# LECTURE NOTES
# ON
# RADAR SYSTEMS

## VI semester (R16)

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# UNIT I

# FUNDAMENTALS OF RADAR

## Lecture 1 :-

### Introduction to Radar :-

Radar is an Electro-magnetic system for the detection and location of objets. It operates by transmitting a perticular type of waveform and detect the nature of echo signal.

→ Radar can't resolve detail as well as the eye, nor able to recognise the color of objets to the degree which the eye is capable.

→ Radar can be designed to see through darkness, haze, fog, rain and snow. In addition radar measure the distance or range to the object.

→ The name radar reflects the emphasis placed by the early experiments on a device to detect the presence of a target and measure its range. Radar is contraction of words radio detection and ranging. It was first developed as a detection device to warn of the approach of hostile aircraft and for directing antiaircraft weapons.

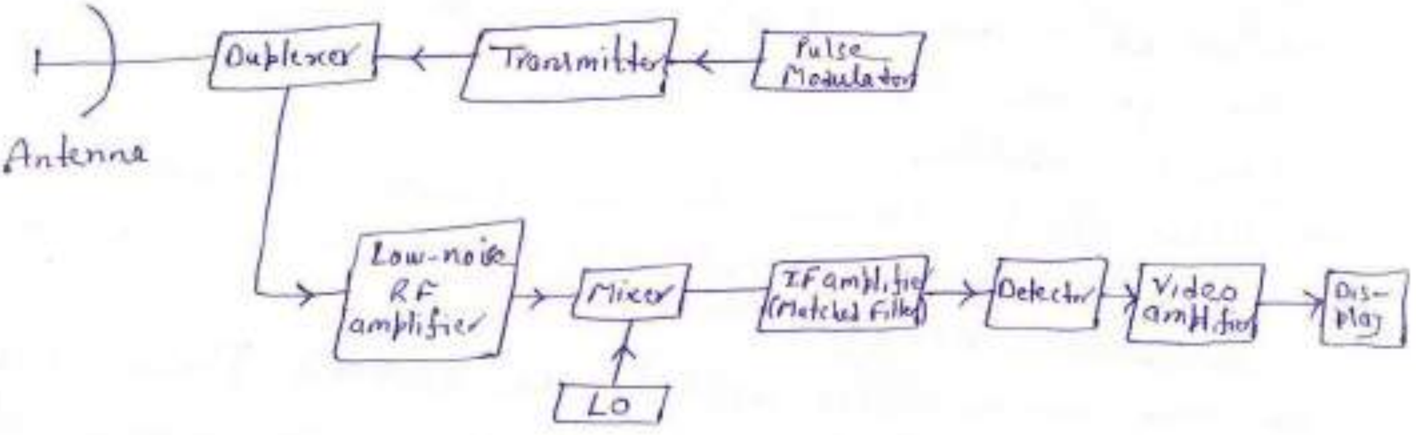
→ Basic radar consist of a transmitting antenna emitting EM waves generated by an oscillator of some sort, a receiving antenna, and a energy detection device, or receiver. A portion of transmitted signal is intercepted by a reflecting object (target) and is re-radiated in all directions. The receiver antenna collect the return energy, where receiver process to detect the presence of target and also its location. The distance to the target is determined by measuring the time taken for the radar signal to travel to the target back. The direction and angular position of the target may be determined from the direction of arrival of reflected wave front.

→ The most common radar waveform is a train of narrow, rectangular-shape pulses modulating a sine carrier. The distance or range to the target is determined by measuring the time taken by the pulse to travel to the target and return.

$$\text{Range } R = \frac{c T_R}{2}.$$

$$R(\text{km}) = 0.15\, T_R\, (\mu s)$$

## Radar Block Diagram :-



Lecture - 2

## Block Diagram operation :-

The transmitter may be an oscillator, such as an magnetron, that is pulsed by the modulator to generate a repetitive train of pulses. Magnetron used most widely of various microwave generators for radar. A typical radar for the detection of aircraft at ranges of 100 or 200 (nmi → nautical miles) might employ a peak power of the order of megawatt, an average power of several kilowatts, a pulse width of several micro-seconds and a pulse repetition frequency of several hundred pulses per second.

→ The waveform generated by the transmitter travels via a transmission line to the antenna.

Duplexer:— ① Protect receiver from damage caused by the high power of transmitter, also ② serves to channel the returned echo signals to the receiver and not to transmitter. Duplexer consist of two gas discharge devices, one known as TR (transmit-receive) and other an ATR( anti-transmit-receive). TR for ① task and ATR for ②. Solid state ferrite Circulator and receiver Protector with gas plasma TR devices employed as duplexer

Receiver: → It is usually of the Super-heterodyne type. First stage might be low-noise RF amplifier but not always desirable. The mixer and LO convert the RF Signal to the intermediate frequency (IF). A typical IF amplifier for radar have a center frequency of 30 to 60 MHz and B.W of the order one MHz. The IF amplifier Should be designed as a matched filter i.e it frequency-response function |H(f)| should ~~be~~ maximize peak-signal to mean noise Power ratio at the output. It occur when magnitude of |H(f)| is equal to magnitude of the echo signal Spectrum |S(f)|, and the Phase Spectrum of matched filter is the negative of Phase Spectrum of echo signal.

After maximizing the SNR in the IF amplifier, the Pulse modulation is extracted by detector and amplified by video amplifier to a level where it can be Properly displayed usually on CRT. Timing signals are also supplied to the indicator to provide the range zero. Angle information is obtained from the Pointing directions of antenna.

Applications of Radar: →

Radar has been employed on the ground, in the air, on the sea, and in Space.

① Ground based radar has been applied chiefly to the detection, location, and tracking of aircraft or space targets.

② Shipboard radar is used for navigation aid and safety device to locate buoys, Shore lines and other ships, as well as for observing aircraft.

③ Airborne radar may be used to detect other aircraft, Ships, or land vehicles, or may be mapping of lands.

Storm avoidance, terrain avoidance, and navigation.

④ In space, radar has assisted in the guidance of space-craft and for remote sensing of the land and sea.

Major area of radar applications are.

- →Air Traffic control (ATC
- → Air craft Navigation
- → Ship Safety
- → Space
- → Remote sensing
- → law Enforcement
- → Military

## Lecture - 3

### Simple Form of Radar equation: →

Radar Equation relates the range of a radar to the characteristics of the transmitter, receiver, antenna, targe and environment. It is useful not just as a means of 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 basic for rada design.

let Power of radar transmitter is $P_t$, and if isotropi antenna is used, the Power density (watt/area) at a distance R from the radar is equal to the transmitte Power divided by the surface area of an imaginary Sphere of radius R.

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

Radar employ directive antenna having gain G.

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

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

where σ ⇒ radar cross section

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

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

The maximum radar range $R_{max}$ is the distance beyond which the target can't be detected. It occurs when the received echo signal power $P_r$ just equals the minimum detectable signal $S_{min}$.

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

↖ Fundamental form of Radar $Eq^n$.

Minimum Detectable ~~Si~~ Signal :→

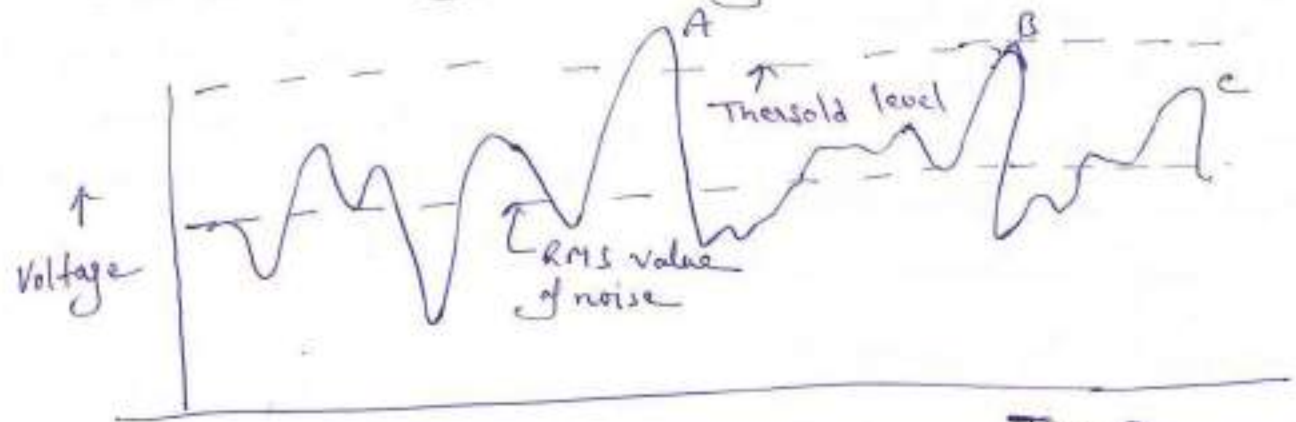
The ability of radar receiver to detect a weak echo signal is limited by noise that occupies the same portion of the frequency spectrum as does the signal energy.

→ The weakest signal the receiver can detect is called minimum detectable signal.

MDS not easy to detect in some times because of statistical nature and criterion to decide whether the target present or not.

Detection is based on thersold level at the O/P of the receiver i.e thersold detection

The thersold level must be low if weak signals are to be detected, but it can't be so low that noise peaks cross the thersold and give a false indication of the Presence of ~~noise~~ targets.



Typical envolpe of the radar receiver O/P as a func$^n$ of time

As shown in figure a target is said to be detected if the envolpe crosses the thersold level. Target A ~~has~~ is not a difficult to decide as in the case of B and C.

## Lecture - 4

**Receiver Noise :→** Noise is the main factor limiting receiver sensitivity, it is necessary to obtain some means of describing it quantitatively. Noise is unwanted EM energy interfere with wanted signal.

The available thermal noise Power generated by a receiver of B.W $B_n$ (Hertz) at temp$^n$ T (°K) is equal to

$$\text{Thermal noise Power} = KTB_n$$

$K = 1.38 \times 10^{-23}$ J/deg = Boltzman constant.

For super-hetrodyne receiver mostly used in radar, the receiver B.W is approximately that of the intermediate frequency B.W.

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

where $H(f)$ = frequency-response characterstic of IF amplifier
$f_0$ = frequency of maximum response

Except the thermal-noise some noise components also present. The exact origin of extra noise components is not important except to know that it exist. No matter wether the noise is generated by a thermal mechanism or by some other mechanism the total noise at the o/p of receiver may be considered to be equal to the thermal noise power obtained from an ideal receiver multiplied by a factor called a noise figure.

$$\text{Noise Figure} = F_n = \frac{N_o}{K T_0 B_n G_a} = \frac{\text{noise o/p of Practical receiver}}{\text{noise out of ideal receiver at std temp}^r\ T_0}$$

where

$N_o$ = noise o/p from receiver
$G_o$ = Gain, $T_0 = 290\,K$, standard temp$^r$.

Also $\quad F_n = \dfrac{S_i/N_i}{S_o/N_o} \qquad$ As $G_a = \dfrac{S_o}{S_i}$ and $K T_0 B_n = N_i$

$$\Rightarrow \quad F_n = \frac{S_i N_o}{S_o K T_0 B_n}$$

$$\Rightarrow \quad S_i = \frac{K T_0 B_n F_n S_o}{N_o}$$

if the minimum detectable signal $S_{min}$ is the value of $S_i$ corresponding to the minimum ratio of o/p signal to noise ratio $(S_o/N_o)_{min}$ necessary for detection, then

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

Also from Radar eq$^n$

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

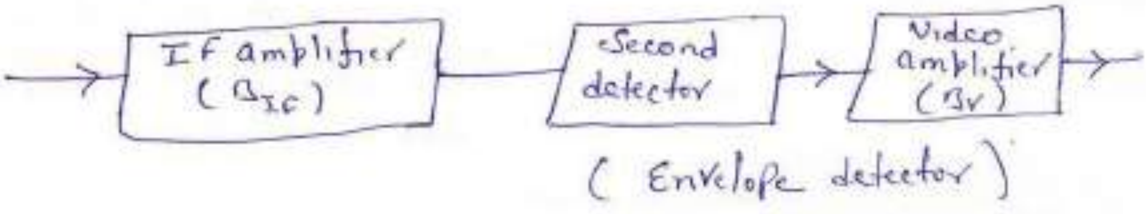
Put value of $S_{min}$

$$\Rightarrow \quad R_{max}^4 = \frac{P_t G A_e \sigma}{(4\pi)^2 K T_0 B_n F_n (S_o/N_o)_{min}}$$

## Signal-to Noise Ratio:→

Signal to noise ratio is very imbortan as far as Radar is concerned, Because Presence of targ or not have small differnce. Statistical noise theory will be applied to obtain S/N at the o/p of the IF amplifier necessary to achieve a Specified Probability of detecti without exceeding a Specified Probability of false alarm.

Consider an IF amplifier with B.W $B_{IF}$ followed by a second detector and a video amplifier with BW $B_v$.



To Extract modulation Envelope, the video B.W must be wide enough to Pass the low frequency component generated by Second detector, but not So wide Enough to Pass high frequency component at or near I.F. The video B.W mu be greater than $B_{IF}/2$ in order to pass all the video mod -lation.

## Lecture - 5

## Signal to noise Ration:→

The noise entering the IF amplifier is assumed to b gaussian, with pdf given by

$$p(v)=\frac{1}{\sqrt{2\pi\sigma}}\,\text{Exp}\left(\frac{-v^2}{2\sigma}\right)$$

with mean zero and variance $\sigma$.

$p(v).dv$ is the Probability of finding noise voltage $v$ b/w the values of $v$ and $v+dv$.

If gaussian noise Passed through a narrowband IF filter one whose B.W is small compared with the mid frequency - the Probabili density of the envelope of noise voltage o/p is given by

$$P(R) = \frac{R}{\sigma_0} \, Exp\left(-\frac{R^2}{2\sigma_0}\right) \quad \text{(Rayleigh Pdf)}$$

where R → amplitude of envelope of the filter o/P.

The probability that the envelope of the noise voltage will lie b/w the values of $v_1$ and $v_2$ is

$$P(v_1 < R < v_2) = \int_{v_1}^{v_2} \frac{R}{\sigma_0} \, Exp\left(-\frac{R^2}{2\sigma_0}\right) dR$$

The probability that the noise voltage envelope will exceed the thersold voltage ($V_T$)

$$P(V_T < R < \infty) = \int_{V_T}^{\infty} \frac{R}{\sigma} \, Exp\left(-\frac{R^2}{2\sigma_0}\right) dR = exp\left(-\frac{V_T^2}{2\sigma_0}\right) = P_{fa} \quad - ①$$

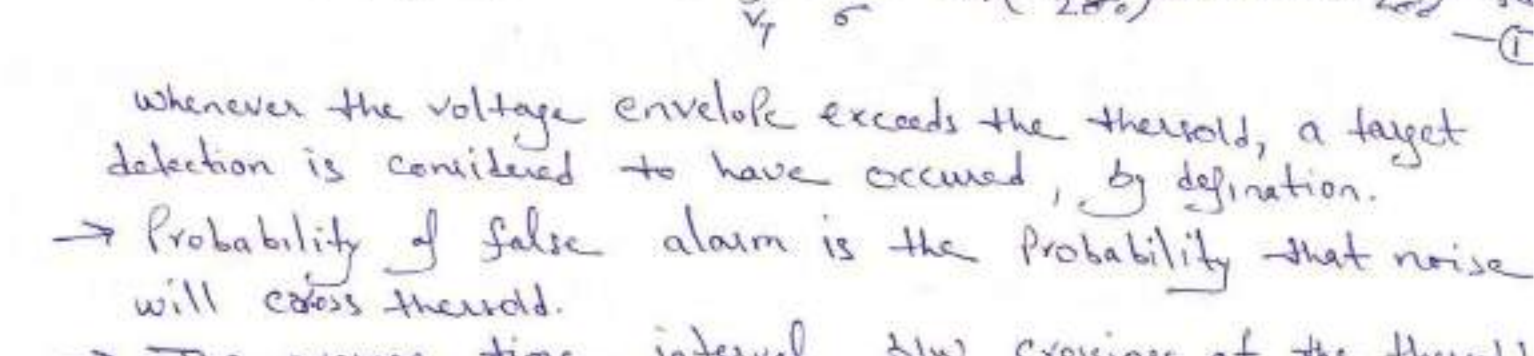
whenever the voltage envelope exceeds the thersold, a target detection is considered to have occured, by defination.

→ Probability of false alarm is the probability that noise will cross thersold.

→ The average time interval b/w crossings of the thersold by noise alone is defined by false alarm time $T_{fa}$

$$T_{fa} = \lim_{N \to \infty} \frac{1}{N} \sum_{K=1}^{N} T_K$$

Probability of false alarm $P_{fa} = \frac{1}{T_{fa} B} = \frac{\sum_{K=1}^{N} t_K}{\sum_{K=1}^{N} T_K} \quad - ②$



Envelope of receiver o/P

From ① and ②

$$T_{fa} = \frac{1}{B_{IF}} \, Exp\left(\frac{V_T^2}{2\sigma_0}\right)$$

A receiver with only noise i/P has been discussed. Now assume a sine wave signal of amplitude A present with noise.

Frequency of the signal is the same as the IF midband freq
The O/P of the envelope detector has a probability-density func

$$P_s(R) = \frac{R}{\sigma_0} \operatorname{Exp}\left(-\frac{R^2+A^2}{2\sigma_0}\right) I_0\left(\frac{RA}{\sigma_0}\right) \quad - ③$$

where $I_0(Z)$ is the modified Bessel funcn of zero ord
and argument Z. Eqn ③ sometime called Rice PDF.
Now Probability of detection $P_d$ is given by

$$P_d = \int_{V_T}^{\infty} P_s(R)\,dR = \int_{V_T}^{\infty} \frac{R}{\sigma_0} \operatorname{Exp}\left(-\frac{R^2+A^2}{2\sigma_0}\right) I_0\left(\frac{RA}{\sigma_0}\right)$$

Solving this and assuming $RA/\sigma_0 >> 1$ and $A >> |R-A|$

$$P_d = \frac{1}{2}\left(1 - \operatorname{erf}\frac{V_T - A}{\sqrt{2\sigma_0}} + \frac{\operatorname{Exp}\left[-(V_T-A)^2/2\sigma_0\right]}{2\sqrt{2\pi}\left(A/\sqrt{\sigma_0}\right)} \times \left[1 - \frac{V_T - A}{4A} + \frac{1-(V_T-A)^2}{8A^2/\sigma_0}\right.\right.$$
$$+ \cdots$$

$$\text{where } \operatorname{erf} Z = \frac{2}{\sqrt{\pi}} \int_0^Z e^{-u^2}\,du \qquad — ④$$

Although the receiver designer prefer to operate with voltages, it is more convenient for radar system engineering to employ power relationships.

$$\frac{A}{\sigma_0^{1/2}} = \frac{\text{Signal amplitude}}{\text{r.m.s noise voltage}} = \frac{\sqrt{2}\ (\text{r.m.s signal voltage})}{\text{r.m.s noise voltage}}$$

$$= \left(2\,\frac{\text{Signal Power}}{\text{noise Power}}\right)^{1/2} = \left(\frac{2S}{N}\right)^{1/2}$$

Lecture-6

Transmitter Power:-

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

where transmitted Power $P_t$ also called Peak Power. This is not instantaneous Power of Sine wave. The average radar Power $P_{av}$ is also of interest in radar and defined as the average Power over the Pulse repetition Period.

