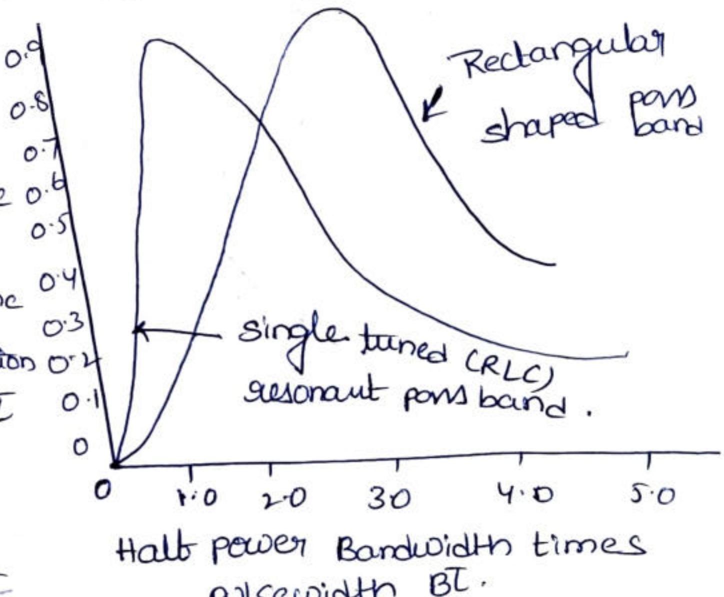
→ If the filter is a matched filter then $h(\lambda) = s(t_1 - \lambda)$ & eqn ② becomes

$$y_o(t) = \int_{-\alpha}^{\alpha} y_{in}(\lambda)\, s(t_1 - t + \lambda)\, d\lambda = R(t - t_1).$$

→ thus the matched filter forms the cross correlation b/w the received signal corrupted by noise & a replica of the transmitted signal.

→ the replica of the transmitted signal is "built in" to the matched filter via the freq response function.

→ If the I/P signal $y_{in}(t)$ were the same as the signal $s(t)$ for which the matched filter was designed, the O/P would be the autocorrelation function. the autocorrelation function of a rectangular pulse of width $\tau$ is a triangle whose base width $2\tau$.




**Efficiency of Nonmatched filters :-**

→ In practice the matched filter cannot always be obtained exactly. the measure of efficiency is taken as the peak signal-to noise ratio from the nonmatched filter divided by the peak signal to noise ratio ($2E/N_0$) from the matched filter.

## Efficiency of nonmatched filters compared with the matched filter.

| I/P signal | filter | optimum BT | 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 ckt | 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 |

## Matched filter with non-white Noise:-

In the derivation of the matched filter characteristic, the spectrum of the noise accompanying the signal was assumed to be white. ie, it was Independent of freq. If this assumption were not true, the filter which max the O/P signal to noise ratio would not be the same as the matched filter.

→ If the I/P power spectrum of the Interfering noise is given by $[N_i(f)]^2$, the freq response function of the filter which maximizes the O/P signal-to noise ratio is

$$H(f) = \frac{G_a S^*(f) \exp(-j2\pi f t_1)}{[N_i(f)]^2} \rightarrow ①$$

→ when the noise is nonwhite, the filter which maximizes the O/P signal to noise ratio is called the NWN (nonwhite noise) matched filter. for white noise $[N_i(f)]^2$ = constant and the NWN matched filter freq response of eqn ① can be written as

$$H(f) = \frac{1}{N_i(f)} \times G_a \left(\frac{S(f)}{N_i(f)}\right)^* \exp(-j2\pi f t_1) \rightarrow ②$$

→ this Indicates that the NWN matched filter can be considered as the cascade of two filters. the first filter, with freq response function $1/N_i(f)$ acts ~~as~~ to Make the noise spectrum uniform, or white. It is sometimes called the whitening ~~Noise~~ filter. the seccond is the matched filter when the I/P is white Noise & a signal whose spectrum is $S(f)/N_i(f)$.

