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Internal Assessment Test 2 – June 2021(Scheme & Solutions)

Sub:	RADAR ENGINEERING					Sub Code:	17EC833/ 15EC833	Branch:	ECE
Date:	20-06-2021 (Sunday)	Duration:	90 mins (1pm-2.30pm)	Max Marks:	50	Sem/Sec:	VIII - A,B,C,D		OBE

<u>Answer any **FIVE FULL** Questions.</u>

		MARKS	CO	RBT
1	a) Derive the modified RADAR equation in terms of signal-to-noise ratio (SNR).	[10 = 7+3]	CO4	L3

Solution :

- Even if we assume ideal conditions –

1) The Radar operates in a perfectly noise free environment (no external sources of noise present with Target Signal).
2) The Receiver of Radar is perfect (does not generate any excess noise).

- Noise will still be present which is generated by thermal agitation of the conduction electrons in the ohmic portion of the receiver input stages (**Thermal Noise** or **Johnson Noise**).

Generated at the input of a Radar Receiver.

If receiver has Bandwidth Bn(hertz) at Temperature T(degrees Kelvin), then,

$$\text{available thermal-noise power} = kTB_n \qquad [2.2]$$

where k = Boltzmann's constant = 1.38×10^{-23} J/deg. (The term *available* means that the device is operated with a matched input and a matched load.) The bandwidth of a superheterodyne receiver (and almost all radar receivers are of this type) is taken to be that of the IF amplifier (or matched filter).

In Eq. (2.2) the bandwidth B_n is called the *noise bandwidth*, defined as

$$B_n = \frac{\int_0^{\infty} |H(f)|^2 \, df}{} \qquad [2.3]$$

Where H(f)= frequency response function of the IF amplifier (filter), and fo = frequency of the maximum response (usually occurs at midband).

- Half power bandwidth B is usually used to approximate the noise bandwidth B_n.
- Noise power in practical receivers is greater than that from Thermal Noise alone.
- Noise Figure (F_n) –
  - ➢ The measure of the noise out of a real receiver (or network) to that from the ideal receiver with only Thermal Noise.

$$F_n = \frac{\text{noise out of practical receiver}}{\text{noise out of ideal receiver at std temp } T_0} = \frac{N_{out}}{kT_0 B G_a} \qquad [2.4]$$

Where

- ➢ Nout = noise out of the receiver,
- ➢ Ga = available gain,
- ➢ To= standard temperature, defined by IEEE as 290 K (62°F). (Close to Room Temperature) [kTo thus becomes $4*10^{-21}$ W/Hz].

If

- ➢ Sout= Signal Out , and Sin= Signal In, with both the output and input matched to deliver maximum output power, then,
- ➢ Ga=Sout/Sin, and Input Noise Nin , in an ideal receiver = kToBn.

Therefore, Eq.2.4 can be rewritten as,

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

This equation shows that the noise figure may be interpreted as a measure of the dation of the signal to noise ratio as the signal passes through the receiver.

Rearranging Eq. (2.5), the input signal is

$$S_{in} = \frac{kT_0 B F_n S_{out}}{N_{out}} \qquad [2.6]$$

If the minimum detectable signal S_{min} is that value of S_{in} which corresponds to the minimum detectable signal-to-noise ratio at the output of the IF, $(S_{out}/N_{out})_{min}$, then

$$S_{min} = kT_0 B F_n \left(\frac{S_{out}}{N_{out}}\right)_{min} \qquad [2.7]$$

Substituting the above into Eq. (2.1), and omitting the subscripts on S and N, results in the following form of the radar equation:

$$R^4_{max} = \frac{P_t G A_e \sigma}{(4\pi)^2 kT_0 B F_n (S/N)_{min}} \qquad [2.8]$$

For convenience, R_{max} on the left-hand side is usually written as the fourth power rather than take the fourth root of the right-hand side of the equation.

The minimum detectable signal is replaced in the radar equation by the *minimum detectable signal to noise ratio* $(S/N)_{min}$. The advantage is that $(S/N)_{min}$ is independent of the receiver bandwidth and noise figure, and, as we shall see in Sec. 2.5, it can be expressed in terms of the probability of detection and the probability of false alarm, two parameters that can be related to the radar user's needs.

b) Explain the need for replacing S_{min} with the term having SNR in the simple form of the radar equation.