If transmitted Pulse is rectangular with width $\tau$ and Pulse repetition Period $T_p = 1/f_p$, the average Power is related to the Peak Power by

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

The ratio $P_{av}/P_t$, $\tau/T_p$ or $\tau f_p$ is called duty cycle

writting Eq$^n$ ① in terms of average Power

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

The B.W and Pulse width are grouped together since the Product of two is usually of the order of unity in most Pulse-radar appl$^n$.

if the transmitted waveform is not a rectangular Pulse, it is sometimes more convenient to express radar Eq$^n$ in terms of Energy.

$$E_\tau = P_{av}/f_p$$

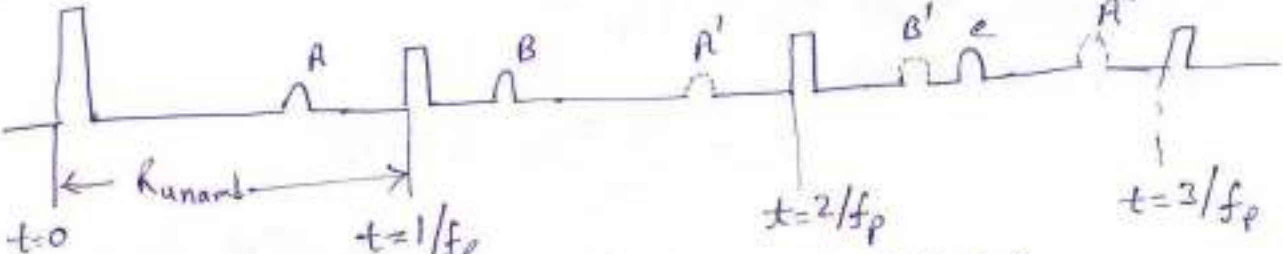
$$\Rightarrow R^4_{max} = \frac{E_\tau G A_e \sigma n E_i(n)}{(4\pi)^2 K T_0 F_n (B_n \tau)(S/N)_1}$$

The important Parameters affecting range are the total transmitted Energy $nE_\tau$, Transmiter gain $G$, effective receiver aperture $A_e$, and receiver noise Figure $F_n$.

# Pulse Repetition Frequency and Range Ambiguities

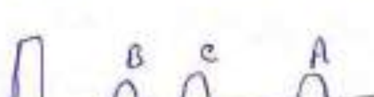
The Pulse rebetion frequency (Prf) is determined primarily by the maximum range at which targets are expected. if the Prf is made too high, the likehood of obtaining target echoe from the wrong Pulse transmission is increased. Echo signals received after an interval exceeding the Pulse repetition perio are called multiple-time around echoes.

Consider the three targets labeled A, B and C as shown in Figure.



Time (or range) (Fig 1)

Multiple-time around echoes that give rise to Ambiguiti



Fig(2)

In above Fig 1, Three targets A, B, C, where A within Runamb and B and C are mutiple time around target.

Fig(2) Shows three targets on A-Scope.

Fig(3)

Fig(3) Shows three targets on A-Scope with changing Prf.

Ambiguities may theoretically be resolved by observing the variatio of echo signal with time (range). This is not always practica technique, however since the echo signal amplitude can fluctuate strongly for reasons other than a change in range. Instead the range ambiguities in mutiple Prf radar can be conveniently decoded and the true range find by computional algorithms.

## Lecture-7

System losses:→ one of the main factors omitted from the simple radar $Eq^n$ was the losses that occur throughout the radar system. The losses reduce the S/N at the receiver o/p. They may be of two kinds depending upon wether or not they can be predicted with ~~the~~ any degree of precision beforehand. The antenna beam shape loss, collapsing loss, and losses in microwave plumbing are example of losses which can be calculated if the system configuration is known. These losses are real and can't be ignored in any serious prediction of radar performance.

Following are the main losses occured in Radar

Plumbing loss:— There is always some finite loss in the transmission lines which connect the o/p of the transmitter to the antenna. The losses in decibels per 100ft for radar transmission lines. At the lower radar frequencies the transm-ssion line introduce little loss, unless its length is exceptionall long. At the higher radar frequencies attenuation may not alway small and may have to be taken into accont. one more loss that can occur at each connection or bend in the line and at the antenna rotary joint if used.

connector losses are usually small but if the connectors are poorly made, it can contribute significant attenuation.

Since the same Tx line is generally used for both receiving and transmission, the loss to be inserted in the radar equation is twice the one way loss.

The signal suffers attenuation as it passes through the duplexe Generally, the greater the isolation required from the duplexer o transmission, the larger will be insertion loss.

In an S-band (3000 MHz) radar, plumbing losses might be

| | |
|---|---|
| 100ft of waveguide Tx line (two-way) | 1.0dB |
| loss due to Poor connection | 0.5dB |
| Rotary joint loss | 0.4dB |
| Duplexer loss | 1.5dB |

Beam Shape loss:→ The antenna gain in radar equation was assumed constant equal to maximum value. But in reality the train of pulse returned from a target with scanning radar is modulated in amplitude by the ~~sat~~ Shape of antenna beam. Instead of beam shape loss is added to radar equation to account for the fact that maxi. gain is employed in radar equation rather than a gain that changes Pulse to Pulse.

When the antenna Scan rapidly enough that the gain on transmitter is not the same as the gain on receiver, this loss is scanning loss.

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 circum-stances. Some receiver, however might employ limiting for some special Purpose, as for Pulse compression processing for Example.

Limiting results in a loss of only a fraction of dB for a large no. of pulses integrated, provided the limiting ratio.

Collapsing loss:— if the radar were to integrate additional noise samples along with a wanted S/N pulses, the added noise results in degradation called collapsing loss. It can occur in displays which ~~can~~ collapse the range information. A collapsing loss can occur when the O/P of high resolution radar is displayed on a device whose resolution is coarser than that inherent in radar. A collapsing loss also result if

O/P of two or more radar receiver are combined and only one contain signal while other contain noise.

## Lecture - 7

Nonideal equipment:— The transmitter power in radar equation was assumed to be O/P power. However transmitting devices (or components) not uniform in quality, nor should it be expected that any individual BJT or JFET (FET) remain at same level of performance through-out its usefull life. Also all the power is usually not uniform over the operating band of the devices. Thus for one or more reasons a loss factor may be introduced.

Operator loss:— Distracted, tired, overloaded, or not properly trained operator performance will decrease that will cause losses.

Field degradation:→ Factors which contribute to field degradation are poor tunning, weak components, water in Tx lines, incorrect mixer-crystal current, deterioration of receiver noise figure, loose cable connection etc.

Other loss factors:→ A radar designed to discriminate b/w moving targets and stationary objects may introduced additional loss over a radar without this facility. This discrimination technique results in complete loss of sensitivity for certain values of target velocity relative to the radar. These are called blind speeds.

The stradding loss accounts for the loss in SNR for targets not at the center of the range gate or at the center of the filter in multiple filter bank processor.

## Propagation Effects :-

In analyzing radar performance it is convenient to assume that the radar and target are both located in free space. However there are very few radar applications which approximate free space condition.

In most cases of practical interest, the earth surface and medium in which radar wave propagate can have a significan effect on radar performance. In some instant propagation factor might be important enough to overshadow all other factors that contribute to abnormal radar performance.

The effect of non-free space propagation on the radar are of three categories

1) Attenuation of the radar wave as it propagates throu the earth's atmosphere,
2) Refraction of radar wave by the earth's atmosphere. and
3) Lobe structure caused by interference b/w the direct wave from radar to target and the wave which arrives at the target via reflection from the ground.

# UNIT II

# CW AND FREQUENCY MODULATED RADAR

# The Doppler Effect

A radar detects the presence of objects and locates their position in the space by transmitting EM energy and observing the return echo. Presence of echo not only indicates the presence of target, but the time that elapses b/w transmission of pulse and the receipt of the echo is a measure of the distance to the target. Sepration of the echo signal and transmitted signal is made on the basis of the difference time.

It is well known in the fields optics and acoustics that if either the source of oscillation or the observer of the oscillations is in motion, an apparent shift in frequency will result. This is the Doppler effect and basis of CW radar.

$R \rightarrow$ Distance b/w target & radar

$\Rightarrow \frac{2R}{\lambda}$ No. of wavelengths $\lambda$ covered (contained) in the two way path b/w radar and target.

One wavelength corresponds to an angular excursion of $2\pi$ radian, ⇒ total excursion made by wave during it transmit to and from the target $\Rightarrow \frac{4\pi R}{\lambda} = \phi$ (Total excursion)

if the target is in motion $R$ and phase $\phi$ continually changing. In change in $\phi$ w.r.t time equal to frequency. This is the doppler angular frequency $\omega_d$.

$$\omega_d = 2\pi f_d = \frac{d\phi}{dt} = \frac{4\pi}{\lambda}\frac{dR}{dt} = \frac{4\pi v_r}{\lambda} \quad — ①$$

$f_d \rightarrow$ doppler frequency shift

$v_r \rightarrow$ relative (or radial) velocity of target w.r.t radar

$$\Rightarrow f_d = \frac{2v_r}{\lambda} = \frac{2v_r f_0}{c}$$

$f_0 \rightarrow$ transmitter frequency

$$\Rightarrow f_d = \frac{1.03\, v_r}{\lambda}$$

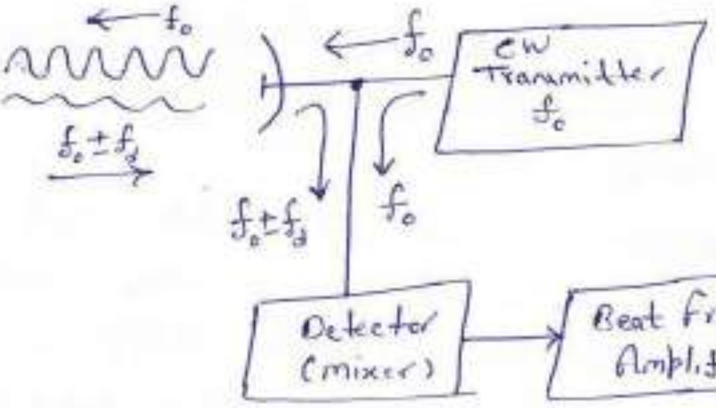
relative velocity can be written as $v_r = v\cos\theta$

$v \rightarrow$ target speed and $\theta$ angle made by target trajectory and

the line joinging radar and target.
When θ=0 doppler frequenty is maximum.

## CW Radar: →

CW stands for continuous ~~radar~~ wave radar.



(Simple CW Radar)

~~Tran~~ Transmitter generates a unmodulated wave of frequency $f_0$, which is radiated by antenna. if the target is in motion with a velocity $v_r$, relative to radar, the received signal will be shifted in frequency from the transmitted frequency $f_0$. the purpose of doppler amplifier is to eliminate echoes from sationary targets and to amplify the doppler echoes signal to a level where it can operate an indicating device. the indicator might be a pair of earphones or a frequency meter.

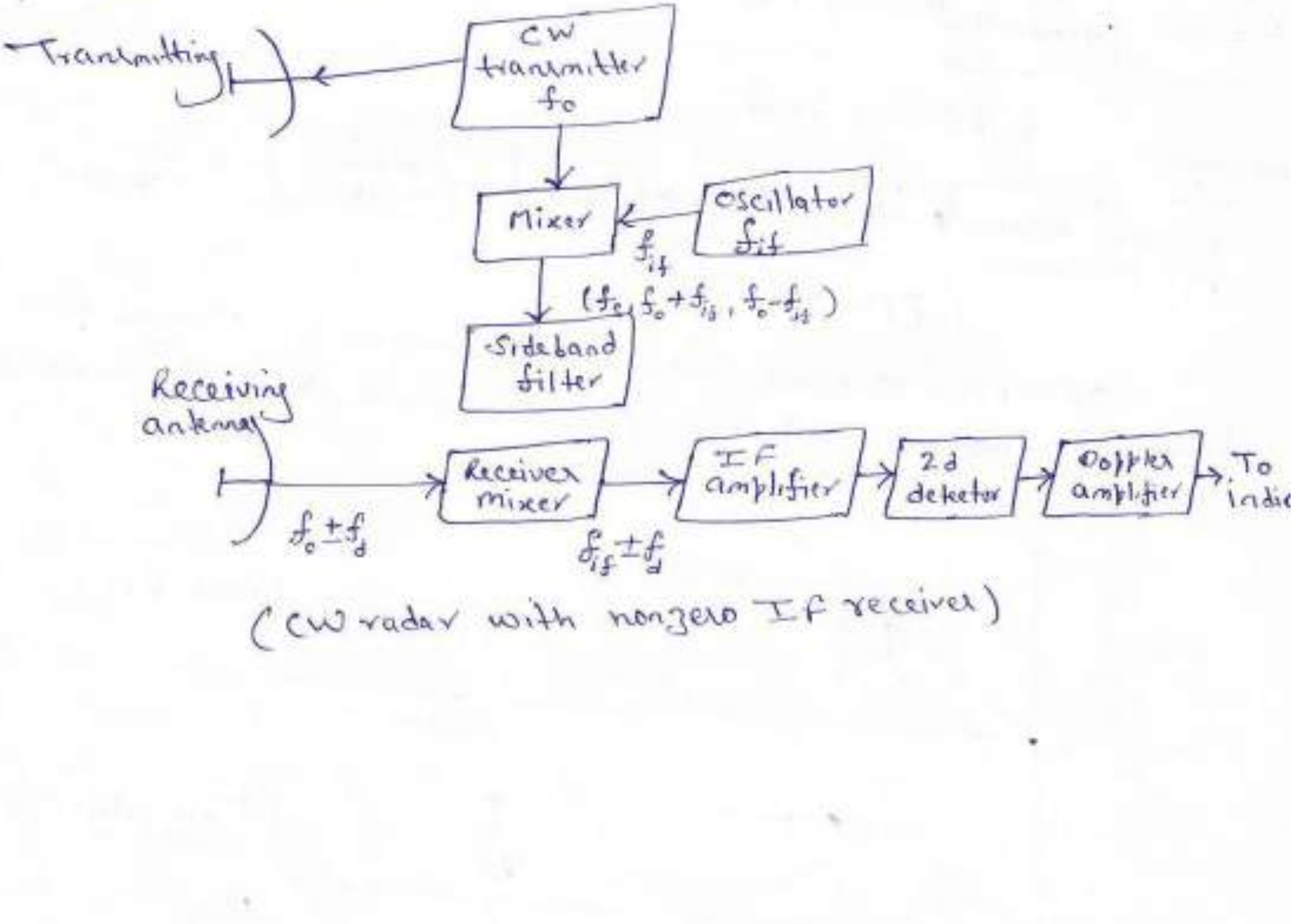
isolation b/w transmitter and receiver — A single antenna serves the purpose of transmission and reception in the simple CW radar. The necessary isolation b/w the transmitter and receiver is acheived via sepration in the frequency as a result of doppler effect.

The amount of isolation required depends on the transmitter power and the accompanying transmitter noise (as well as) and the sensitivity of the receiver.

Intermediate - Frequeng Receiver

flicker noise is the main effect which produce distortion in the received signal. flicker noise occurs in semiconductors such a diode detectors, transistor etc. The noise power produced by flicker effect varies as $1/f$.

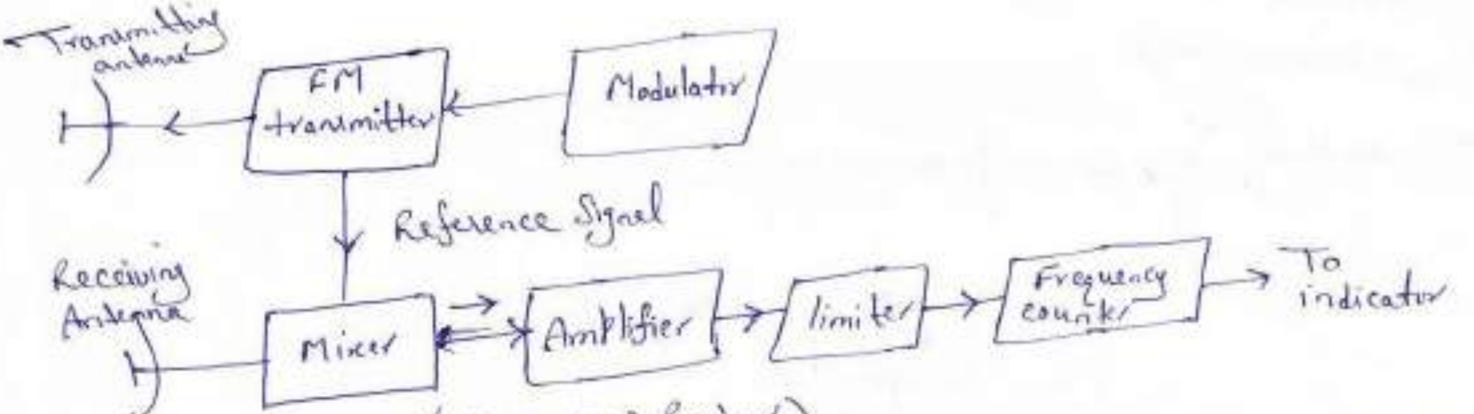
The effect of flicker noise are overcome in norma Superhydrodyne receiver by using an intermediate frequency hi enough to render flicker noise small as compare to norm receiver noise.