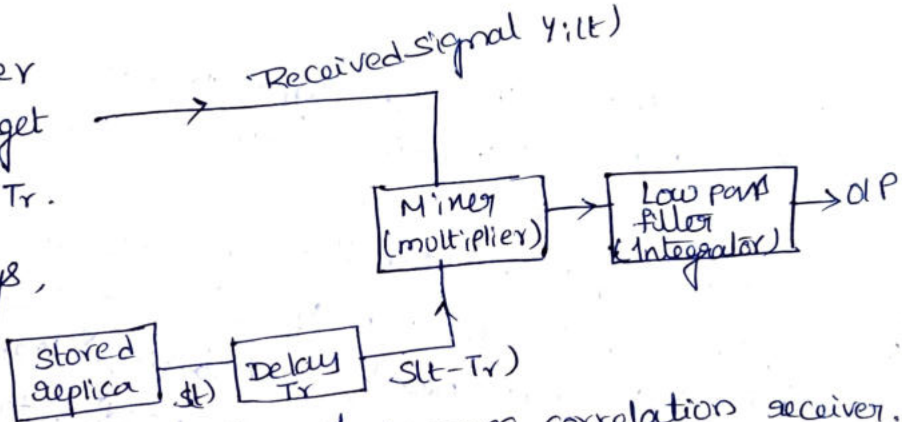
## Correlation Detection :-

→ the O/P of the matched filter as the cross correlation b/w the I/P signal & 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.

→ the I/P signal $y(t)$ is multiplied by a delayed replica of the transmitted signal $s(t-T_r)$, and the product is passed through a lowpass filter to perform the Integration.

→ the cross correlation receiver tests for the presence of a target at only a single time delay $T_r$.

→ Targets at other time delays, & ranges might be found by varying $T_r$.




Block diagram of a cross correlation receiver.

→ 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 $T_r$, as well as a multiplier and low pass filter.

→ In some applications it may be possible to record the signal on some storage Medium, & at a higher playback speed perform the search sequentially with different values of $T_r$, i.e., the playback speed is Increased in proportion to the no of time delay Intervals $T_r$ that are to be tested.

## Radar displays & Duplexers :-

**The Radar Receiver :-** The function of the radar receiver is to detect desired echo signal in the presence of noise, Interference, & clutter. It must separate wanted from unwanted signals, & 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, & clutter echoes with which the desired echo signals must compete.

→ here we are considering the receiver design is considered mainly
as a problem of extracting desired signal from noise. 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.
→ Good receiver design is based on maximizing the O/P signal-to noise
ratio. to maximize the o/p signal-to-noise ratio, the receiver must be
designed as a matched filter or its equivalent.
→ Receiver design also must be concerned with achieving sufficient gain,
phase and amplitude stability, dynamic range, tuning, ruggedness &
simplicity. protection must be provided against overload or saturation,
and burnout from nearby interfering transmitters.
→ Timing & reference signals are needed to properly extract target Infor-
mation. specific applications such as MTI radar, tracking radar, or radar
designed to minimize clutter place special demands on the receivers.
→ Although the superregenerative, crystal video & tuned radio freq
(TRF) receivers have been employed in radar systems, the superhetero-
dyne has seen almost exclusive applications bcoz of its good sensitivity
high gain, selectivity & reliability. NO other receiver type has been
competitive to the superheterodyne.

<u>Noise Figure</u> :-
→ 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 n/w may be defined as

$$F_n = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{N_{out}}{KT_0B_nG}$$

where $S_{in}$ = available I/P signal power.
$N_{in}$ = available I/P noise power (equal to $KT_0B_n$)
$S_{out}$ = available O/P signal power.
$N_{out}$ = available O/P noise 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 = $1.38 \times 10^{-23}$ J/deg. $T_0$ = standard temperature of
290 K (room temp) & $B_n$ is the noise Bandwidth. the product
$KT_0 \simeq 4 \times 10^{-21}$ W/Hz.