Solution :

In practice, the simple form of Radar Equation fails to accurately predict range R due to the following 4 reasons :

- The statistical nature of S_{min} (usually determined by receiver noise).
- Fluctuations and Uncertainties in the Target's Radar Cross Section.
- The Losses experienced throughout a radar system.
- Propagation Effects caused by the Earth's Surface and Atmosphere.

- S_{min} (Minimum Detectable Signal) – The weakest signal that can just be detected by a receiver.
- Use of S_{min} not common in Radar.
- Use of S_{min} is not the preferred method.
- S_{min} not used for describing the ability of a Radar Receiver to detect echo signals from Targets.

- Threshold detection & SNR is popular term in Radar Studies compared to S_{min} in the detection of echo signals.
- Radar echo signal detection is based on establishing a Threshold at the output of the receiver.
- Target is said to be present if receiver output exceeds the Threshold.
- Noise is said to be present if receiver output does not cross the Threshold.

2	Make use of a portion of the radar receiver block diagram, and discuss with necessary equations, the probability of false alarm and probability of detection.	[10]	CO4	L3

Solution :

- The basic concepts for the detection of signals in noise are found in a classical review paper by S.O.Rice or a text book on detection theory.
- In order to find R_{max}, we need $(S/N)_{min}$ to be computed first.
- (S/N) computation further needs a knowledge of P_d and P_{fa}.
- Hence, we need to find P_d and P_{fa} to compute (S/N) using which we ca compute R_{max}.

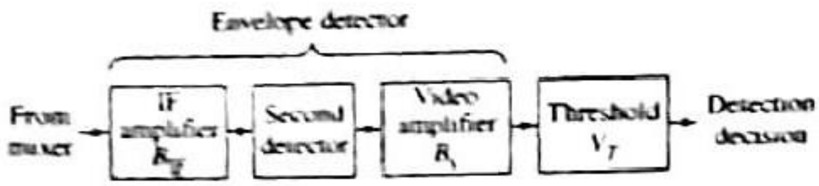


Figure 2.3 Portion of the radar receiver where the echo signal is detected and the detection decision is made.

- Figure 2.3 is a portion of a superheterodyne radar receiver.
- IF amplifier shown has a bandwidth B_{IF}.
- Second Detector is a diode stage (the name 'Second Detec distinguishes it from other detectors like 'Phase Detector' etc.).
- Video amplifier has a bandwidth B_v.
- Envelope detector is collective name for the 3 above stages as it gi output which is the envelope or modulation of the IF Signal (passes modulation and rejects the carrier).
- Requirement for Envelope Detector – 1) $B_v \geq B_{IF}/2$ and 2) $f_{IF} >>$ (both conditions are usually met in radar).
- If input to Threshold Detector crosses the threshold V_T , a signa declared to be present.

Receiver Noise at the input of IF Filter (terms 'filter' and 'amplifier' interchangeably used here) is described as follows:
It is considered to have a Gaussian Probability density function with mean value of zero, or ,

$$p(v) = \frac{1}{\sqrt{2\pi\Psi_0}} \exp\left(-\frac{v^2}{2\Psi_0}\right) \qquad [2.20]$$

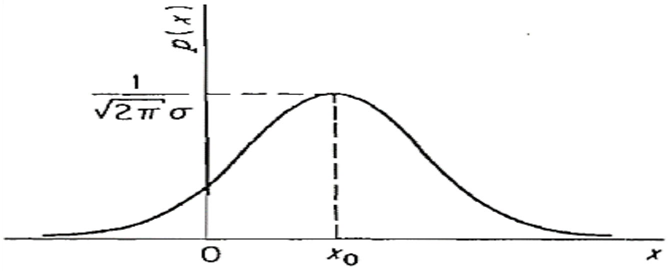
- p(v) dv = probability of finding noise voltage v between the values
- v and v+dv and = mean square value of noise voltage (mean noise power).