(CW radar with nonzero IF receiver)

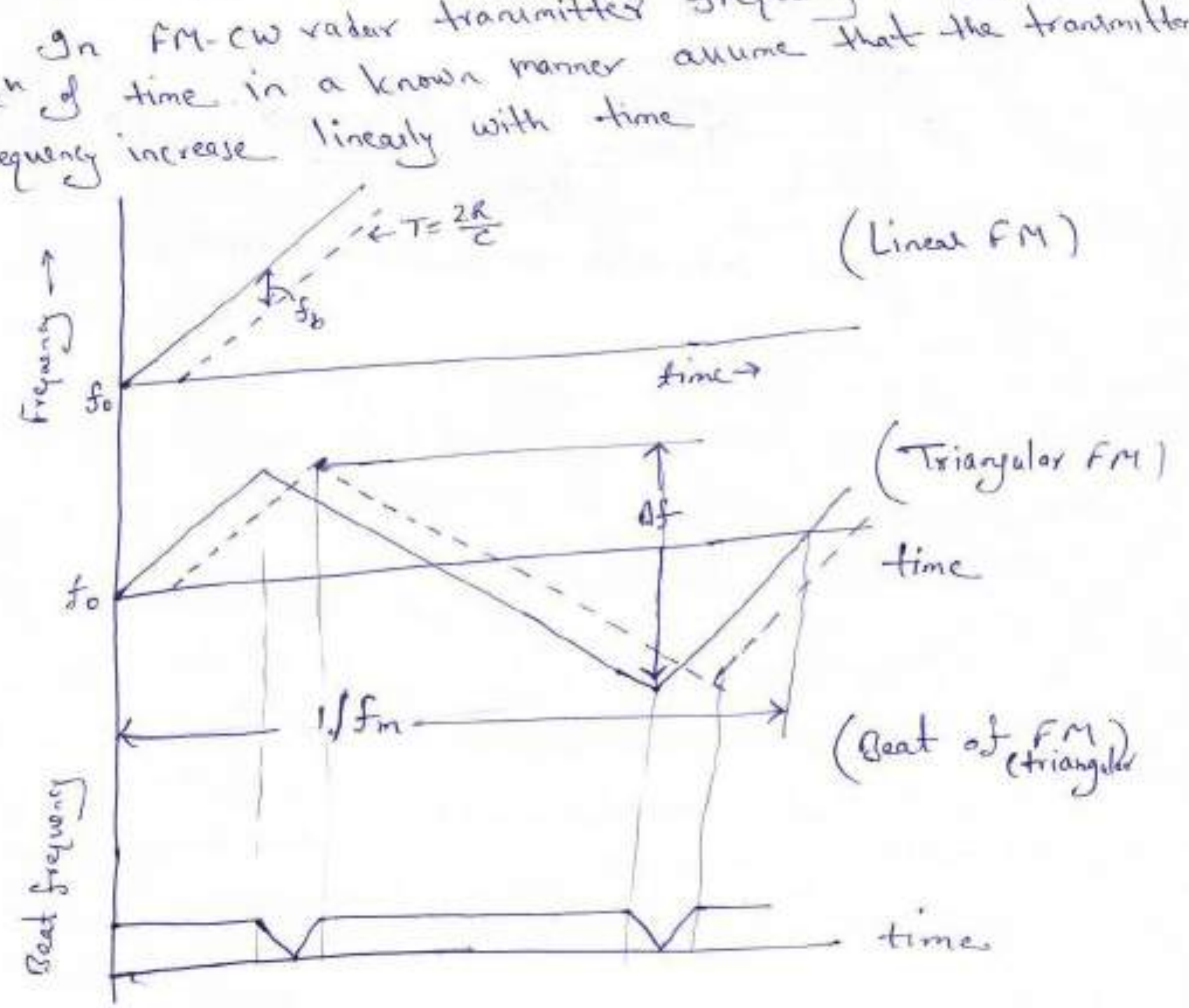
## FM-CW Radar:(Frequency modulated) CW Radar:—

The inability of a simple CW radar to measure range is related to the relatively narrow Spectrum (B.W) of its transmitted waveform. Some sort of timing mark must be applied to CW carrier if range is to be measured. The timming mark permits the time of transmission and the time of return to be recognised. The sharper ~~the~~ or more distinct the mark, the more accurate the measurement of transit time



(FM-CW Radar)

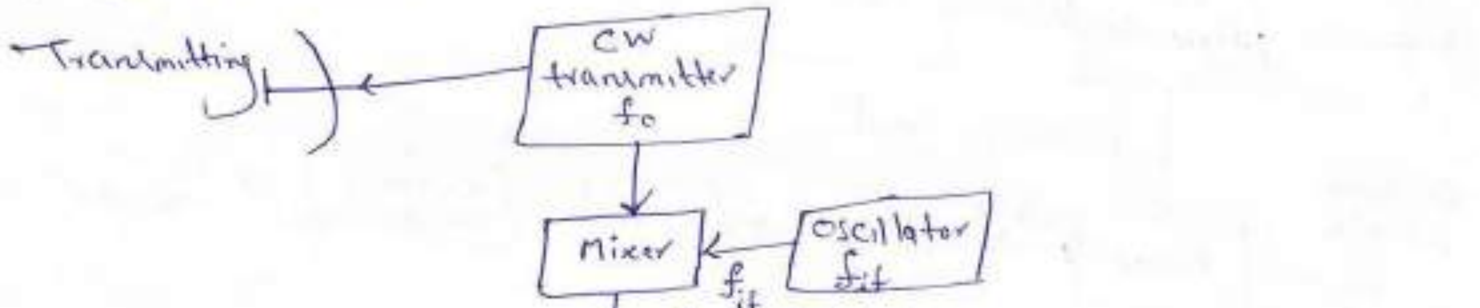
In FM-CW radar transmitter frequency changed as a func$^{n}$ of time in a known manner assume that the transmitter frequency increase linearly with time



# Intermediate - Frequeny Receiver

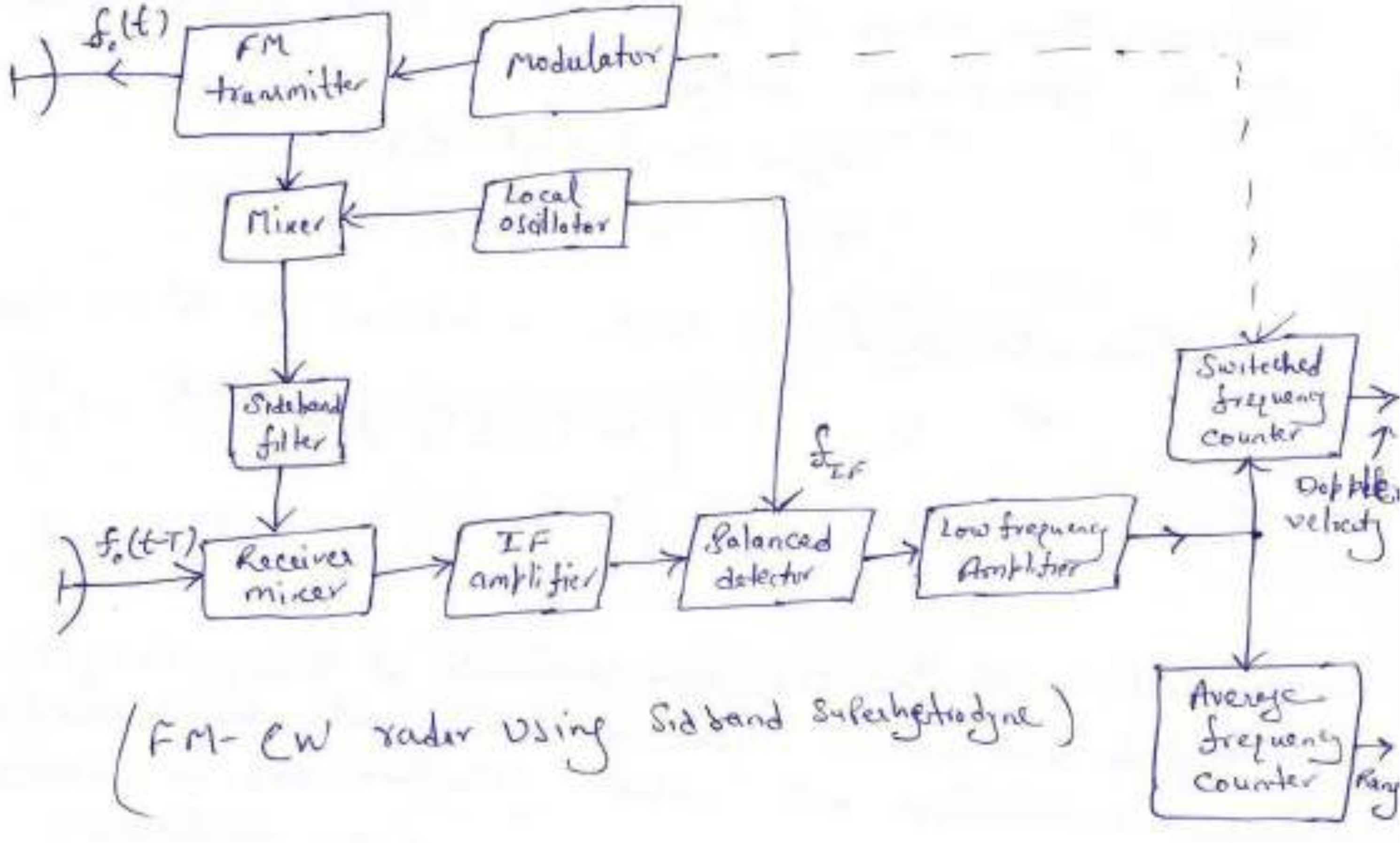
Flicker noise is the main effect which produce distortion in the received Signal. flicker noise occurs in Semiconductors Such a diode detectors, transistor etc. The noise power produced by flicker effect varies as 1/f.

The effect of flicker noise are overcome in normal Superhydrodyne receiver by using an intermediate frequency hig enough to render flicker noise small as compare to norm receiver noise.



(CW radar with nonzero IF receiver)

**FM-CW altimeter :—** The FM-CW radar pricible is used in aircraft radio altimeter to measure height above the surface of each. The laye backscatter cross section and the relative short ranges required of altimeter permit low transmitter power and low antenna gain.



(FM-CW radar using Sidband Superhetrodyne)

## MULTIPLE FREQUENCY CW RADAR:—

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 transmitted signal.

Consider a CW radar radiating a single frequency sine wave of form $\sin 2\pi f_0 t$. The signal travel to the target at a range R and returns to the radar after a time $T = 2R/c$. The echo signal received at the radar is in $[2\pi f_0(t-T)]$. If the transmitted and received signals are compared in phase detector, the O/P is proportional to the phase difference b/w 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, or $R = \frac{C\Delta\phi}{4\pi f_0} = \frac{\lambda}{4\pi}\Delta\phi$ —①

When put $\Delta\phi = 2\pi$ into ① gives the maximum unambig -ous range as $\lambda/2$.

The transmitted waveform is assumed to consist of two continuous Sine waves of frequency $f_1$ and $f_2$ Seprated by amount $\Delta f$. ⇒ Corresponding voltages

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

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

The echo Signal is shifted in frequency by doppler effect

$$\Rightarrow \quad v_{1R} = \sin\left[2\pi(f_1 \pm f_{d1})t - \frac{4\pi f_1 R_0}{c} + \phi_1\right]$$

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

The receiver Seprates the two components of the echo signal and heterodynes each received Signal component with the corresponding transmitted waveform and extract the two doppler frequency components are

$$v_{1D} = \sin\left(\pm 2\pi f_d t - \frac{4\pi f_1 R_0}{c}\right)$$

$$v_{2D} = \sin\left(\pm 2\pi f_d t - \frac{4\pi f_2 R_0}{c}\right)$$

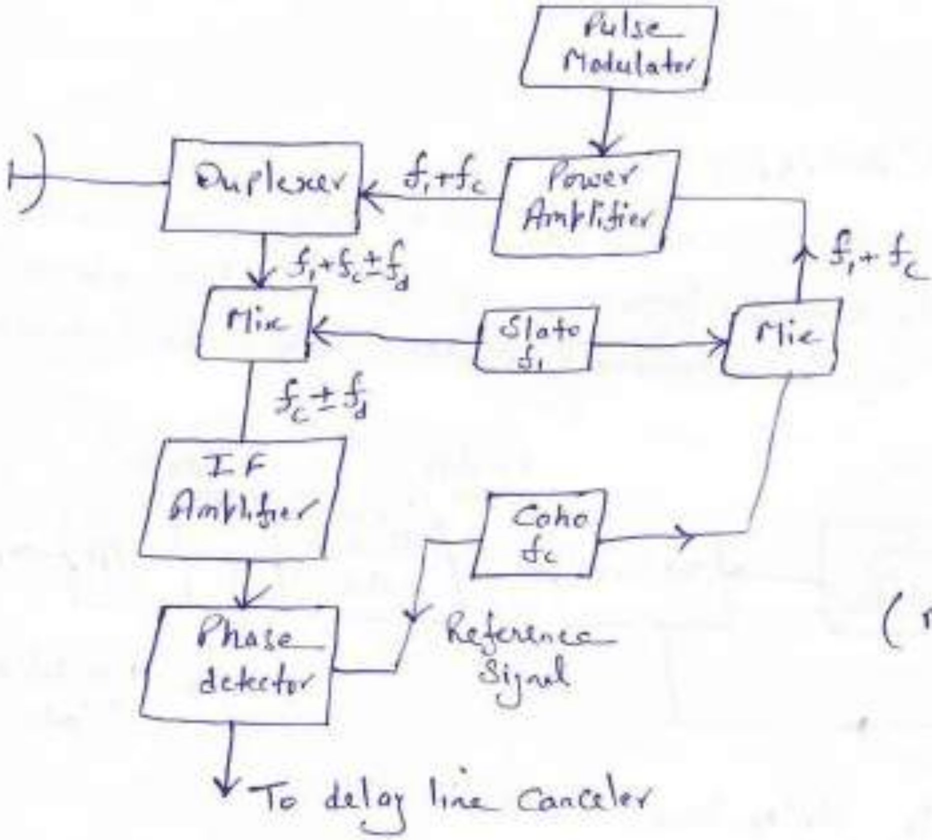
The Phase difference b/w two components is

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

$$\Rightarrow R_0 = \frac{c\,\Delta\phi}{4\pi\,\Delta f}$$

# MTI RADAR AND PULSE DOPPLER RADAR:-

A pulse radar that employs the doppler shift for detecting moving targets is either an MTI (moving target indication) radar or pulse doppler radar. The MTI radar has a pulse repetition frequency (prf) low enough to not have any range ambiguities as in $R_{un} = \frac{CT_r}{2} = \frac{C}{2f}$



(MTI Radar)

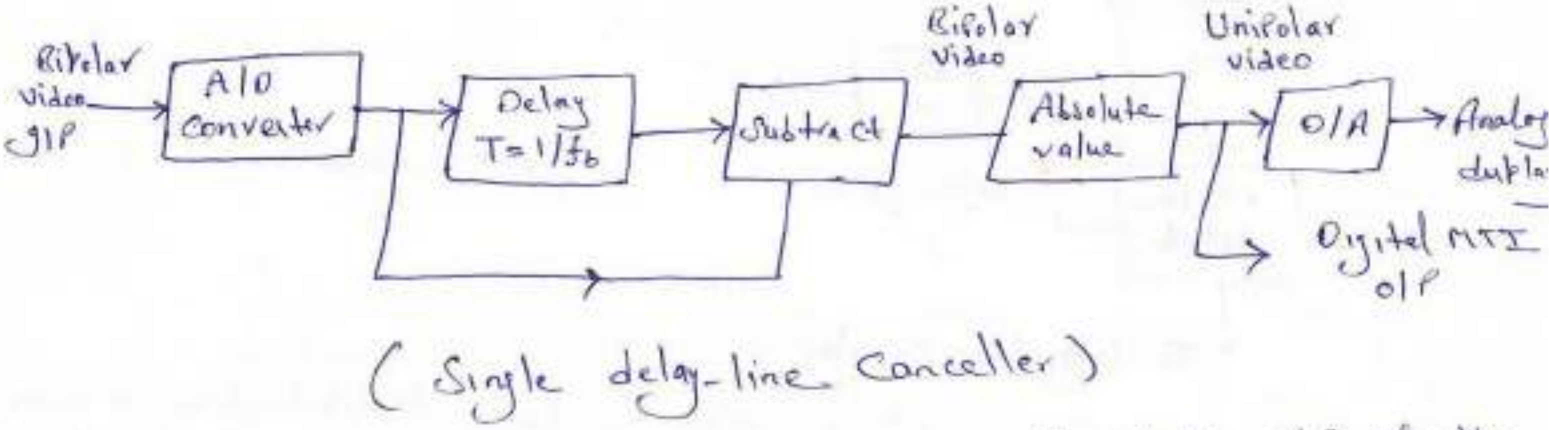
The local oscillator of an MTI radar's superheterodyne receiver must be more stable than the local oscillator for a radar that does not employ doppler. To recognize the need for high stability, the local oscillator of an MTI receiver is called stalo, which stand for stable local oscillator. The IF stage is designed as the matched filter, as is usually the case in radar. Instead of amplitude detector, there is a phase detector following the IF stage. This is a mixer like device that combines the received signal and reference signal from coho. The name coho stands for coherent oscillator to signify that the reference signal that has phase of the transmitter signal.

coherency with the transmitted Signal is obtained by using the Sum of the coho and the Stalo Signals as the I/P Signal to Power amplifier.

The Power amplifier is a good transmitter for MTI radar Sinc it can have high Stability and is capable of high Power. The pulse modulator turns the amplifier on and off to genera the radar pulses.

## DELAY - LINE CANCELERS :-

Simple MTI delay line canceler (DLC) of previously defined figure is an Example of a time domain filter that rejects Stationary clutter at zero frequency.



( Single delay-line Canceller)

The Signal from a target at range $R_0$ at the O/P of the phase detector can be written

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

we can write delayed version of $V_1$

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

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

The frequency response func^n of the single delay-line Canceler is then

$$H(f) = 2\sin(\pi f_d T_p) \quad \text{— ①}$$

Blind Speeds:- The response of the Single delay line canceler wil be zero whenever the magnitude of $\sin(\pi f_d T_p)$ in ① is zero.