→ It may be considered as the degradation of the signal-to noise ratio caused by the n/w (receiver), or it may be Interpreted as the ratio of the actual available o/p noise power to the noise power which would be available if the n/w merely Amplified the thermal noise. the noise figure may also be written

$$F_n = \frac{KT_0 B_n G + \Delta N}{KT_0 B_n G} = 1 + \frac{\Delta N}{KT_0 B_n G}.$$

→ where $\Delta N$ is the additional noise Introduced by the n/w itself. the noise figure is commonly expressed in decibels, ie., $10\log F_n$.

→ The definition of noise figure assumes the I/P & O/P of the n/w are matched. In some devices, less noise is obtained under mismatched, rather than matched conditions. Inspite of definitions, such n/w's would be operated so as to achieve the max signal-to-noise ratio.

Noise figure of n/w's in cascade :-

→ consider two n/w's in cascade, each with the same noise bandwidth $B_n$, but with different noise figures and available gain. let $F_1, G_1$ be the noise figure & available gain, respectively, of the first n/w & $F_2, G_2$ be similar parameters for the second n/w.

→ the problem is to find $F_0$, the overall noise figure of the two ckts in cascade. from the defination of noise figure the O/P noise $N_0$ of the two ckts in cascade is

$N_0 = F_0 G_1 G_2 KT_0 B_n$ = noise from n/w 1 at to I/P of n/w 2 + Noise $\Delta N_2$ Introduced by n/w 2.

$N_0 = KT_0 B_n F_1 G_1 G_2 + \Delta N_2 = KT_0 B_n F_1 G_1 G_2 + (F_2 - 1) KT_0 B_n G_2$. or

$$F_0 = F_1 + \frac{F_2 - 1}{G_1}.$$

→ the contribution of the second n/w to the overall noise figure may be made negligible if the gain of the first n/w 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 n/w is not an Amplifier but is a n/w with loss (as in a crystal mixer), the gain $G_1$ should be Interpreted as a number less than unity.

The Noise figure of N n/w's in cascade may be shown to be

$$F_0 = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots + \frac{F_N - 1}{G_1 G_2 \cdots G_{N-1}}.$$

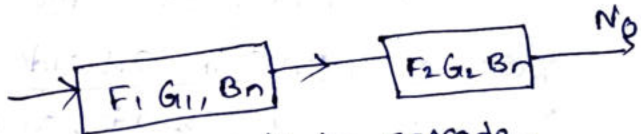
Noise Temperature :- The noise introduced by a n/w may also be expressed as an effective noise temperature, $T_e$, defined as that (fictional) temperature at the I/P of the n/w which would account for the noise $\Delta N$ at the O/P.

$$\therefore \Delta N = k T_e B_n G_1 \text{ \& } F_n = 1 + \frac{T_e}{T_0}, \quad T_e = (F_n - 1) T_0.$$

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. Two n/w's in cascade.

→ The effective noise temperature of a receiver consisting of a no of n/w's in cascade is

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

where $T_i$ & $G_i$ are the effective noise temperatures & gain of the $i^{th}$ n/w.

→ the effective noise temperature & the noise figure both describe the same characteristics of a n/w. In general, the effective noise temperature has been preferred for describing low noise devices, & the noise figure is preferred for conventional receivers. for radar receivers the noise figure is the more widely used term.

Radar Displays :-

→ The purpose of the displays is to visually present in a form suitable for operator interpretation & action the information contained in the radar echo signal. when the display is connected directly to the video O/P of the receiver, the information displayed is called raw video. this is the "traditional" type of radar presentation.

→ when the receiver video O/P is first processed by an automatic detector or automatic detection & tracking processor (ADT), the O/P displayed is sometimes called synthetic video.

→ The cathode ray tube (CRT) has been almost universally used as a radar display. there are two basic cathode-ray tube displays. one is the deflection-modulated CRT, such as the A-scope, in which a target is indicated by the deflection of the electron beam.

→ the other is Intensity Modulated CRT, such as the PPI, in which a target is Indicated by Intensifying the electron beam & presenting a luminous spot on the face of the CRT. In general, deflection-modulated CRT; have the advantage of simpler ckts than those of Intensity modulated displays & the targets may be more readily discerned in the presence of noise or Interference.