- The gaussian density function has a bell-shaped appearance anc defined by :

$$p(x) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp \frac{-(x - x_0)^2}{2\sigma^2}$$

- x_o is the mean

(=0 in Eqn.2.20)



(b)

- When above gaussian noise is passed through the IF Filter, the probabi density function (pdf) of the envelope R is a form of Rayleigh pdf (shc by S.O.Rice) as follows :

$$p(R) = \frac{R}{\Psi_0} \exp\left(-\frac{R^2}{2\Psi_0}\right) \quad [2.21]$$

- Integrating p(R) from V_T to infinity gives the probability that R will exceed V_T and we get :

$$\text{Probability } (V_T < R < \infty) = \int_{V_T}^{\infty} \frac{R}{\Psi_0} \exp\left(-\frac{R^2}{2\Psi_0}\right) dR = \exp\left(\frac{-V_T^2}{2\Psi_0}\right)$$

- The above integration of p(R) represents that when only noise is pres noise will cross the V_T value with the given probability and be fals called the Target.
- Thus, it is same as the **Probability of False Alarm** (P_{fa}) , and hence, have :

$$P_{fa} = \exp\left(-\frac{-V_T^2}{2\Psi_0}\right) \quad [2.23]$$

- But, to measure the effect of noise on radar performance (measurin excessive false alarms are troubling or not) , T_{fa} (false alarm time better than P_{fa}.

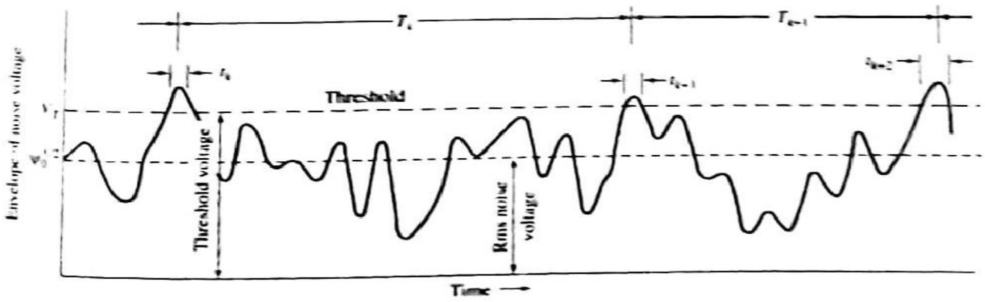


Figure 2.4 Envelope of the receiver output with noise alone, illustrating the duration of false alarms and the time between false alarms.

- Figure 2.4 illustrates the occurrence of false alarms.
- **False Alarm Time** (T_{fa}) is the average time between the crossings of decision Threshold when Noise alone is present, and is given by :

$$T_{fa} = \lim_{N \to \infty} \frac{1}{N} \sum_{k=1}^{N} T_k \qquad [2.24]$$

- T_k is the time between the crossings of the threshold V_T by the no envelope.
- T_{fa} is easier to relate to for a radar customer or operator than P_{fa}.

- Noting that P_{fa} is the ratio of the time the envelope(R) is actually ab the threshold (sum of all t_k from k=1 to N), to the total time it could h been above the threshold(T_{fa}) :

$$P_{fa} = \frac{\sum_{k=1}^{N} t_k}{\sum_{k=1}^{N} T_k} = \frac{\langle t_k \rangle_{av}}{\langle T_k \rangle_{av}} = \frac{1}{T_{fa}\, B}$$

- tk and Tk are shown in Fig.2.4
- B = bandwidth of the IF amplifier and approx. equals reciprocal of "s of all t_k from k=1 to N".

Equating Eqs.2.23 and 2.25, we get :

$$T_{fa} = \frac{1}{B} \exp\left(\frac{V_T^2}{2\psi_0} \right) \qquad [2.26]$$

- Let the echo signal be a sine wave of amplitude A, appearing along v gaussian noise at envelope detector input.
- The pdf of envelope R at video output is then:

$$p_s(R) = \frac{R}{\Psi_0} \exp\left(-\frac{R^2 + A^2}{2\Psi_0}\right) I_0\left(\frac{RA}{\Psi_0}\right) \quad [2.27]$$

where $I_0(Z)$ is the modified Bessel function of zero order and argument Z. For large Z, an asymptotic expansion for $I_0(Z)$ is

$$I_0(Z) = \frac{e^Z}{\sqrt{2\pi Z}}\left(1 + \frac{1}{8Z} + \cdots\right) \quad [2.28]$$

When the signal is absent, $A = 0$ and Eq. (2.27) reduces to Eq. (2.21), the pdf for noise alone. Equation (2.27) is called the *Rice probability density function.*

The probability of detecting the signal is the probability that the envelope R will exceed the threshold V_T (set by the need to achieve some specified false-alarm time). Thus the probability of detection is

$$P_d = \int_{V_T}^{\infty} p_s(R)\, dR \quad [2.29]$$

When the probability density function $p_s(R)$ of Eq. (2.27) is substituted in the above, the probability of detection P_d cannot be evaluated by simple means. Rice[9] used a series approximation to solve for P_d. Numerical and empirical methods have also been used.