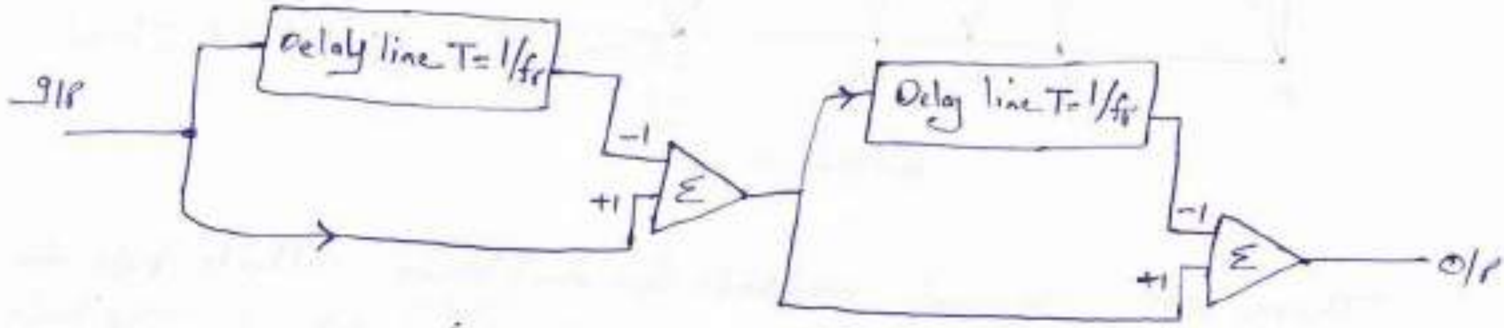
which occurs when $\pi f_d T_p = 0, \pm\pi, \pm 2\pi \ldots$

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

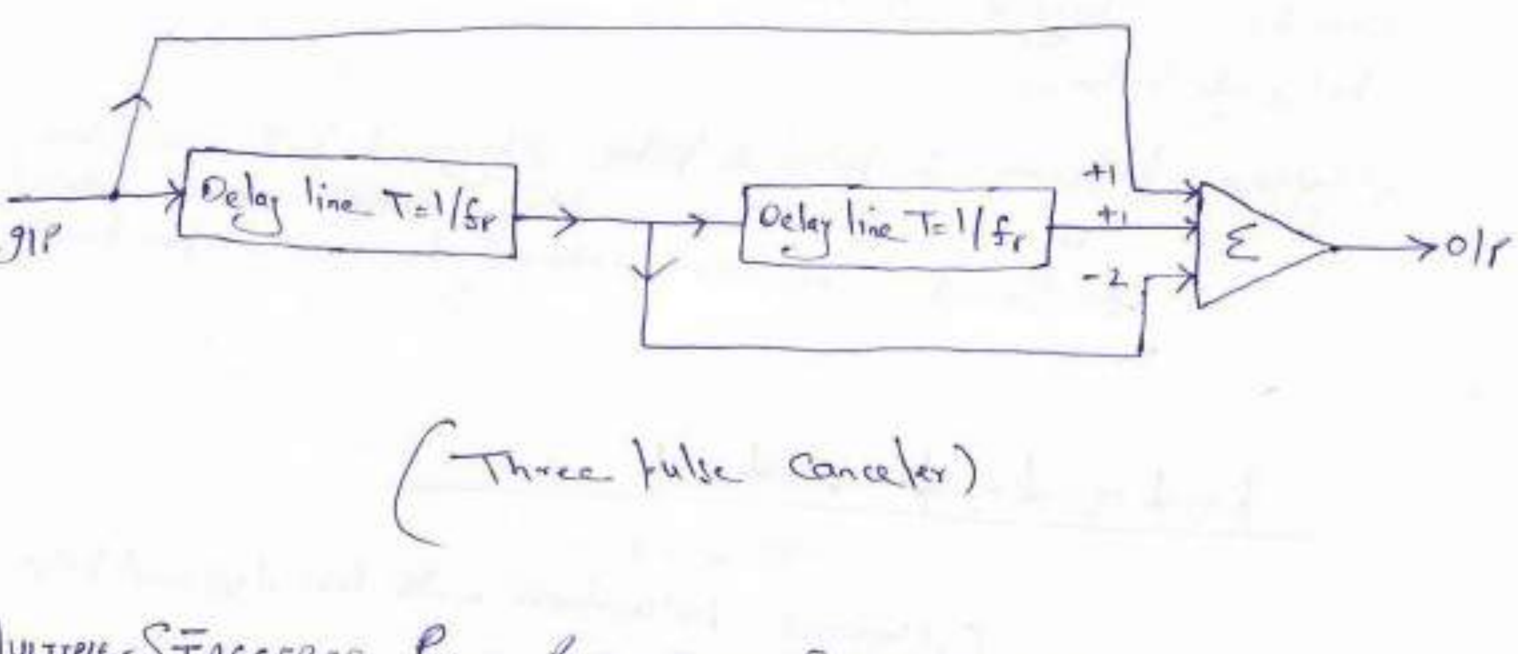
The radial velocities that produce blind Speed are

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

where $v_r$ has been replaced by $v_n$, the nth blind Speed.



(Double delay line canceler)

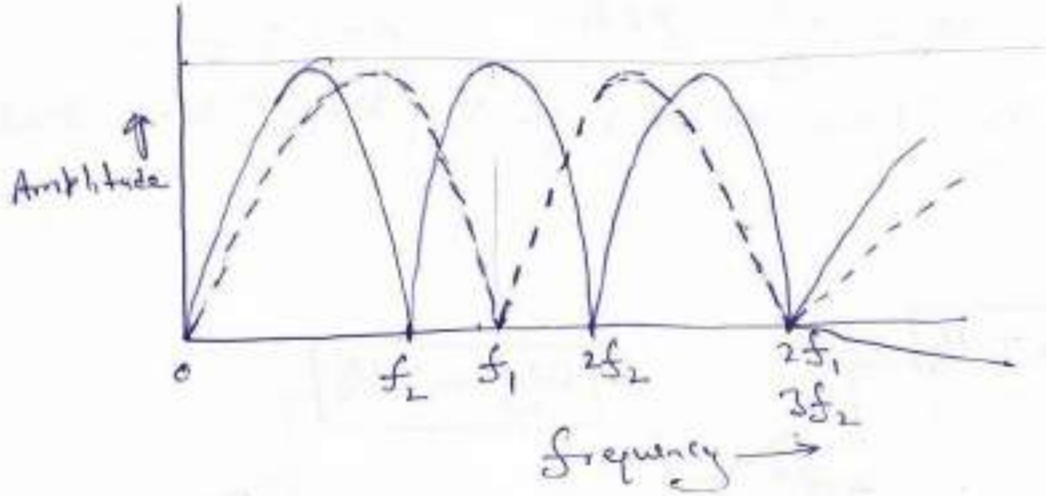


(Three pulse canceler)

## MULTIPLE, STAGGERED PULSE REPETITION FREQUENCIES

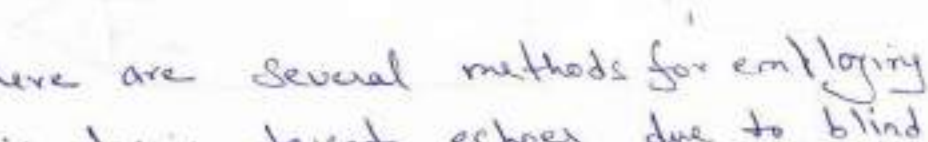
The use of multiple waveforms with different pulse repetition frequencies allow the detection of moving targets that would otherwise be eliminated with a constant-prf waveform if their radial velocities were at, or in vicinity of, a blind Speed ($v_n = \frac{n\lambda f_r}{2}$). As shown in figure frequency response of of a single delay line canceler with two different pulse repetition frequencies. At prf $f_1$

blind speeds (nulls) occur when the doppler frequency is $f_1$ or $2f_1$. with prf $f_2 = 2f_1/3$ blind speed occur when the doppler frequency equals $f_2$, $2f_2$ or $3f_2$.



There are several methods for employing multiple prfs to avoid losing target echoes due to blind speeds. The prf can be changed 1) scan to scan ② dwell to dwell ③ pulse to pulse.

Staggered prfs. → In pulse to pulse staggered prfs as shown in fig, the time b/w pulses is an interval or period. The term interval is more appropriate



(Staggered pulse train with four different pulse period)

# RANGE GATED DOPPLER FILTERS

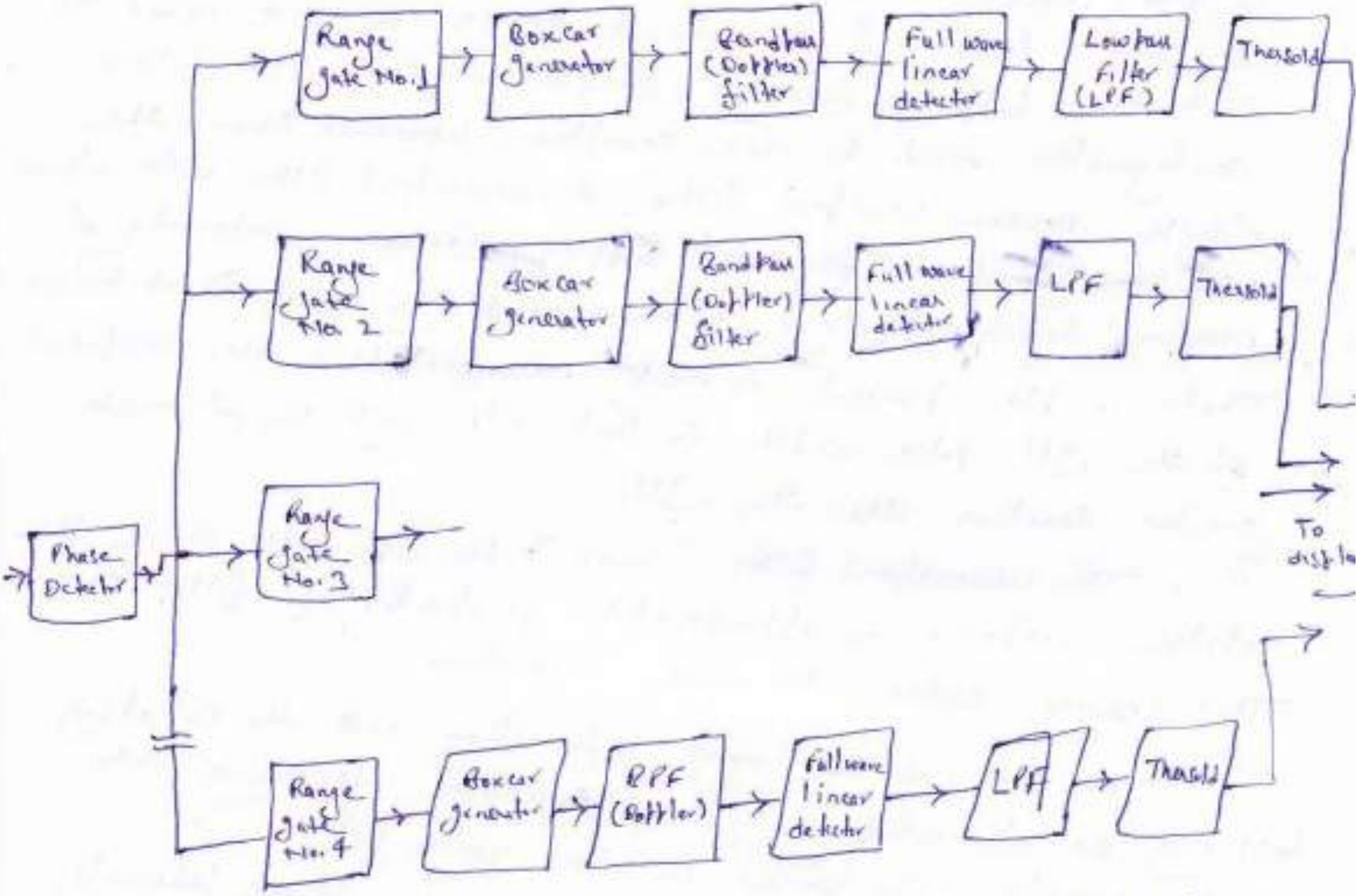
The delay line Canceler, which can be considere as a time-domain filter, has widely used in MTI radar as th mean for seprating moving targets from Stationary clutter. It is also possible to employ the more usual frequency domain bandpass filters of conventional design in MTI radar to sort the doppler-frequency-shifted targets. The filter configuration must be more complex, however than the single narrow bandpass filter. A narrowband filter with a pa-band designed to pass the doppler frequency components of moving target will "ring" when excited by usual short pulse That is its passband is much narrower than the reciproca of the I/P pulse width so that o/p will be of much greater duration than the I/P.

The narrowband filter "Smears" the I/P pulse since the impulse response is approximately reciprocal of filter B.W. This searing destroys the range resolution.

The loss of the range information and the collapsing loss may be eliminated by first quantizing the range into small intervals. The process is called range gatting. Once the radar return is quantized into range intervals, the O/P from gate may be applied to a narrow band filter since the pulse shape need no longer be preserved for range resolution. A collapsing loss does not take place since noise from other range intervals is excluded.

A block diagram of the video of an MTI radar with multiple range gates followed by clutter-rejection filter shown in Fig. The O/P of phase detector is sampled sequentially by the range gates. Each range gate opens in sequen just long enough to sample range gate acts as a switch or gate which opens and closes at proper time.

An echo from a moving target produces a series of pulses which vary in amplitude according to the doppler frequency.



(MTI radar Using range gates and filter)

The O/P of range gates is Strecked in a ckt called the boxcar generator or sample and hold ckt, whose purpose is to aid in the filtering and detection process and by emphasizing the fundamental of the modulation frequency and eliminate harmonics of prf.

# RANGE GATED DOPPLER FILTERS

The delay line canceler, which can be considered as a time-domain filter, has widely used in MTI radar as the mean for seprating moving targets from stationary clutter. It is also possible to employ the more usual frequency domain bandpass filters of conventional design in MTI radar to sort the doppler-frequency-shifted targets. The filter configuration must be more complex, however than the single narrow bandpass filter. A narrowband filter with a pass-band designed to pass the doppler frequency components of moving target will "ring" when excited by usual short pulses. That is its passband is much narrower than the reciprocal of the I/P pulse width so that O/P will be of much greater duration than the I/P.

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# LIMITATIONS TO MTI PERFORMANCE

The limitations to MTI performance to be cause the clutter spectrum to widen. More clutter energy is then passe by the doppler filter, which lower the improvement factor if the clutter power spectral density can be expressed as a gaussian func^n with a standard deviation $\sigma_c$ in $H_z$, i can be represented by

$$W(f) = W_0 \exp\left(-\frac{f^2}{2\sigma_c^2}\right)$$

Antenna Scanning Modulation:— The frequency spectrum has a B.W inversely proportional to the time duration to. Consequently, even if the clutter scatters were prefectly stationary and there were no instability in the radar equipment, there would still be a finite spectral spread due to the finite duration of echo signal. This is called antenna scanning modulation, this basically due to finite time on target. The longer the time on the target the less will be the spread in the clutter spectrum.

System Instabilities:→ Changes in the amplitude, frequency, or phase of the stalo or coho oscillators as well as changes in the pulse to pulse characteristics of the transmitted signal or error in the timing can result in uncanceled clutter echoes and cause a limit to the improvement factor that can be achieved.

Amplitude Changes:- If the single delay line canceler, has amplitude of the first pulse received from a stationary clutter scatterer is A and the second pulse is A+ΔA, the voltage o/p of the delay-line canceler is ΔA. ⇒ Clutter attenuation $= \frac{(\Delta A)^2}{A^2}$ and the improvement factor is twice of this.

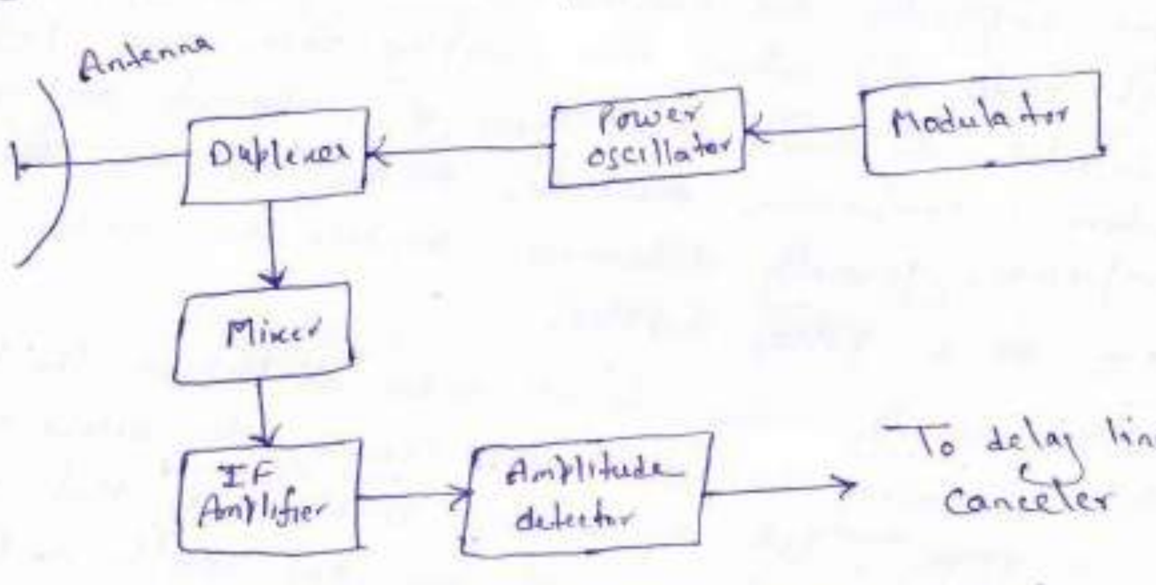
Phase changes:- If the echo received from the first pu from Stationary Clutter is represented by $A \sin(\omega t+\phi)$ a if the echo from the second pulse is $A \sin(\omega t+\phi+\Delta\phi$ there will be an uncanceled residue from a single de line canceler equal to the difference, $2A \sin(\Delta\phi/2)$, For small ph where $\Delta\phi$ → phase changes b/w pulses. changes, the o/p voltage is $A\Delta\phi$.

Phase Noise:- Noise due to phase fluctuations associat with the Stalo and coho oscillators can be a major limitation to the improvement factor of high performance MTI radar. Generally, phase noise has much larger effect than noise caused by amplitude instab -ties. The phase noise from oscillators in the exciter a power amplifier affect the transmitted signal as well as the signal in the receiver.

Limiting in MTI Radar:- clutter echoes often can be larger enough to saturate the radar receiver, obscure target echoes on a display, and cause false alarm. Saturation of the receiver by the clutter echoes also resul in spreading of the clutter spectrum that reduces the improvement factor. if the receiver is of large enough dynamic range, and there are sufficient bit in the A/D converter, and if the improvement factor is large enough to make the uncanceled clutter residue Smaller than receiver noise, there will be no problem Since there will be no limiting. larger dynamic rang and cancellation of all the large clutter, however, are the usual situation. A limiter in the MTI receiver has some times been used to reduce the clutter to the level of receiver noise.

# NONCOHERENT MTI

In an MTI radar there must be a reference signal to recognize that the echo signal of a moving target is shifted in frequency by the doppler effect. The echo signal from clutter also has the characteristics of the transmitted signal and can be used as a reference to extract the doppler frequency shift of the target echo signal. Since the clutter echo and the moving target echo appear together at the i/p to the receiver, an internal reference signal is not needed. A radar that uses the clutter echo as the reference signal to extract the doppler-shifted target echo is known as a noncoherent MTI radar.



(Noncoherent MTI radar)

The advantage of noncoherent MTI (~~application~~) is its relative simplicity. It was used in the past for both land-based and airborne MTI applications. A limitation is that it requires that clutter echo be presented along with target echo.

# PULSE DOPPLER RADAR

Pulse radar that extracts the doppler frequency shift for the purpose of detecting moving targets in the presence of clutter is either an MTI radar or a pulse doppler radar. The distinction b/w them is based on the fact that in a Sampled measurement System like a pulse radar, ambiguitie can arise in both the doppler frequency and the range measurements. Range ambiguities are avoided with a low Sampling rate, and range measurement.