→ on the other hand, Intensity Modulated displays have the advantage of presenting data in a convenient & easily Interpreted form. the deflection of the beam or the target is commonly referred to a blip.

→ the focusing & deflection of the electron beam may be accomplished electrostatically, electromagnetically or by a combination of the two. Deflection ~~of the beam or the~~ Modulated CRT's such as the A-scope, generally employ electrostatic deflection. Intensity-Modulated CRT's such as the PPI, generally employ electromagnetic deflection.

→ the ability of an operator to extract Information efficiently from a CRT display will depends on such factors as the brightness of the display, density & character of the back ground noise, pulse repetition rate, scan rate of the antenna beam, signal clipping, decay time of phosphor, length of time of blip exposure, blip size, viewing distance, ambient illumination, dark adaptation, display size & operator fatigue.

<u>Types of Display presentations</u> :- The various types of CRT displays which might be used for surveillance & tracking radars are defined as follows.

<u>A-scope</u>:- A deflection Modulated display in which the vertical deflection is proportional to <u>target echo</u> strength & the horizontal coordinates is proportional to <u>range</u>.

<u>B-scope</u>:- A Intensity Modulated rectangular display with <u>azimuth angle</u> Indicated by the horizontal coordinates & range by the vertical coordinates.

C-Scope :- An Intensity Modulated Rectangular display with azimuth angle
Indicated by the horizontal coordinate & elevation angle by the vertical
coordinates.
D-Scope :- A C-scope in which blips extend vertically to give a rough
estimate of distance.
E-Scope :- An Intensity Modulated rectangular display with distance Indicated
by the horizontal coordinate & elevation angle by the vertical coordinates.
Similar to the RHI in which target height or altitude is the vertical coor-
dinates.
F-Scope :- A rectangular display in which a target appears as a centralized
blip when the radar antenna is aimed at it. horizontal & vertical aiming
errors are respectively Indicated by the horizontal & 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 &
wings appear to grow on the pip as the distance to the target is diminished.
H-Scope :- A B-scope modified to Include Indication of angle of elevation
A target appears as two closely spaced blips which approximate a short
bright line, the slope of which is in proportion to the sine of 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 & 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 & target
appears 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 equal height, & when not so pointed, the difference in deflection
amplitude is an Indication of the direction & magnitude of the pointing error.
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.

L-scope :- A display in which a target appears as two horizontal blips, one extending to the right from a central vertical time base & the other to the left.

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.

P-scope :- 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 & azimuth angle displayed in polar coordinates, forming Maplike display.

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 as the vertical axis & Range as the horizontal axis.

→ however, the PPI, A-scope, B-scope & RHI are among the more usual displays employed in Radar.

CRT screens :- A number of different cathode-ray tube screens are used in radar applications. they differ primarily in their decay times & persistence.

| phosphor | fluorescent color | phosphorescent color | persistence |
|---|---|---|---|
| $P_1$ | yellowish green | yellowish green | Medium |
| $P_7$ | blue | yellowish green | blue, Medium short, yellow, long. |
| $P_{12}$ | orange | orange | Long. |
| $P_{13}$ | Reddish orange | Reddish orange | Medium. |
| $P_{14}$ | purplish blue | yellowish orange | blue, medium short, yellowish orange - Medium. |
| $P_{17}$ | Blue | yellow | blue, short, yellow long. |
| $P_{19}$ | orange | orange | Long |
| $P_{21}$ | Reddish orange | Reddish orange | Medium. |
| $P_{25}$ | orange | orange | Medium. |
| $P_{26}$ | orange | orange | very long. |
| $P_{28}$ | yellowish green | yellowish green | long. |
| $P_{32}$ | purplish blue | yellowish green | long. |

| | | | |
|---|---|---|---|
| P33 | orange | orange | very long |
| P34 | bluish green | yellowish green | very long. |
| P38 | orange | orange | very long. |
| P39 | yellowish green | yellowish green | Long. |

persistence - short = 1 to 10 µs, Medium short = 10 µs to 1 ms;
Medium = 1 to 100ms ; long = 100ms to 1s ; very long = > 1 s.