The expression for P_d, Eq. (2.29), along with Eq. (2.27), is a function of the signal amplitude A, threshold V_T, and mean noise power Ψ_0. In radar systems analysis it is more convenient to use signal-to-noise power ratio S/N than $A^2/2\Psi_0$. These are related by

$$\frac{A}{\Psi_0^{1/2}} = \frac{\text{signal amplitude}}{\text{rms noise voltage}} = \frac{\sqrt{2}\,(\text{rms signal voltage})}{}$$

$$= \left(2\frac{\text{signal power}}{\text{noise power}}\right)^{1/2} = \left(\frac{2S}{N}\right)^{1/2}$$

The probability of detection P_d can then be expressed in terms of S/N and the ratio of the threshold-to-noise ratio $V_T^2/2\Psi_0$. The probability of false alarm, Eq. (2.23) is also a function of $V_T^2/2\Psi_0$. The two expressions for P_d and P_{fa} can be combined, by eliminating the threshold-to-noise ratio that is common to each, so as to provide a single expression relating the probability of detection P_d, probability of false alarm P_{fa}, and the signal-to-noise ratio S/N. The result is plotted in Fig. 2.6.

Albersheim[11,12] developed a simple empirical formula for the relationship between S/N, P_d, and P_{fa}, which is

$$S/N = A + 0.12AB + 1.7B \qquad [2.30]$$

where

$$A = \ln\,[0.62/P_{fa}] \quad \text{and} \quad B = \ln\,[P_d/(1 - P_d)]$$

3 What do you understand by "Radar Cross Section of Targets" ? Give the relevant equation and definition. Give an account of the radar cross section of simple targets like Sphere and Cone-Sphere. [10] CO1 L1

Solution:

- The radar cross section of a target is the property of a scattering object, or target, that is included in the radar equation to represent the magnitude of the echo signal returned to the radar by the target.
- The radar cross section was defined as:

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

- Another definition:

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

where

- R = distance between radar and target
- Er = reflected field strength at radar
- Ei= strength of incident field at target
- Larger the target size, the larger the cross section is likely to be.

Real targets do not scatter the incident energy uniformly in all directions.

- In theory, the scattered field, and hence the radar cross section, can be determined by solving **Maxwell's equations** with the proper boundary conditions applied or by computer modeling.
- Unfortunately, the determination of the radar cross section with Maxwell's equations can be accomplished only for the most **simple of shapes**, and solutions valid over a large range of frequencies are not easy to obtain.

<u>Sphere:</u>

- Since the sphere is a sphere no matter from what aspect it is viewed, its cross section will not be aspect-sensitive.
- The cross section of other objects, however, will depend upon the direction as viewed by the radar
- The radar cross section of a simple sphere is shown in Fig. 2.9 as a function of its circumference measured in wavelengths ($2\pi a/\lambda$, where a is the radius of the sphere and λ I is wavelength).

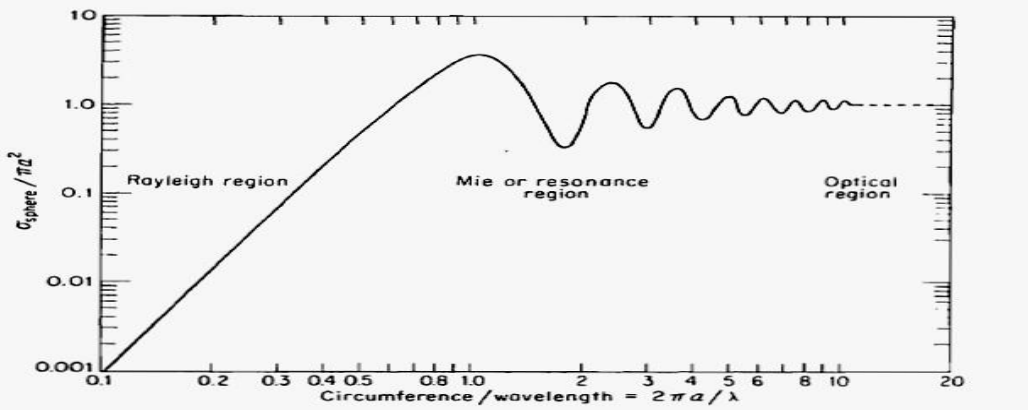


Figure 2.9 Radar cross section of the sphere. a = radius; λ = wavelength.