Range ambiguities are avoided with a low sampling rate and dopple frequency ambiguities are avoided with a high sampling rate, Howe in most radar applications the sampling rate, or prf can't be selected to avoid both types of measurement ambiguities. Therefore a compromise must be made and the ~~the~~ nature of compromise generally determines wether the radar is calle an MTI or a pulse doppler.

MTI usually refers to a radar in which the pulse repetition frequency is chosen low enough to avoid ambiguities in range, but with the consequence that the frequency measurement is ambiguous and result in blind Speed.

The pulse doppler radar is more likely to use range-gate doppler filter banks than delay line- cancellers. Also a, power amplifier such as a klystron is more likely to be used than a power oscillator like the magnetron. A pulse dopler radar operates at a higher duty cycle than does an MTI.

# MTI FROM A MOVING PLATFORM

When a radar is in motion, as when mounted on a ship or an aircraft, the detection of a moving target in the presence of clutter is either an MTI radar or a pulse doppler radar. But it is not easy to detect the presence of target in such situation. The doppler frequency shift of the clutter is no longer at dc. It varies with speed of the radar platform, the direction of antenna in azimuth and the angle of elevation to the clutter. Thus the clutter rejection notch needed to cancel clutter can't be fixed, but must vary. The design of MTI is more difficult with airborn radar than a ship borne radar because the higher speeds and the greater range of elevation angles result in a greater variation of the clutter spectrum.

In addition to shifting the center frequency of the clutter, its spectrum is also widened. An approximate measure of the spectrum width can be found by taking the differential of the doppler frequency

$f_d = 2(v/\lambda)\cos\theta$

$$\Rightarrow \Delta f_d = \frac{2v}{\lambda}\sin\theta\,\Delta\theta \quad -①$$

$v \rightarrow$ plateform speed

Compensation for clutter doppler shift:- Two methods are used for compensation. In one implementation the frequency of coho is changed to compensate for the shift in clutter doppler frequency.

Compensation in some cases can also be made open loop by using the a priori knowledge of the velocity of the platform carrying the radar and the direction of the antenna pointing.

## Compensation for clutter doppler spread:—

As shown in eq$^n$ ① the spread in the clutter spectrum is a func$^n$ of angle θ b/w the velocity vector of the moving platform and the antenna beam-pointing direction.

Compensation of clutter doppler spread is given by DPCA (Displaced Phase Center Antenna).

## OTHER MTI DELAY LINES : →

MTI radar used acoustic delay lines in which EM signals were converted into acoustic waves. The acoustic signals were delayed, and then converted back into EM signals. The process was lossy (50 to 70 dB) of limited dynamic range, and spurious responses were generated that could be confused for legitimate echoes. Since acoustic waves travel with a speed about $10^{-5}$ that of EM waves, an acoustic line can be of practical size whereas an EM delay is not. However acoustic delay lines are larger ~~than~~ and heavier than digital lines and must usually be kept in a temp$^r$-controlled environment to prevent unwanted changes in delay time.

# UNIT III

# MOVING TARGET INDICATION AND PULSE DOPPLER RADAR

# TRACKING WITH RADAR

A radar not only recognizes the presence of target, but it determines the target's location in the range and in one or two angle co-ordinate. As it continues to observe a target over a time, the radar can provide the target's trajectory. There are four types of radar that can provide the tracks of targets.

1) Single target tracker:- This tracker is designed to continuously track a single target at a relatively rapid data rate. The data rate, of course, depends on the amplification, but 10 observations per second might be "typical" of a military guided missile weapon control radar.

The antenna beam of a single target tracker follows the target by obtaining an angle error signal and employing a (Single target tracker) closed loop servo system to keep error signal small.

2) Automatic detection and track (ADT) This performs tracking as part of an air surveillance radar. It is found in almost all modern civil air-traffic control radars as well as military air surveillance radar. The rate at which observations are made depends on the time for the antenna to make a rotation. The ADT, therefore, has a lower data rate than that of STT, but its advantage is that it can simultaneously track a large no. of targets.

3) Phased array radar tracking:- A large no. of targets can be held in track with a high data rate by an electronically steered phased array radar. Multiple targets are tracked

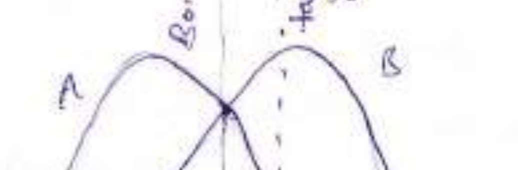
on a time shared basis under computer control since the beam of an electronically scanned array can be rapidly switche from one angular direction to another, sometimes in a few micro seconds.

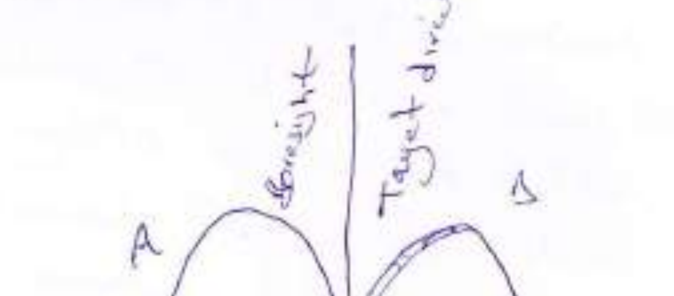
4) Track while Scan (TWS):-

This radar rapidly scans a limited angular sector to maintain tracks, while a modera data rate, on more than one target within the coverage of the antenna. It has been used in past for air defense radars, aircraft landing radars, and in some airborne intercept radars to hold multiple target in track.

Angle Tracking:- In an simple pencil-beam radar the detection of a target provides its location in angle as bein somewhere within the antenna beamwidth; but more inform is needed to determine the direction the antenna shoul be moved to maintain the target within its beam.

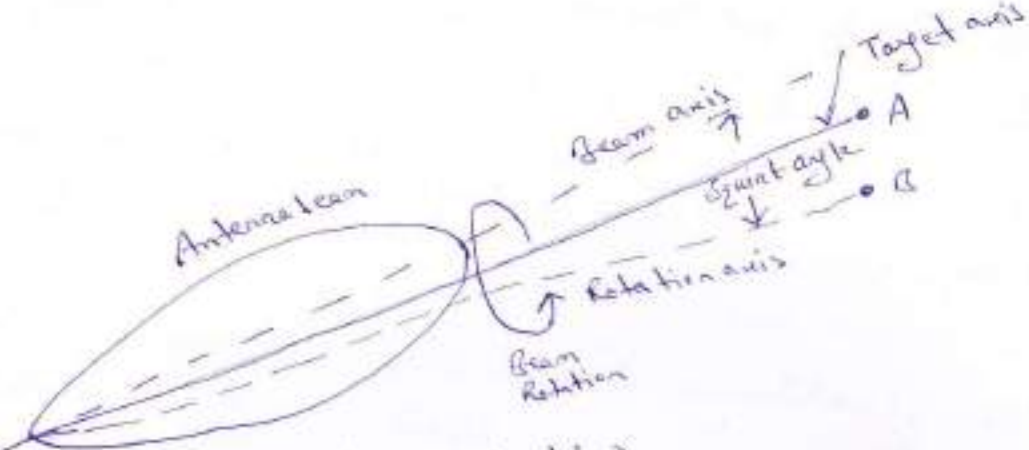
- In order to determine the direction in which the anten beam needs to be moved, a measurement has to be mad at two different beam position.





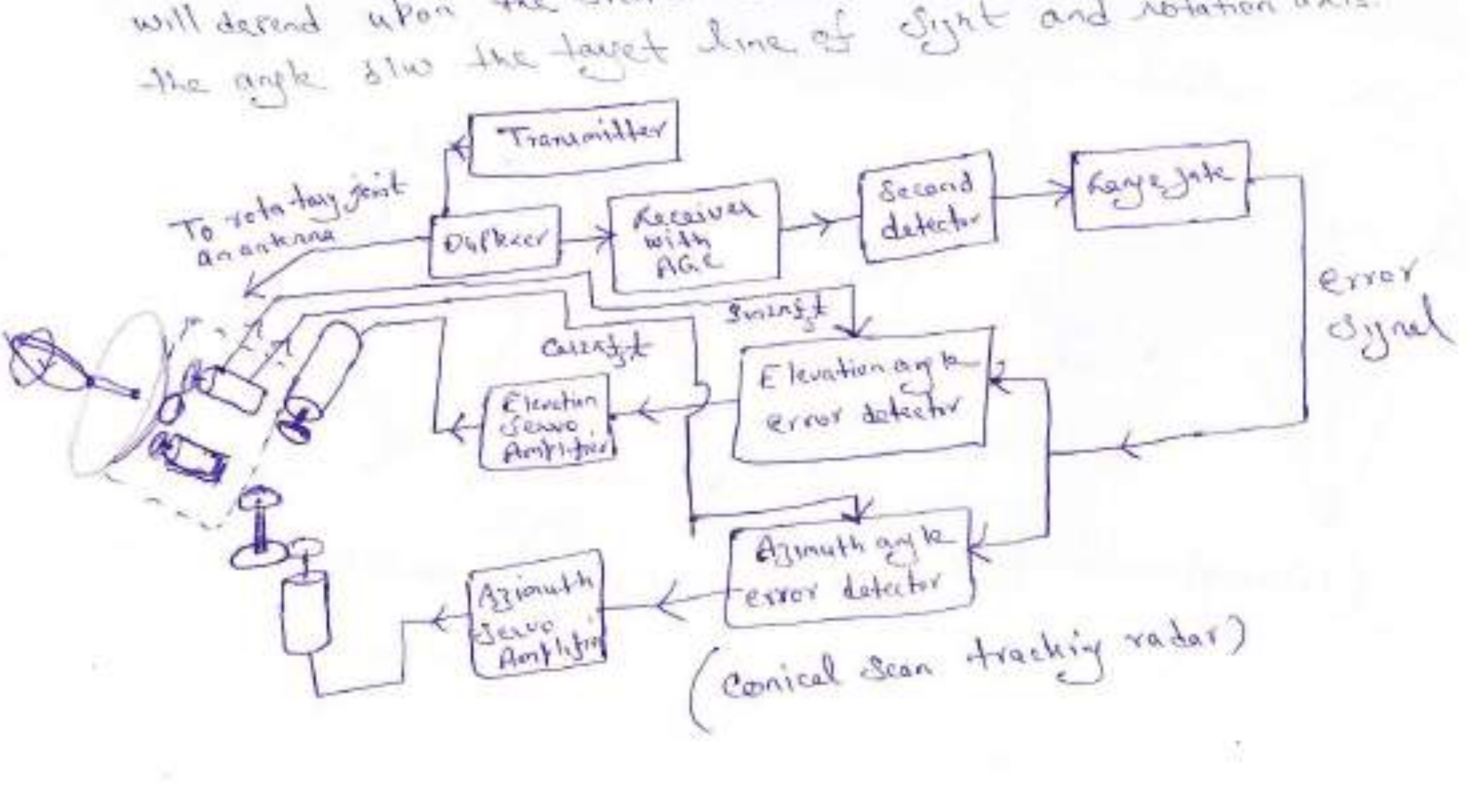
# CONICAL SCAN

The logical extention of the Simultaneous lobing techniques is to rotate continuously an offset antenna beam rather than discontinuously step the beam b/w the four discrete positions. This is known as Conical Scanning.



(conical scan tracking)

The angle b/w the axis of rotation which is usually but not always the axis of the antenna reflector and the axis of the antenna beam is called squint angle. Consider a target at position A. The echo signal will be modulated at a frequency equal to rotation frequency of the beam. The amplitude of the echo signal modulation will depend upon the shape of antenna pattern, the squint angle, and the angle b/w the target line of sight and rotation axis.

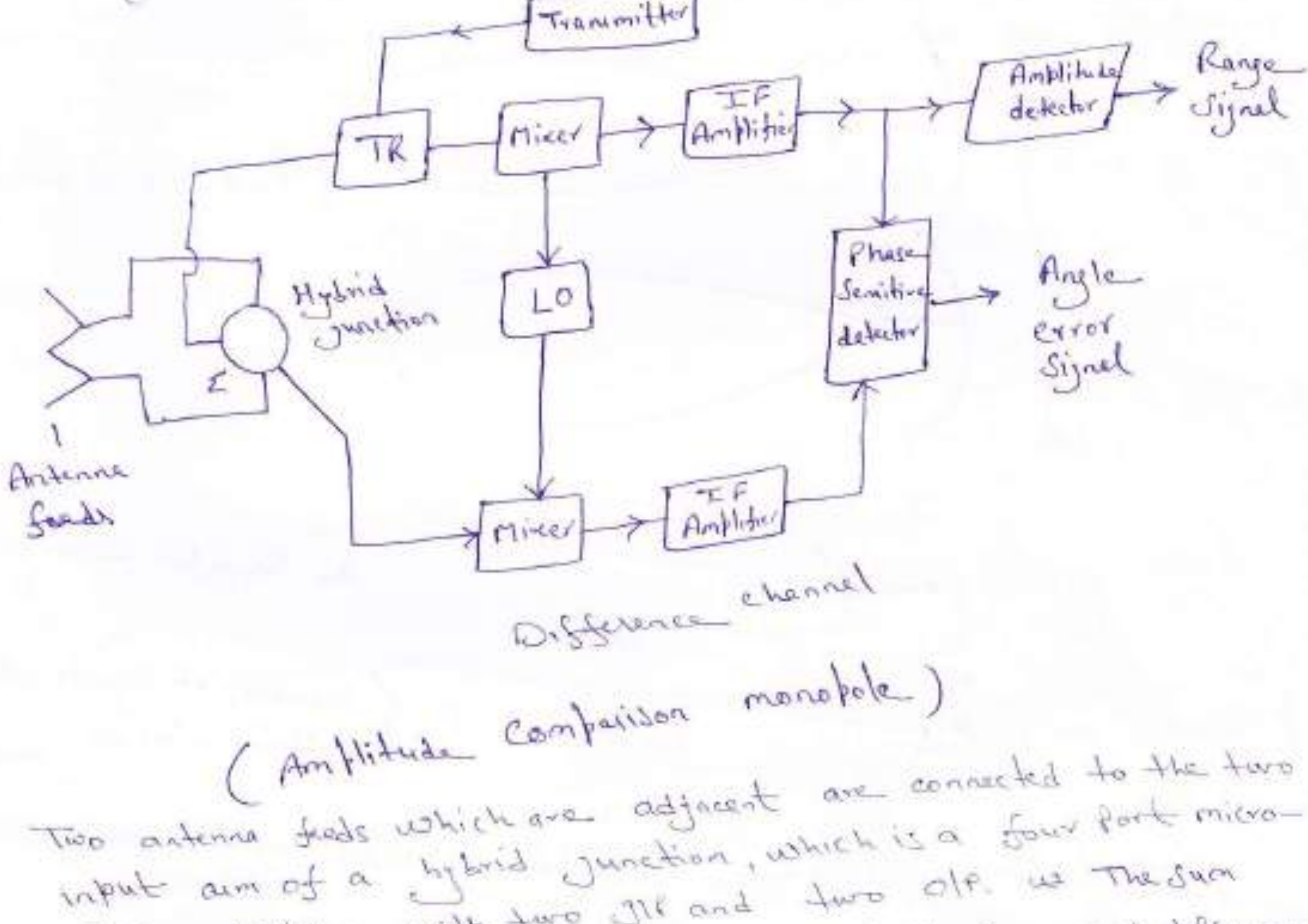


(Conical scan tracking radar)

# MONOPULSE TRACKING

A monopole tracker is defined as one in which information concerning the angular location of a target is obtained by comparison of signals received in two or more simultaneous beams. A measurement of angle may be made on the basis of a single pulse, hence the name monopulse.

There are several methods by which a monopulse angle measurement can be made. The most popular by far has been the amplitude-comparison monopulse which compares the amplitudes of the signals simultaneously received in multiple squinted beams to determine the angle.



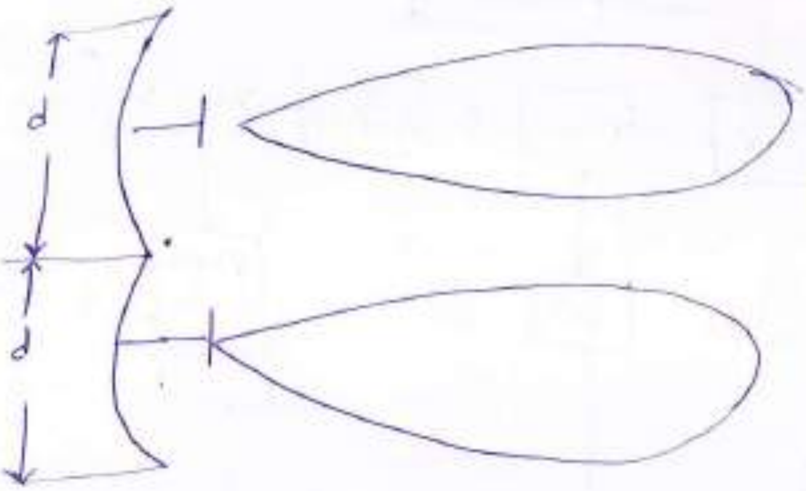
( Amplitude Comparison monopole )

Two antenna feeds which are adjacent are connected to the two input arm of a hybrid junction, which is a four port micro-wave device with two I/P and two O/P. ut The sum pattern is used for transmission while both sum and difference patterns are used for reception. The signal received with difference pattern provide the magnitude and angle error.

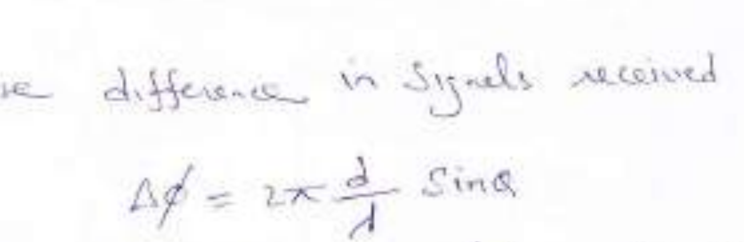
The sign of the difference signal is determined by comparing the phase of the difference signal with the phase of the sum signal.
]

## Phase Comparison monopulse :→

In this two antenna beams are used to obtain an angle measurement in one co-ordinat just as in amplitude comparision monopulse. The two bea however, look in the same direction and cover the same region of space rather than be squinted to look in two slightly different directions. In order for the two beams to look in same direction, two antenn have to be used in the phase comparison monopulse.