→ Resolution on the CRT is limited by the phosphor characteristics as well as the electron beam. A double-layer phosphor will have poorer resolution that a single layer phosphor.

Duplexers :- A Duplexer is the device that allows a single antenna to serve both the transmitter & the receiver. on transmission it must protect the receiver from burnout or damage, & on reception it must channel the echo signal to the receiver.

→ Duplexers, especially for high-power applications, sometimes employ a form of gas-discharge device. solid state devices are also utilized.

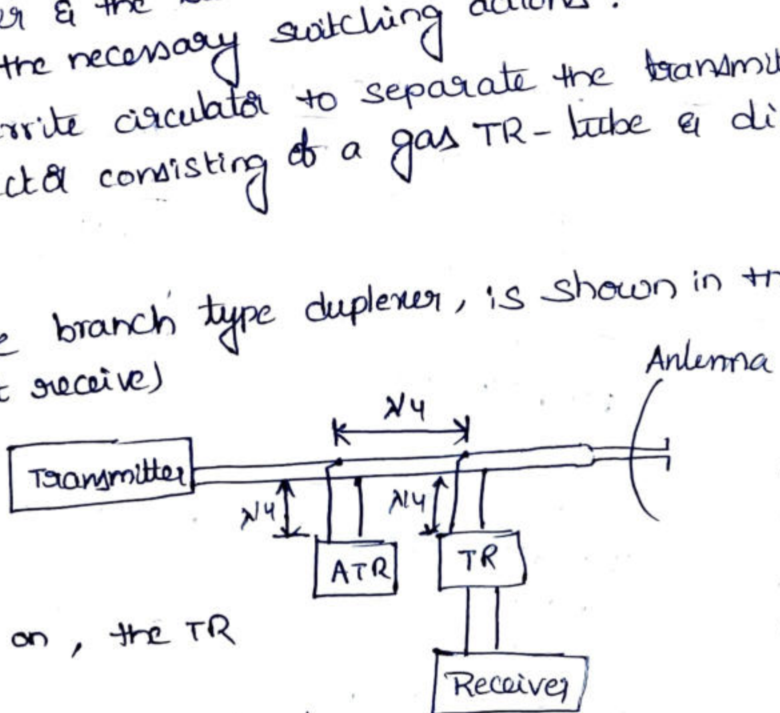
→ In a typical duplexer application the transmitter peak power might be a megawatt or more & the max safe power that can be tolerated at the receiver might be less than a watt.

→ there have been two basic Methods employed that allow the use of a common antenna for both transmitting & receiving. the older method is represented by the branch-type duplexer & the balanced duplexer which utilizes gas TR-tubes for accomplishing the necessary switching actions.

→ the other method uses a ferrite circulator to separate the transmitter & receiver, and a receiver protector consisting of a gas TR-tube & diode limiter.

Branch-type Duplexer :- The branch type duplexer, is shown in the fig. It consist of a TR (transmit receive) switch & ATR (Anti transmit receive) switch, both of which are gas discharge tubes.

→ when the transmitter is turned on, the TR & the ATR tube Ionize.




principle of branch-type duplexer.

ie, 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 wavelength from the main transmission line, so that it appears as a short ckt at the receiver but as an open ckt at the transmission line so that it does not impede the flow of transmitter power.

→ since the ATR is displaced a quarter wavelength from the main transmission line & thus has no effect on transmission. During reception, the transmitter is off & neither the TR nor the ATR is fired. the open ckt of the ATR, being a quarter wave from the transmission line, appears as a short ckt across the line.

→ since this short ckt is located a quarter wave from the receiver branch line the transmitter is effectively disconnected from the line & the echo signal power is directed to the receiver.

→ The branch type duplexer is of limited bandwidth & power handling capability, & has generally been replaced by the balanced duplexer & other protection devices. it is used in spite of these limitations in some low cost radars.