- **<u>Rayleigh scattering</u>** is the predominantly elastic scattering of light or other electromagnetic radiation by <u>particles much smaller than the wavelength</u> of the radiation. It does not change the state of material. ($2\pi a/\lambda << 1$)
- Rayleigh scattering is strongly dependent upon the <u>size of the particle and the wavelengths</u>.

❖ The intensity of the Rayleigh scattered radiation increases rapidly as the ratio of particle size to wavelength increases.

- For wave frequencies well below the resonance frequency of the scattering particle, the amount of scattering is inversely proportional to the fourth power of the wavelength.

- At radar frequencies, echo from rain is usually described by

Rayleigh scattering.

- The Mie solution to Maxwell's equations (also known as **Mie scattering**) describes the scattering of an electromagnetic plane wave by a homogeneous sphere.
- More broadly, "Mie scattering" suggests situations where the size of the scattering particles is comparable to the wavelength of the light, rather than much smaller or much larger.
- Dust, pollen,smoke and microscopic water droplets that

form clouds are common causes of Mie scattering.

- Mie scattering occurs mostly in the lower portions of the atmosphere, where larger particles are more abundant, and dominates in cloudy conditions.
- The cross section is **oscillatory** with frequency within this region.
- At the other extreme is the **optical region**, where the dimensions of the sphere are large compared with the wavelength ($2\pi a/\lambda >> 1$).
- For large $2\pi a/\lambda$, the radar cross section approaches the

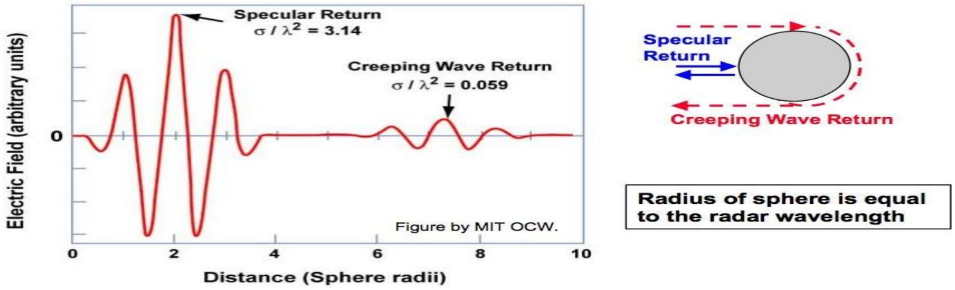
optical cross section πa^2 .

- This unique circumstance can mislead one into thinking that the geometrical area of a target is a measure of its radar cross section --------> It applies to only sphere.
- In the optical region scattering does not take place over the entire hemisphere that faces the radar, but only from a small *bright spot at the tip* of the smooth sphere (ex. Polished metallic sphere).

- The RCS of sphere in the resonance (Mie) region oscillates as a

function of frequency or $2\pi a/\lambda$.

- Its maximum occurs at $2\pi a/\lambda = 1$ and is 5.6 dB greater than its value in the optical region.
- Changes in cross section occur with changing frequency because there are two waves that interfere *constructively and destructively*.
- One is the direct reflection from the front face of the sphere.
- Other is the creeping wave that travels around the back of the sphere and returns to the radar where it interferes with the reflection from the front of the sphere.