(Two antenna radiati



{Geometry of Signals at two antennas

Phase difference in Signals received in the two anten is

$$\Delta\phi = 2\pi \frac{d}{\lambda} \sin\theta$$

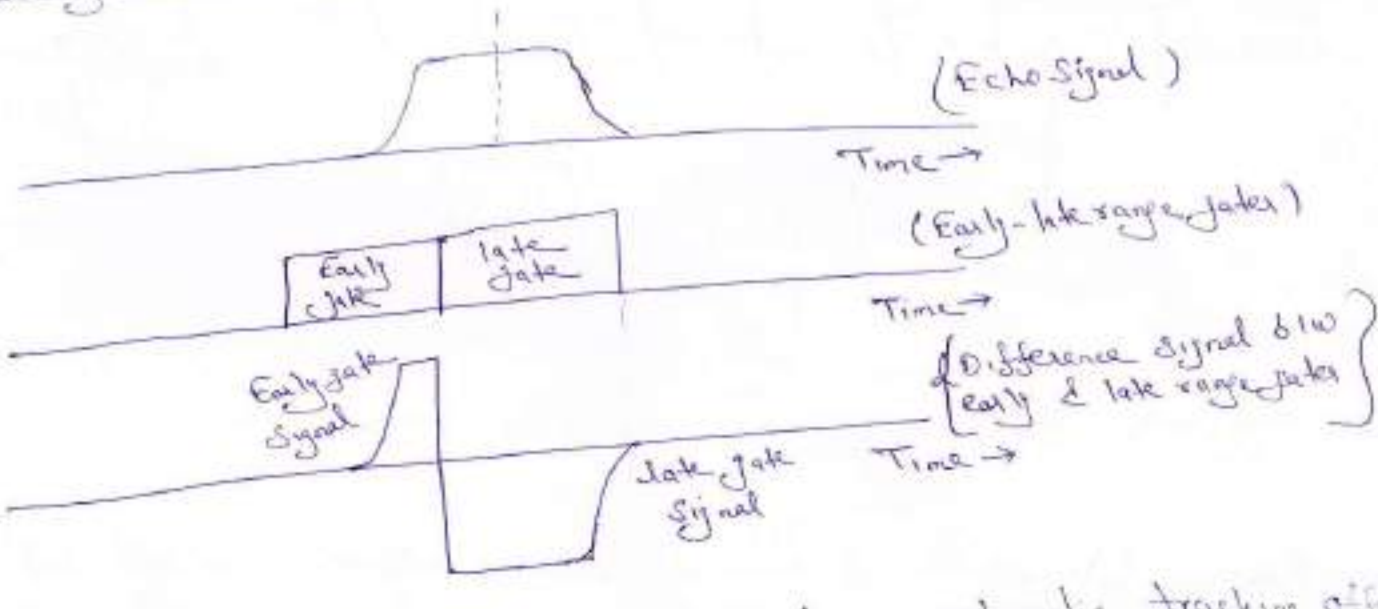
The phase - comparison monopulse is sometimes known as an interferometer radar.

# 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 or J-scope representation and manually positions a handwheel in order to maintain a marker over the desired target pip. The setting of handwheel is a measure of the target range and may be converted to a voltage that is supplied to a data processor.

As the target speeds increases it is increasingly difficult for an operator to perform at the necessary level of efficiency over a sustained period of time, and automatic tracking become a necessity.

The technique for automatically tracking in range is based on the Split range gate. Two range gates are generated as shown in Figure

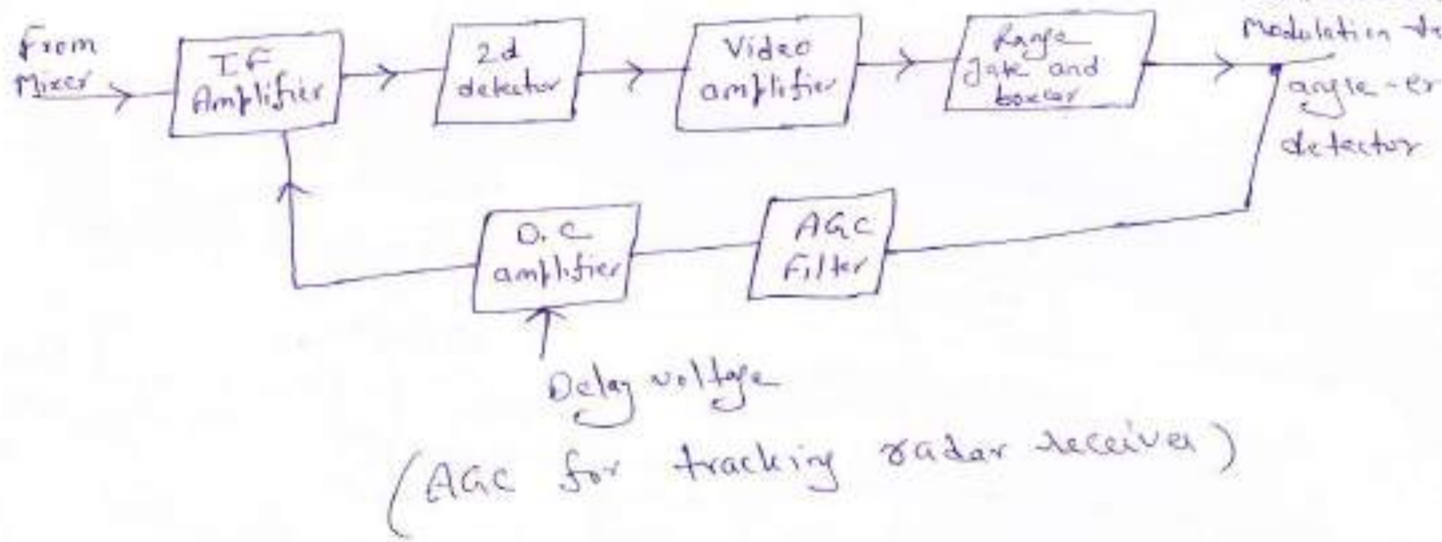


The range gating necessary to perform automatic tracking offe several advantages as by products. it isolates one target, excluding targets at other ranges. This permits the boxcar generator to be employed.

one of the simplest conical scan antenna is a parabola w an offset rear feed rotated about the axis of reflector. if the feed maintains the plane of polarization fixed as it rotates

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.

Automatic gain control: → The echo signal amplitude at the tracking radar receiver will not be constant but will vary with time. The function of AGC is to maintain the d.c level of the o/p of receiver constant and to smooth or eliminate as much of the noiselike amplitude fluctuat as possible without disturbing the extraction of desired err signal at the conical scan frequency.



(AGC for tracking radar receiver)

Range Glint: → A target with multiple scatters distributed in range can cause tracking errors because of glint

$$\Delta T_R = \frac{\Delta T}{2} \cdot \frac{1-a^2}{1+a^2+2a\cos(2\pi f_0 \Delta T)}$$

# ACQUISITION

A tracking radar must first find and aquire (look on it) its target before it can operate as a tracker. Most tracking radars employ a narrow pencil beam for accurate tracking in angle but it can be difficult to search a large volume for targets when using a narrow antenna beamwidth. Some other radar, therefore must first find the target to be tracked and then designate the target's co-ordinates to the tracker. These radars has been called acquisition radars or designation radar.

The tracker is slewed to the direction of target based on the target co-ordinates supplies by the aquisition radar. These co-ordinate are always accurate enough to bring the tracker directly onto the target. Some searching in both azimuth and elevation angle might have to be done by the tracker in order to find a target. There have been several different types of patterns employed to search a limited angular region

If a 2D air-surveillance radar (range and azimuth) is used for designing a target to a surface based mechanical tracking radar, the tracker might acquire its target with nodding beam scan in elevation, which is raster scan in vertical rather than horizontal.

The target must be find in range as well as in angle. During the aquisition process, the tracking radar receiver range gate is scanned in range as well as the pulse propagate outwards in space.

# UNIT IV

## TRACKING RADAR AND RADARDETECTION THEORY

# RADAR RECEIVER

The function of the receiver in early radar System was to Extract the weak echo Signals that appeared at the antenna terminals and amplify them to a level where they could be displayed to a radar operator who then make the decision as to amplify whether or not a target echo Signal present. It employs a matched filter whose purpose is to maximize the Peak Signal to mean noise ratio and discriminate against unwanted Signals whose waveform are different from those transmitted by the radar. In modan radars the decision wether a target is present or absent is seldom made by an operator viewing on a display the unprocessed output of a receiver. Information about a target's location in range and angle can also be extracted automatically instead of manually by an operator. In an operational air surveillance radar, tracking of targets is no longer performed by an operator making with a grease pencil on a radar display the location of blips (target) from Scan to scan and calculating the target sp and estimating its direction.

When a radar can not remove all the clutter, echoes, Cont -ant false alarm rate (CFAR) circuitry is employed to prevent the tracking computer from becoming overloade when trying to establish tracker using clutter echoes.

Thus in addition to detection and amplification of Signals a radar receiver perfoms many other functions either directly as a part of receives or in conjunction with it. Ther other functions include Signal processing, information extraction, data processing, electromagnetic compatibility a

Electronic Counter-Countermeter (The modern receiver mignt thought of as the receiver processor). Sometimes the dipla is Considered part of receiver System.

The radar receiver is almost always a Superheterodyne o Superhet. The essential Characteristic of superheterodyne is that it converts the RF input signal to an intermediate frequency where it is easier than at RF to acheive the necessary filter shape, bandwidth, gain, and Stability.

The first Stage, or front end, of a radar Superheterodyne receiver can be an RF low-noise amplifier such as a transist

Before the avalibility of low-noise transistors, the receiver front end was the mixer Stage without an RF amplifier preceding.

# NOISE FIGURE

The noise figure of a linear network may be defined as either

$$F_n = \frac{N_{out}}{KT_0 B_n G} \quad \text{or} \quad \frac{S_{in}/N_{in}}{S_{out}/N_{out}} \qquad -①$$

where $N_{out}$ → Available O/P noise power

$KT_0B_0 = N_{in}$ → Available I/P noise power

$K$ ⇒ Boltzmann's constant $= 1.38\times10^{-23}$ J/deg

$T_0$ → Standard temprature of 290 K

$G = S_{out}/S_{in}$ = Available gain

$S_{out}$ → Available O/P Signal Power

$S_{in}$ → Available I/P Signal Power

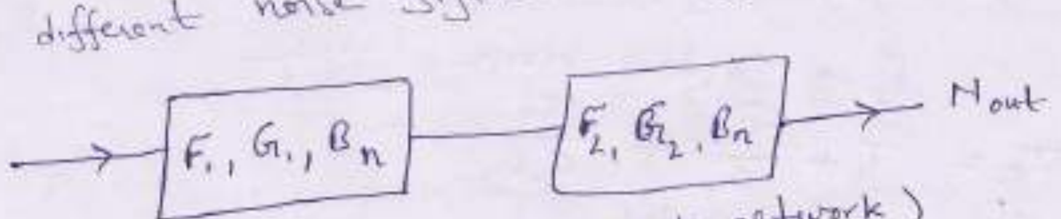
— The term avilable power refers to the power that would be delivered to a matched load.

Eqⁿ ① permits two different, but equivalent, interpretation of the noise figure. It may be considered as the degradation of the signal to noise ratio as the signal passes through the network.

$$\Rightarrow F_n = \frac{KT_0B_nG + \Delta N}{KT_0B_nG} = 1 + \frac{\Delta N}{KT_0B_nG}$$

Noise figure of network in cascade :—

Considered two networks in cascade, each with the same noise bandwidth $B_n$, but with different noise figure and gain.



( Cascade network )

The problem is to find $F_o$, the overall noise-figure of the two networks in cascade.

The output noise $N_{out}$ of the two network in cascade is

$N_{out} =$ noise from network 1 at o/p of network 2 + noise $\Delta N_2$ intro-duced by network 2.

$$N_{out} = F_o K T_o B_n G_1 G_2 = F_1 K T_o B_n G_1 G_2 + \Delta N_2 \Rightarrow$$

$$N_{out} = F_1 K T_o B_n G_1 G_2 + (F_2 - 1) K T_o B_n G_2$$

$$\Rightarrow F_o = F_1 + \frac{F_2 + 1}{G_1}$$

The noise figure of N networks in cascade may be given b

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

Noise temprature:— The noise introduced by a network ma also be Exprened as the effective noise temprature $T_e$,

It is given $\Rightarrow$ $\Delta N = K T_e B_n G$

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

$$\Rightarrow T_e = (F_n - 1) T_o$$

The System noise temprature $T_s$ is defined as the effective noise temprature of the receiver including the effects of antenna temprature $T_a$. if the receiver effective noise temprature is $T_e$ then

$$T_s = T_a + T_e = (F_s - 1) T_o$$

$F_s \rightarrow$ System noise figure

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

For N- networks

# MIXERS

Many radar Superheterodyne receivers do not employ a low noise RF amplifier. Instead, the first Stage is simply the mixer. Although the noise figure of a mixer front end may not be as low as other devices that can be used as receiver front ends, it is acceptable for many radar applications when other factors besides low noise are important.

The function of the mixer is to convert RF energy to IF energy with minimum loss and without ~~Sup~~ Spurious responses. An integral part of the mixer is the local oscillator. The IF amplifier is also of importance in mixer design because of the its influence on the overall noise figure.

Conversion loss and noise temprature ratio:-

The conversion loss of a mixer is defined as

$$L_c = \frac{\text{available RF power}}{\text{available IF power}}$$

It is the measure of efficiency of the mixer in converting RF signal power into IF. The conversion loss of typical microwave crystals in a conventional ~~by~~ Single ended mixer Configuration varies from about 5 to 6.5 dB. A crystal mixer is called broadband when the signal and image frequencies are both terminated in matched loads.

Short circuiting or open circuiting the image frequency terminals results in narrow band mixer.

The noise temprature ratio of a crystal mixer is defined by

$$t_n = \frac{\text{actual available IF noise power}}{\text{available noise power from an equivalent resistance}}$$

$$t_n = \frac{F_c K T_0 B_n G_c}{K T_0 B_n} = F_c G_c = \frac{F_c}{L_c}$$

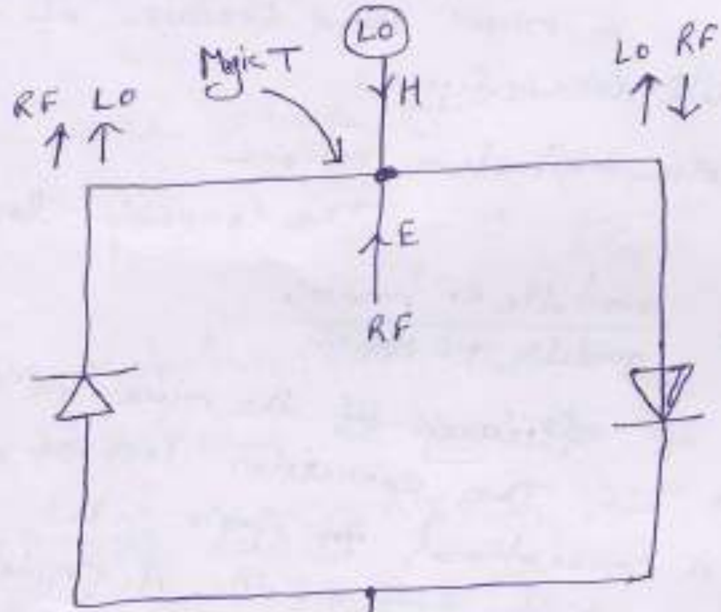
$F_c$ = crystal mixer noise figure

$L_c = 1/G_c$ = Conversion loss

Balanced mixers →

Noise that accompanies the local oscillator (LO) sign can appear at the IF frequency because of the nonlinear action of the mixer. The LO noise must be removed, if receiver sensi-vity is to be maximized. One method for eliminating LO nois that interferes with the desired signal is to insert a narrow bandpass RF filter b/w the local oscillator and the mixer.

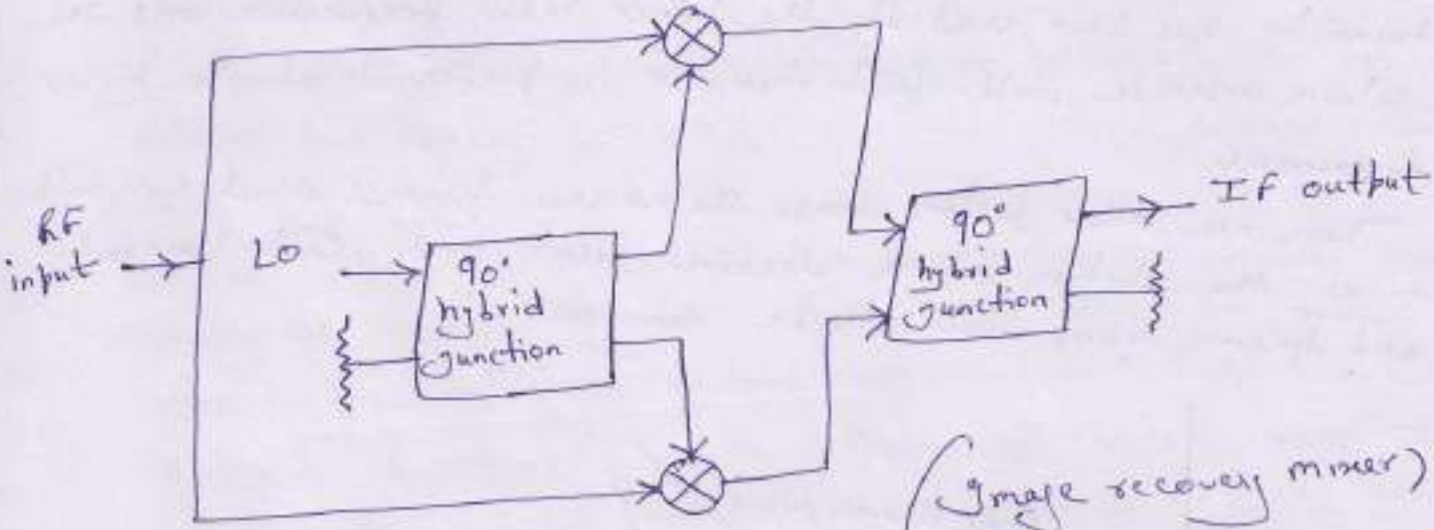
A method of eliminating local oscillator noise without the disadv-ntage of a narrowbandwidth filter is the balanced mixer. A balanced mixer uses a hybrid junction, a magic T, or an equivalen These are the four port junctions.



(Balanced mixer)

In a single ended mixer, the mixing action generates all harmon of the RF and LO frequencies, and combination thereof. The output is designed to filter out the frequency of interest, usu the difference frequency. A balanced mixer supresses the even harmonics of the LO signal. A double balanced mixer is basically two single ended mixers connected in parallel and 180° out of phase. It supreses even harmonics of both the RF and the LO signals.

Reactive image termination:- If the Image frequency of a mixer is presented with the proper reactive termination, the conversi loss and the noise figure can be 1 or 2 dB less than with a broad band mixer in which the image frequency is terminated in the matched load.