## Balanced Duplexers :-

→ The balanced duplexer 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 b/w the two. the short-slot hybrid may be considered as a broadband directional coupler with a coupling ratio of 3dB.

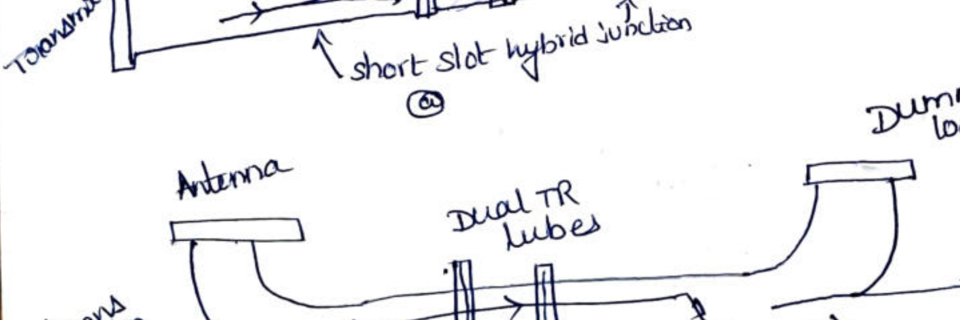
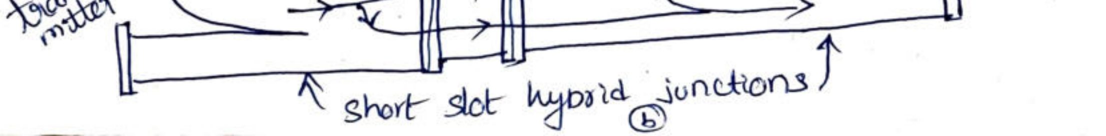



Fig: balanced duplexer using dual TR tubes & two short slot hybrid junctions.

ⓐ Transmitter condition

ⓑ receiver condition.



→ In the transmit condition power is divided equally into each wave guide by the first short slot hybrid junction. Both TR tube breakdown & reflect the incident power out the antenna arm as shown. the short slot hybrid has the property that each time the energy passes through the slot in either direction, its phase is advanced 90°.

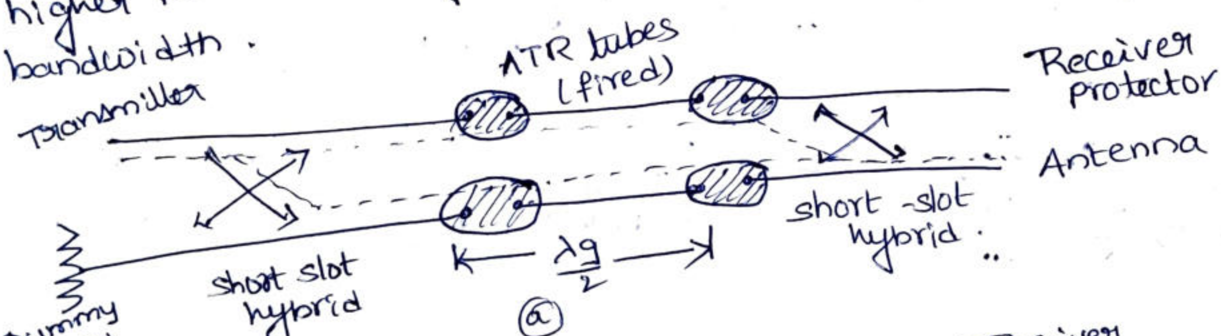
→ therefore, the energy must travel as indicated by the solid lines. Any energy which leaks through the TR tubes (shown by the dashed lines) is directed to the arm with the matched dummy load & not to the rxer.