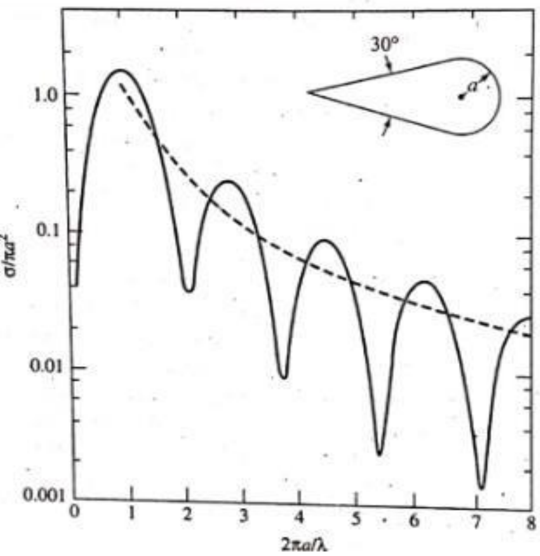


- Longer the electrical path around the sphere, greater the loss, so smaller will be the magnitude of the fluctuation with increasing frequency.
- Figure: illustrates the backscatter that would be produced by a very short pulse radar that can resolve the specular echo from the creeping wave.
- The creeping wave lags the specular return by the time required to travel one half the circumference of the sphere plus the diameter.

Cone Sphere:

- This is a cone whose base is capped with a sphere.
- The first derivatives of the cone and sphere contours are equal at the join between the two.

Figure 2.12 Theoretical normalized nose-on radar cross section of a cone-sphere based on an approximate impulse analysis; 15° half cone-angle (30° included cone-angle), a = radius of the sphere, and λ = wavelength. The dashed curve represents the approximation



4 a) Illustrate the concepts of pulse-repetition frequency and range ambiguities in the case of RADAR with suitable diagrams and equations. [10 = 6+4] CO1 L2

Solution:

- The pulse repetition frequency (prf) is determined primarily by the maximum range at which targets are expected.
- If the prf is made too high, the likelihood of obtaining target echoes from the wrong pulse transmission is increased.
- Echo signals received after an interval exceeding the pulse- repitition period are called *multiple-time-around* echoes.

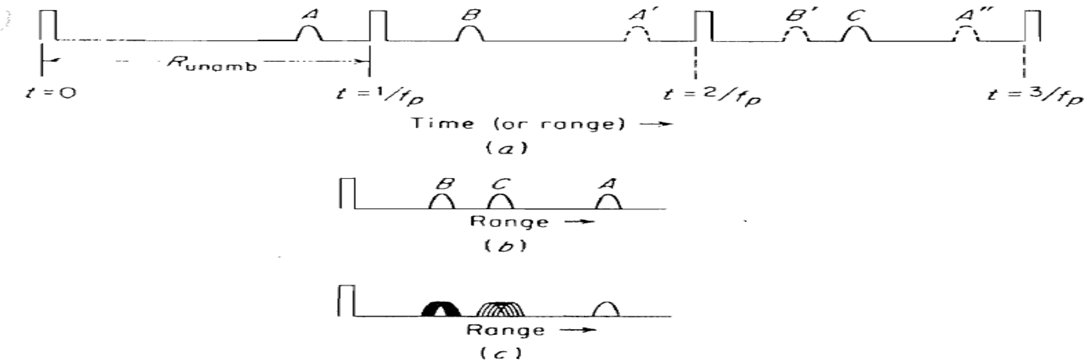


Figure 2.24 Multiple-time-around radar echoes that give rise to ambiguities in range. (a) Three targets A, B, and C, where A is within the unambiguous range R_{un}. B is a second-time-around ech and C is a multiple-time-around echo; (b) the appearance of the three echoes on an A-scope; (c) appearance of the three echoes on the A-scope with a changing prf

- Ambiguous range echoes can be recognized by *changing the prf* of the radar.
- In that case, the unambiguous echo remains at its true range.
- Echoes from multiple-time-around targets will be spread

over a finite range.

- The prf may be changed continuously within prescribed limits, or it may be changed discretely among several predetermined values.