(Image recovery mixer)

A method for achieving a reactive termination without bandwidth components is the image recovery mixer.

Diode burn out:- one of the cause of diode burn out in radar receiv -er has been the increased RF leakage through conventional duplexer due to aging of the TR tube.

Noise figure due to RF losses:- The noise figure due to RF losses may be derived from

$$F_n = \frac{N_{out}}{KT_0 B_n G}$$

The noise figure of a receiver with noise figure $F_x$, preceded by RF losses equal to $L_{RF}$ is

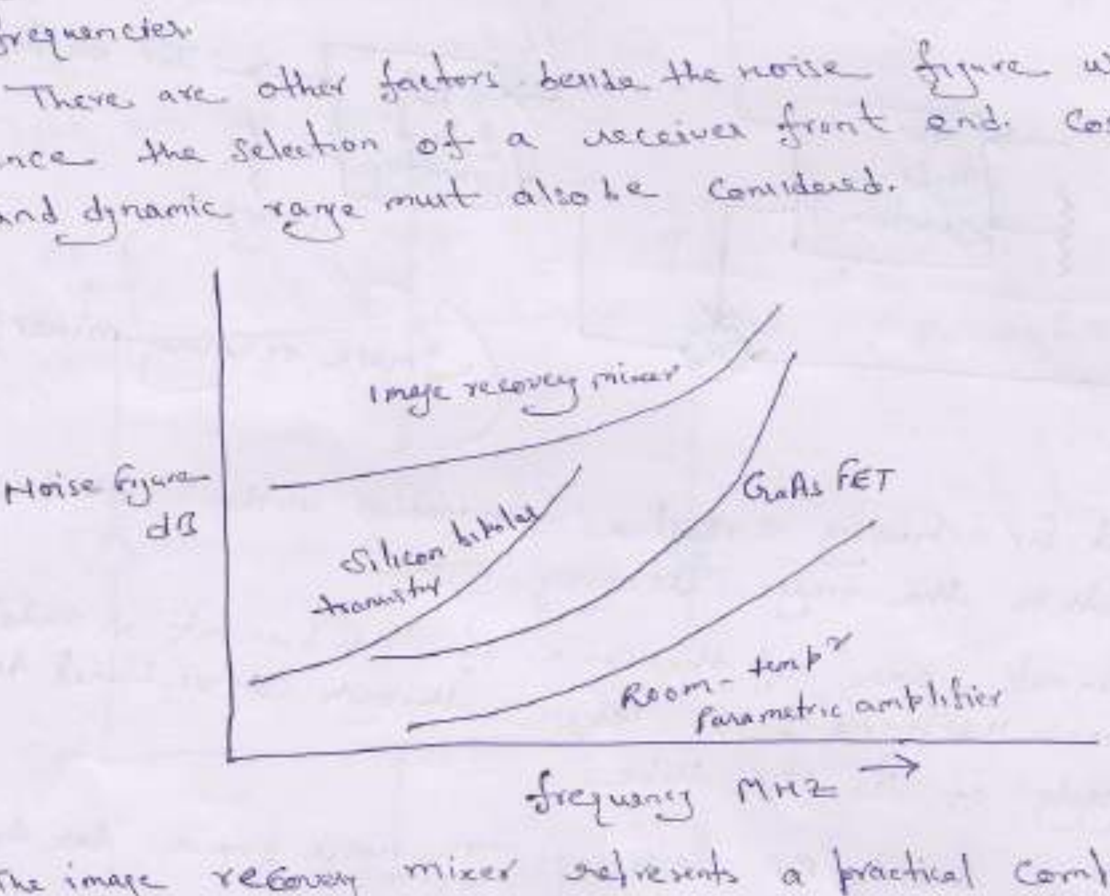
$$F_0 = F_1 + \frac{F_2 - 1}{G_1} = L_{RF} + (F_x - 1) L_{RF} = F_x L_{RF}$$

# LOW NOISE FRONT ENDS

The parametric amplifier has the lowest noise figure as compare to other amplifiers, especially at the higher microwave frequencie

The transistor amplifier can be applied over most of the entire range of frequencies of interest to radar. The silicon bipolar transistor has been used at the lower radar frequencies and the galium arsenide field-effect transistor is preferred at the higher frequencies.

There are other factors beside the noise figure which can inf -ence the selection of a receiver front end. Cost burnout, and dynamic range must also be considered.



The image recovery mixer represents a practical compromise which tends to balance its slightly greater noise figure by its lower cost, greatest ruggedness, and greater dynamic range.

The lower the noise figure of the radar receiver, the less nee be the transmitter power and/or the antenna aperture. Reduction in the size of transmitter and antenna are always desirable if there are no concomitant reductions in performance. A few decibels improvement in receiver noise figure can be obtained a a relatively low cost as compared to cost and complexity of adding the same few decibels to a high power transmitter.

# DISPLAYS: →

The purpose of display is to visually present in a form suitable for operator interpretation and action the informat contained in the radar echo signal. when the display is connected directly to video output of the receiver, the information displayed is called raw video.

When the receiver video output is first processed by an automatic detector and tracking processor (ADT), the output displayed is sometimes called synthetic video.

The cathode ray tube (CRT) has been almost universally used as the radar display. There are two basic cathode ray tube displays. one is the deflection modulated CRT, such as A scope in which a target is indicated by the deflection of electron beam. The other is the intensity modulated CRT, such as PPI in which the target is indicated by intensifying the electronic modulated dis-plays, and targets may be more readily discerned in the presence of noise or interference. On the other hand, intensity modulated displays have the advantages of presenting data in a convenient an easily interpreted form. The deflection of the beam or the appear-ance of an intensity modulated spot on a radar display caused by the presence of a target is commonly refered as a blip.

Electrostatic deflection CRT's use an E-field applied to pairs of deflection electrods, or plates, to deflect the electron beam. Such tubes are usually longer than magnetic tubes, but the overall size, weight, and power dissipation are less. Electromag-netic deflection CRT's require magnetic coils, or deflection yokes positioned around the neck of the tube.

Types of display presentation:- The various types of CRT displays which might be used for surveillance and tracking radars are defined as follow.

A-Scope:- A deflection modulated display in which the vertical deflecti is proportional to target echo strength and the horizontal co-ordina is proportional to range.

B-Scope:- An intensity modulated rectangular display with azimuth angle indicated by the horizontal co-ordinate and range by vertical co-ordinate

C-Scope:- An intensity modulated rectangular display with azimuth angle indicated by the horizontal co-ordinate and elevation angle by the vertical co-ordinate.

D-Scope:- A C-Scope in which blips extend vertically to give a rough estimate of distance.

E-Scope:- A intensity modulated rectangular display with distance indica by the horizontal co-ordinate and elevation angle by the vertical co-ordinate.

F-Scope:- A rectangular display in which a target appears as a centera -zed blip when the radar antenna is aimed it.

G-Scope:- A rectangular display in which a target appear at a laterally centralized blip when radar antenna is aimed at it in azimuth.

H-Scope:- A B-Scope modified to include indication of angle of elevation.

I-Scope:- A display in which a target appear as a complete circle when the radar antenna is pointed at it and in which the radius of the circle is proportional to target distance.

J-Scope:- A modified A-Scope in which the time base is a circle and target appear as radial deflections from the time base.

K-Scope:- A modified A-Scope in which a target appear as a pair of vertical deflection

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

M-Scope:- A type of A-Scope in which the target distance is dete -ned by moving an adjustable pedestal signal along the baseli until it coincides with horizontal position of target signal deflection.

N-Scope:- A K-Scope having adjustable pedestal signal.

O-Scope:- An A-Scope modified by the inclusion of an adjustable notch for measuring distance.

PPI:- Plane Position indicator (also called P-Scope). An intensity modulated circular display on which echo signals produced from reflecting objects, are shown in plan position with range and azimuth angle displayed in polar co-ordnate, forming a map-like display.

R-Scope:- An A-Scope with a segment of the time base expanded near the blip for greater accuracy in distance measurement.

RHI:- Range height indicator:- An intensity modulated display with height as the vertical axis and range as horizontal axis.

CRT Screens:- A number of different cathode ray tubes screens are used in radar application. we also have color CRT's that provides another dimension for the display of target information.

Bright displays:→ There are applications where it is not possible or convenient to use the conventional CRT display that requires a darkened environment; such as in cockpit of an aircraft or an air-field control tower. One form of bright display is the direct view storage tube.

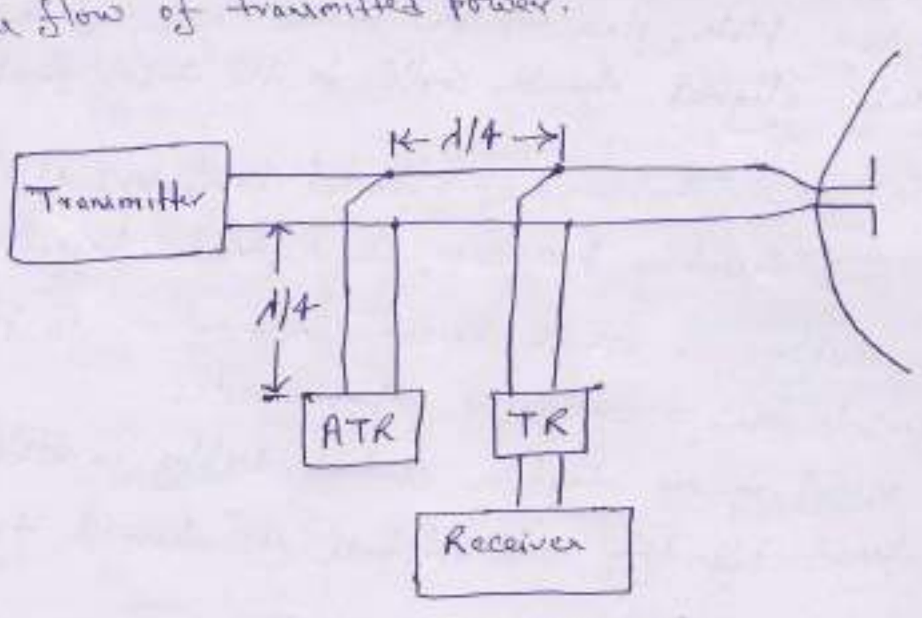
Rear Port:- This is a plate glass window in the cone of a cathode ray tube aligned to be parallel to the tube faceplate.

Synthetic-video Displays:- The use of a digital computer, as in an automatic detection and tracking processor, to extract target informat results in synthetic displays in which target information is presented with standard symbols and accompanying alphanumerics.
This is especially useful in air traffic control display in which such information as target identity and altitude is desired to be displayed.

# DUPLEXERS AND RECEIVER PROTECTORS

The duplexer is the device that allows a single antenna to serve both the transmitter and the receiver. On transmission it must protect the receiver from burnout or damage, and on receptio it must channel the echo signal to the receiver. Duplexer especially for high-power applications, sometimes employ a form of gas discharge device.

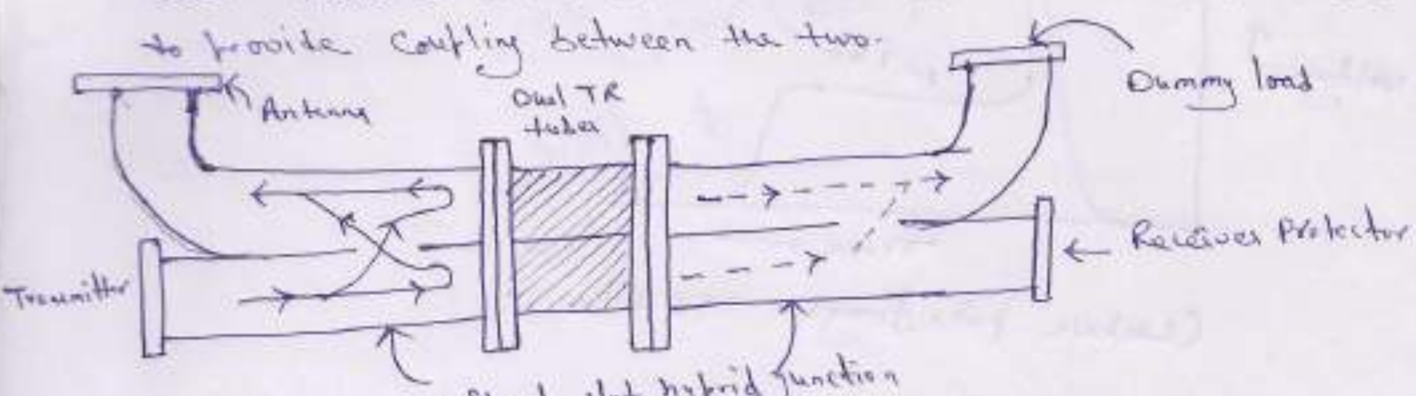
Branch type duplexer:— This the earliest duplexer configuration employed. It consist of a TR (transmit-receiver) switch and an ATR( anti transmit receiver) switch, both of which are gas discharged tub When the transmitter is turned on, the TR and ATR tubes ionize; that is they break down, or fire. The TR in the fired condition acts as a short circuit to prevent transmitter power from entering the receiver. Since the TR is located a quarter wavel from the main transmission line so that it does not impede the flow of transmitted power.



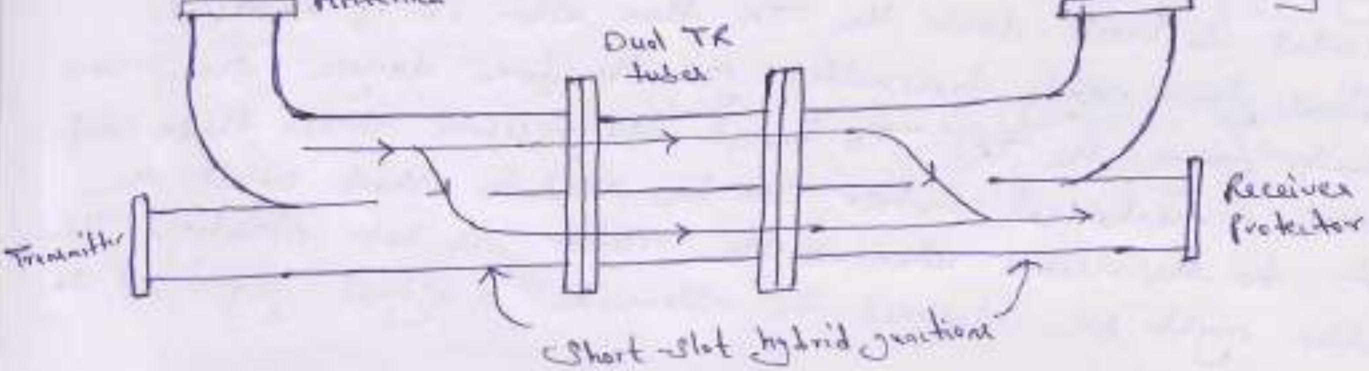
( Branch type duplexer )

The branch type duplexer is of limited bandwidth and power handling capability, and has generally been replaced by the balanced duplexer and other protecting devices. It is used in spite of these limitation, in some low cost radar.

**Balanced duplexer:—** It is based on short slot hybrid junction which consist of two sections of waveguide joined along on of their narrow walls with a slot cut in common narrow wall to provide coupling between the two.



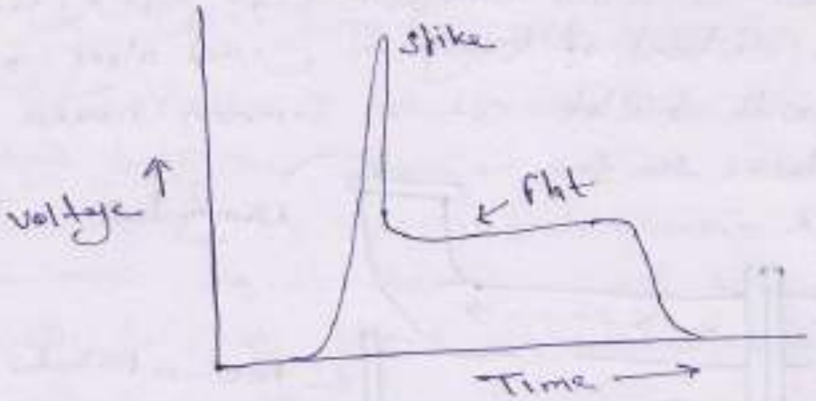
(Balanced duplexer using TR tubes Transmit condition)



(Receiver condition)

Both TR tubes break down and reflect the incident power out the antenna arm. The short-slot hybrid has the property that each time the energy passes through slot in them 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 is directed to the arm with the matched dummy load and not to the receiver.

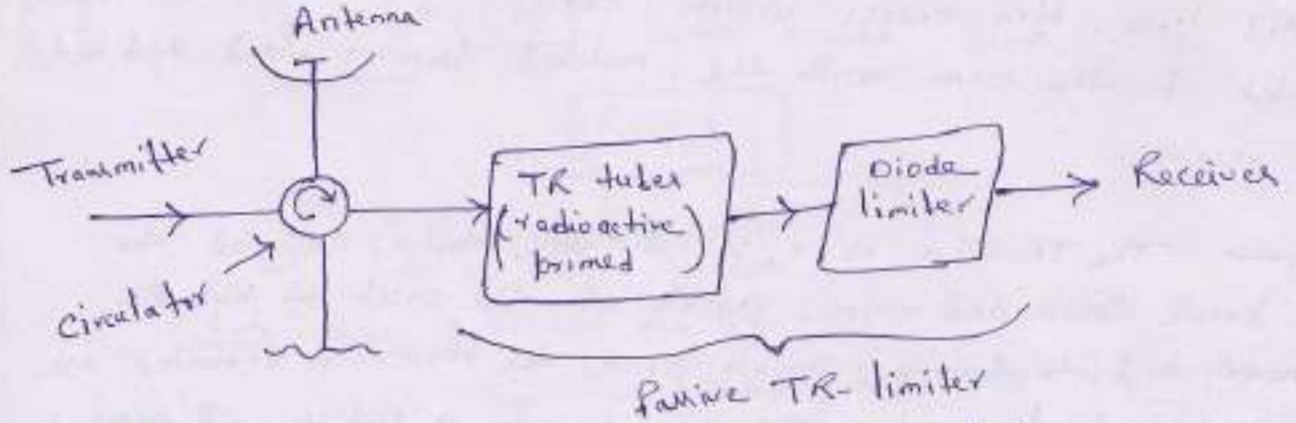
**TR tubes:—** The TR tubes is a gas-discharge device designed to break down and ionize quickly at the onset of high RF power, and to deionize quickly once the power is removed. one common construction of a TR consists of a section of waveguide contained one or more resonant filters and two glass to metal window to seal in the gas at the low pressure.