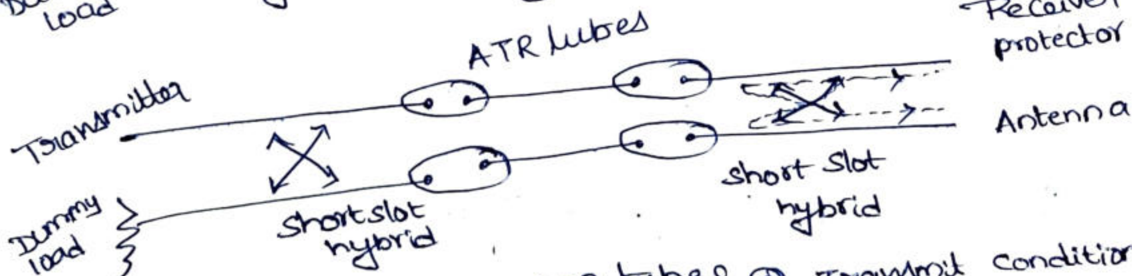
→ on reception the TR tubes are unfired & the echo signals pass through the duplexer & into the receiver as shown in fig (b) the power splits equally at the first junction & bcoz of the 90° phase advance on passing through the slot, the energy recombines in the receiving arm & not in the dummy-load arm.

→ The power handling capability of the balanced duplexer is inherently greater than that of the branch type duplexer & it has wide bandwidth, over ten percent with proper design.

→ Another form of balanced duplexer uses four ATR tubes & two hybrid junctions. During transmission the ATR tubes located in a mount b/w the two short slot hybrids ionize & allow high power to pass to the antenna. During reception the ATR tubes present a high impedance which results in the echo signal power being reflected to the receiver. the ATR type of balanced duplexer has higher power handling capability than that of TR tubes but it has less bandwidth.

→ Dashed lines show the flow of power.




Balanced duplexer using ATR tubes (a) Transmit condition (b) receiver condition.

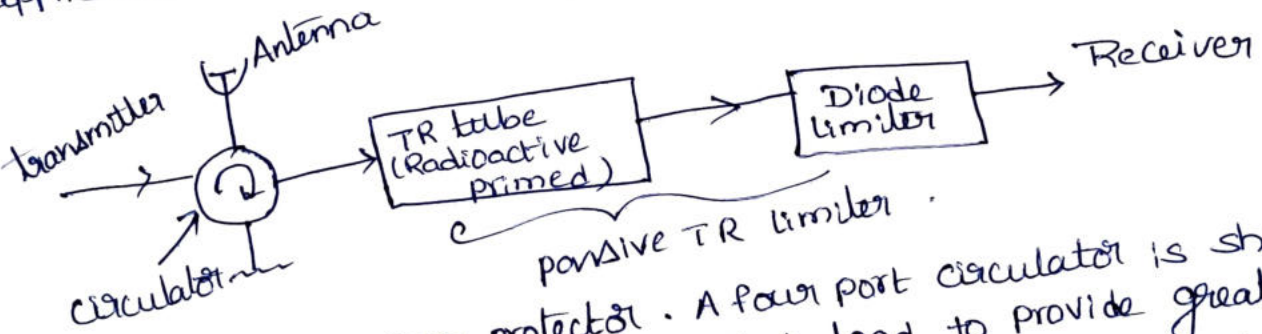
Circulator & Receiver protector :- The ferrite circulator is a three or four port device that can in principle offer seperation of the transmitter & receiver without the need for the conventional duplexer configuration. the circulator does not provide subficient protection by itself & requires a receiver protector as shown in the fig.

→ the isolation b/w the transmitter & receiver ports of a circulator is subficient to protect the receiver from damage. however, it is not the isolation b/w transmitter & 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 the measure of the amount of power reflected by the antenna

→ thus, the receiver protector is almost always required. It also reduces to a safe level radiations from near by transmitters.

→ the receiver protector might use solid-state diode for an all solid state configurations, or it might be a passive TR-limiter consisting of a radioactive primed TR-tube followed by the diode limiter.

→ The ferrite circulator with receiver protector is attractive for radar applications bcoz of its long life, wide bandwidth, & compact design.




→ circulator & receiver protector. A four port circulator is shown with the fourth port terminated in matched load to provide greater isolation b/w the transmitter & receiver than provided by a three port circulator.