(Leakage pulse through a TR tubes)

**Receiver Protectors:→** Since the keep alive in the TR is not usually energized when the radar turned off, considerally more power is needed to break down the TR than when it is energized. Radiations from nearby transmitters may therefore damage the receiver without firing the TR. To protect the receiver under these conditions, a mechanical shutter can be used to short circuit the input to the receiver whenever the radar is not operating. The shutter might be designed to attenuate a signal by 25 to 50 dB.

**Circulator and receiver protector:→** The ferrite circulator is a three or four port device that can in principle, offer separation of the transmitter and receiver without the need for the convention duplexer configurations.



(Circulator and receiver protector)

# UNIT V

# RADAR RECEIVERS

**Matched-Filter Receiver:**

A network whose frequency-response function maximizes the output peak-signal-to-mean-noise (power) ratio is called a matched filter. This criterion, or its equivalent, is used for the design of almost all radar receivers.

The frequency-response function, denoted H(f), expresses the relative amplitude and phase of the output of a network with respect to the input when the input is a pure sinusoid. The magnitude $|H(f)|$ of the frequency-response function is the receiver amplitude passband characteristic.

If the bandwidth of the receiver passband is wide compared with that occupied by the signal energy, extraneous noise is introduced by the excess bandwidth which lowers the output signal-to-noise ratio. On the other hand, if the receiver bandwidth is narrower than the bandwidth occupied by the signal, the noise energy is reduced along with a considerable part of the signal energy.

The net result is again a lowered signal-to-noise ratio. Thus there is an optimum bandwidth at which the signal-to-noise ratio is a maximum. This is well known to the radar receiver designer.

The rule of thumb quoted in pulse radar practice is that the receiver bandwidth B should be approximately equal to the reciprocal of the pulse width $\tau$. This is a reasonable approximation for pulse radars with conventional superheterodyne receivers. It is not generally valid for other waveforms, however, and is mentioned to illustrate in a qualitative manner the effect of the receiver characteristic on signal-to-noise ratio.

The exact specification of the optimum receiver characteristic involves the frequency-response function and the shape of the received waveform.

The receiver frequency-response function, is assumed to apply from the antenna terminals to the output of the IF amplifier. (The second detector and video portion of the well designed radar superheterodyne receiver will have negligible effect on the output signal-to- noise ratio if the receiver is designed as a matched filter.) Narrow banding is most conveniently accomplished in the IF.

The bandwidths of the RF and mixer stages of the normal superheterodyne receiver are usually large compared with the IF bandwidth. Therefore the frequency-response function of the portion of the receiver included between the antenna terminals to the output of the IF amplifier is taken to be that of the IF amplifier alone. Thus we need only obtain the frequency-response function that maximizes the signal-to-noise ratio at the output of the IF. The IF amplifier may be considered as a filter with gain. The response of this filter as a function of frequency is the property of interest. For a received waveform s(t) with a given ratio of signal energy E to noise energy No (or noise power per hertz of bandwidth), North showed that the frequency-response function of the linear, time-invariant filter which maximizes the output peak-signal-to-mean-noise (power) ratio for a fixed input signal-tonoise (energy) ratio is

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

where $S(f) = \int_{-\infty}^{\infty} s(t) \exp(-j2\pi f t)\, dt$ = voltage spectrum (Fourier transform) of input signal

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

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

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

The noise that accompanies the signal is assumed to be stationary and to have a uniform spectrum (white noise). It need not be gaussian. The filter whose frequency-response function is given by Eq. above has been called the North filter, the conjugate filter, or more usually the matched filter. It has also been called the Fourier transform criterion. It should not be confused with the circuit-theory concept of impedance matching, which maximizes the power transfer rather than the signal-to-noise ratio.

The frequency-response function of the matched filter is the conjugate of the spectrum of the received waveform except for the phase shift exp (- $j2\Pi f_{t1}$). This phase shift varies uniformly with frequency. Its effect is to cause a constant time delay. A time delay is necessary in the specification of the filter for reasons of physical realizability since there can be no output from the filter until the signal is applied.

The frequency spectrum of the received signal may be written as an amplitude spectrum

|S(f) | ( and a phase spectrum exp [- jφs (f )]. The matched- filter frequency-response function may similarly be written in terms of its amplitude and phase spectra |H(f) | and exp [- jφm(f)]. Ignoring the constant Ga, Eq. above for the matched filter may then be written as

$$|H(f)| \exp[-j\phi_m(f)] = |S(f)| \exp\{j[\phi_s(f) - 2\pi f t_1]\}$$

or

$$|H(f)| = |S(f)|$$

and

$$\phi_m(f) = -\phi_s(f) + 2\pi f t_1$$

Thus the amplitude spectrum of the matched filter is the same as the amplitude spectrum of the signal, but the phase spectrum of the matched filter is the negative of the phase spectrum of the signal plus a phase shift proportional to frequency.

The matched filter may also be specified by its impulse response h(t), which is the inverse Fourier transform of the frequency-response function.

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

Physically, the impulse response is the output of the filter as a function of time when the input is an impulse (delta function).

Since $S^*(f) = S(-f)$, we have

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

A rather interesting result is that the impulse response of the matched filter is the image of the received waveform; that is, it is the same as the received signal run backward in time starting from the fixed time $t_1$. Figure 1 shows a received waveform s(t) and the impulse response h(t) of its matched filter. The impulse response of the filter, if it is to be realizable, is not defined for t< 0. (One cannot have any response before the impulse is applied.) Therefore we must always have $t < t_1$. This is equivalent to the condition placed on the transfer function H(f) that there be a phase shift exp $(-j2\Pi f_{t1})$. However, for the sake of convenience, the impulse response of the matched filter is sometimes written simply as s(- t).

## Derivation of the matched-filter characteristic:

The frequency-response function of the matched filter has been derived by a number of authors using either the calculus of variations or the Schwartz inequality. We shall derive the matched-filter frequency-response function using the Schwartz inequality.

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

We wish to show that the frequency-response function of the linear, time-invariant filter which maximizes the output peak-signal-to-mean-noise ratio is

When the input noise is stationary and white (uniform spectral density). The ratio
we wish to maximize is

R.

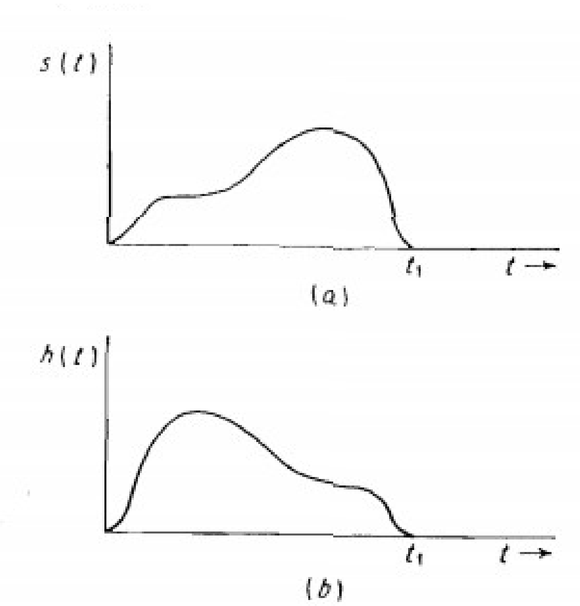


**Fig.1 (a**) Received waveform s(t); (b) impulse response h(t) of the matched filter.

Where |so(t) |max = maximum value of output signal voltage and N = mean noise power at receiver output. The ratio $R_f$ is not quite the same as the signal-to-noise ratio which has been considered in the radar equation. The output voltage of a filter with frequency-response function H(f) is

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

Where S(f) is the Fourier transform of the input (received) signal. The mean output noise power is

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

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

Where $N_0$ is the input noise power per unit bandwidth. The factor appears before the integral because the limits extend from - ∞ to +∞, whereas No is defined as the noise power per cycle of bandwidth over positive values only. Assuming that the maximum value of |so(t) |2 occurs at time t = t1, the ratio Rf becomes

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

$$\int P^*P \, dx \int Q^*Q \, dx \geq \left| \int P^*Q \, dx \right|^2$$

The equality sign applies when P = kQ, where k is a constant. Letting

$$P^* = S(f) \exp{(j2\pi f t_1)} \quad \text{and} \quad Q = H(f)$$

**and recalling that**

$$\int P^*P \, dx = \int |P|^2 \, dx$$

We get, on applying the Schwartz inequality to the numerator of Eq. earlier, we get

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

From Parseval's theorem,

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

Therefore we have

$$R_f \leq \frac{2E}{N_0}$$

The frequency-response function which maximizes the peak-signal-to-mean-noise ratio $R_f$ may be obtained by noting that the equality sign in Eq. applies when P = kQ, or

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

Where the constant k has been set equal to $1/G_a$.

**Relation between the matched filter characteristics and correlation function:**

The matched filter and the correlation function. The output of the matched filter is not a replica of the input signal. However, from the point of view of detecting signals in noise, preserving the shape of the signal is of no importance. If it is necessary to preserve the shape of the input pulse rather than maximize the output signal-to-noise ratio, some other criterion must be employed.

The output of the matched filter may be shown to be proportional to the input signal cross-correlated with a replica of the transmitted signal, except for the time delay $t_1$. The crosscorrelation function R(t) of two signals $y(\lambda)$ and $s(\lambda)$, each of finite duration, is defined as

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

The output $y_o(t)$ of a filter with impulse response h(t) when the input is $y_{in}(t) = s(t) + n(t)$ is

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

If the filter is a matched filter, then $h(\lambda) = s(t_1 - \lambda)$ and Eq. above becomes

$$y_0(t) = \int_{-\infty}^{\infty} y_{in}(\lambda) s(t_1 - t + \lambda)\, d\lambda = R(t - t_1)$$

Thus the matched filter forms the cross correlation between the received signal corrupted by noise and a replica of the transmitted signal. The replica of the transmitted signal is "built in" to the matched filter via the frequency-response function. If the input signal yin (t) were the same as the signal s(t) for which the matched filter was designed (that is, the noise is assumed negligible), the output would be the autocorrelation function. The autocorrelation function of a rectangular pulse of width $\tau$ is a triangle whose base is of width $2\tau$.

**Efficiency of non-matched filters:**

In practice the matched filter cannot always be obtained exactly. It is appropriate, therefore, to examine the efficiency of non matched filters compared with the ideal matched filter. The measure of efficiency is taken as the peak signal-to-noise ratio from the non matched filter divided by the peak signal-to-noise ratio ($2E/N_o$) from the matched filter. Figure. Plots the efficiency for a single-tuned (RLC) resonant filter and a rectangular-shaped filter of half-power bandwidth $B_\tau$ when the input is a rectangular pulse of width $\tau$. The maximum efficiency of the single-tuned filter occurs for $B_\tau \approx 0.4$. The corresponding loss in signal-to-noise ratio is 0.88 dB as compared with a matched filter.

Table lists the values of $B_\tau$ which maximize the signal-to-noise ratio (SNR) for various combinations of filters and pulse shapes. It can be seen that the loss in SNR incurred by use of these non-matched filters is small.

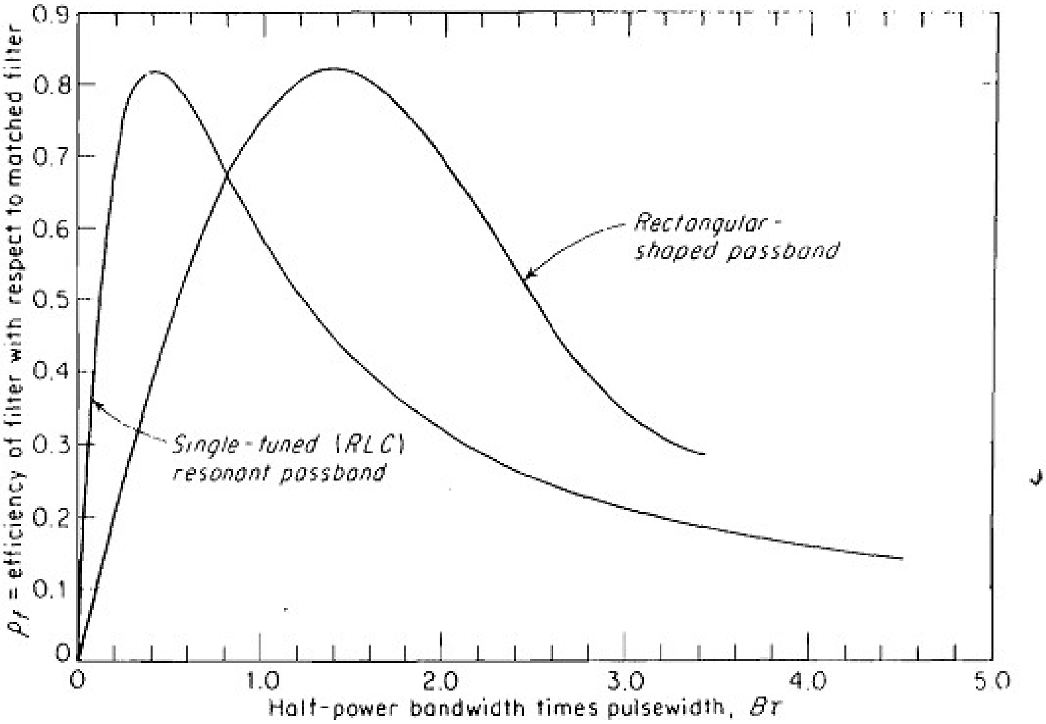
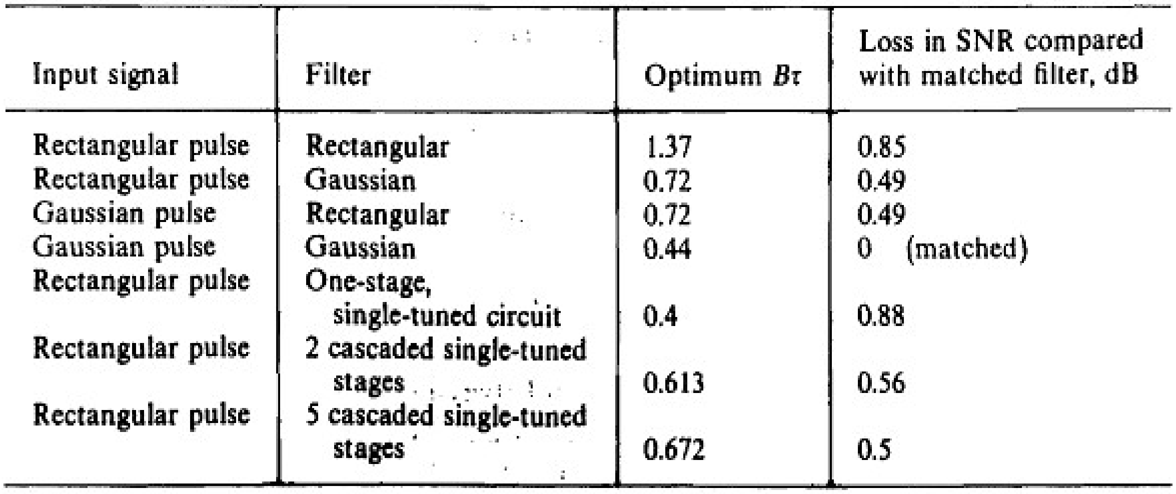


**Fig:** Efficiency, relative to a matched filter, of a single-tuned resonant filter and a rectangular shaped filter, when the input signal is a rectangular pulse of width $\tau$. B = filter bandwidth



| Input signal | Filter | Optimum $B\tau$ | Loss in SNR compared with matched filter, dB |
|---|---|---|---|
| Rectangular pulse | Rectangular | 1.37 | 0.85 |
| Rectangular pulse | Gaussian | 0.72 | 0.49 |
| Gaussian pulse | Rectangular | 0.72 | 0.49 |
| Gaussian pulse | Gaussian | 0.44 | 0 (matched) |
| Rectangular pulse | One-stage, single-tuned circuit | 0.4 | 0.88 |
| Rectangular pulse | 2 cascaded single-tuned stages | 0.613 | 0.56 |
| Rectangular pulse | 5 cascaded single-tuned stages | 0.672 | 0.5 |

**Table:** Efficiency of nonmatched filters compared with the matched filter

**Matched filter with nonwhite noise:**

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

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

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

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

This indicates that the NWN matched filter can be considered as the cascade of two filters. The first filter, with frequency-response function $1/N_i(f)$, acts to make the noise spectrum uniform, or white. It is sometimes called the whitening filter. The second is the matched filter when the input is white noise and a signal whose spectrum is $S(f)/N_i(f)$.

## Correlation Detection:

$$y_0(t) = \int_{-\infty}^{\infty} y_{1n}(\lambda) s(t_1 - t + \lambda)\, d\lambda = R(t - t_1)$$

Equation above describes the output of the matched filter as the cross correlation between the input signal and a delayed replica of the transmitted signal. This implies that the matched-filter receiver can be replaced by a cross-correlation receiver that performs the same mathematical operation as shown in Fig.5. The input signal y (t) is multiplied by a delayed replica of the transmitted signal s(t - Tr), and the product is passed through a low-pass filter to perform the integration. The cross-correlation receiver of Fig.5 tests for the presence of a target at only a single time delay Tr. Targets at other time delays, or ranges, might be found by varying $T_r$. However, this requires a longer search time. The search time can be reduced by adding parallel channels, each containing a delay line corresponding to a particular value of Tr, as well as a multiplier and low-pass filter. In some applications it may be possible to record the signal on some storage medium, and at a higher playback speed perform the search sequentially with different values of $T_r$. That is, the playback speed is increased in proportion to the number of time-delay intervals $T_r$ that are to be tested.

Since the cross-correlation receiver and the matched-filter receiver are equivalent mathematically, the choice as to which one to use in a particular radar application is determined by which is more practical to implement. The matched-filter receiver, or an approximation, has been generally preferred in the vast majority of applications.

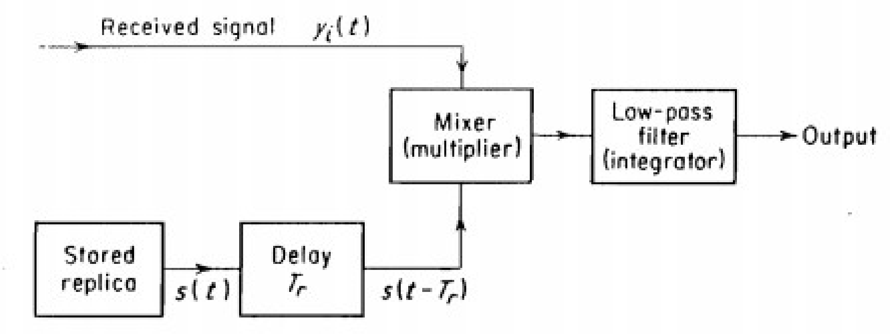


**Fig:** Block diagram of a cross-correlation receiver.