If the first pulse repetition frequency f_1 has an unambiguous range R_{un1}, and if the apparent range measured with prf f_1 is denoted R_1, then the true range is one of the following

$$R_{true} = R_1, \text{ or } (R_1 + R_{un1}), \text{ or } (R_1 + 2R_{un1}), \text{ or} \ldots$$

Anyone of these might be the true range. To find which is correct, the prf is changed to f_2 with an unambiguous range R_{un2}, and if the apparent measured range is R_2, the true range is one of the following

$$R_{true} = R_2, \text{ or } (R_2 + R_{un2}), \text{ or } (R_2 + 2R_{un2}), \text{ or} \ldots$$

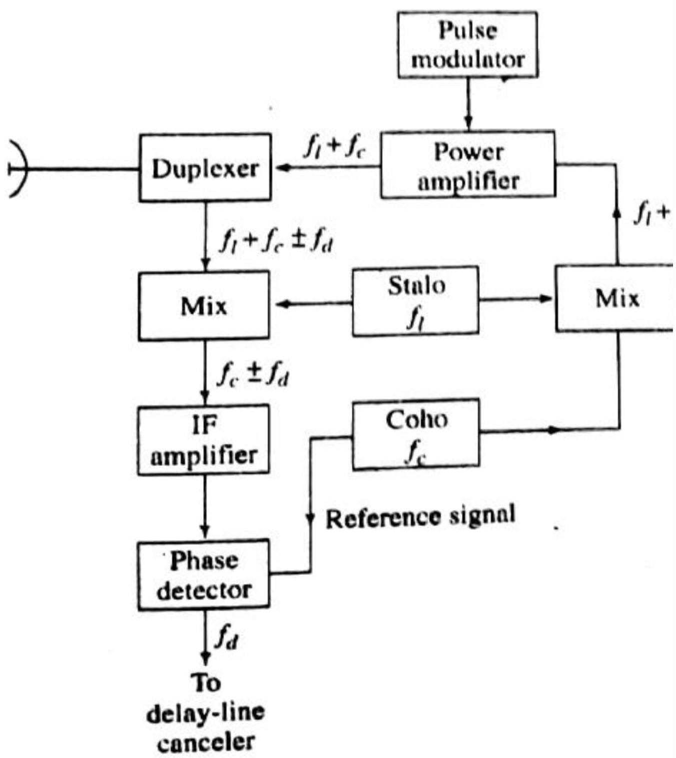
The correct range is that value which is the same with the two prfs.

	b) Discuss the various types of signal losses in Radar. Solution : ➢ System loss, Ls is a number greater than 1. ➢ Ls is inserted in the denominator of the radar equation. ➢ It is the reciprocal of efficiency (a number less than 1). ➢ Sometimes, Loss and efficiency are used interchangeably. 1) Microwave Plumbing Losses (Transmission Line Loss, Duplexer Loss). 2) Antenna Losses (Beam Shape Loss, Scanning Loss, Radome Loss, Phased Array Losses). 3) Signal Processing Losses (Nonmatched Filter, Constant False Alarm Rate(CFAR) Receiver, Automatic Integrators, Threshold Level, Limiting Loss, Straddling Loss, Sampling Loss). 4) Losses in Doppler-Processing Radar. 5) Collapsing Loss. 6) Operator Loss. 7) Equipment Degradation. 8) Propagation Effects. 9) Radar System Losses – the Seller and Buyer.			
5	a) Explain the principle and working of MTI Radar with neat block diagram. Solution : • Elaborate block diagram of an MTI radar employing a power amplifie transmitter is shown in Fig.3.7. • The local oscillator (LO) of an MTI radar's superheterodyne receiver r more stable than the LO for a radar that does not employ doppler. • If the phase of the local oscillator were to change significantly b pulses, an uncancelled clutter residue can result. • This will appear at the output of the delay-line canceler which mi mistaken for a moving target even though only clutter were present. • To recognize the need for high stability, the LO of an MTI receiver is the *stalo*, which stands for *stable local oscillator*. • The IF stage is designed as a matched filter, as is usually the case in rac • Instead of an amplitude detector, there is a *phase detector* following stage. • This is a mixer-like device that combines the received signal (at IF) reference signal from the *coho.* • This is done such that the difference between frequencies of received & the reference signal is produced.	[10 = 8+2]	CO2	L2

- This difference is the doppler frequency.
- The name *coho* stands for *coherent oscillator* to signify that it is the re signal that has the phase of the transmitter signal.
- Coherency with the transmitted signal is obtained by using the sum coho & the stalo signals as the input signal to the power amplifier.
- Thus the transmitter frequency is the sum of the stalo frequency f_l & th frequency f_c.
- This is accomplished in the mixer shown on the upper right side of Fig.
- The combination of the stalo & coho is sometimes called the *receiver-* portion of the MTI radar.

- Using the receiver stalo & coho to also generate the transmitter signal better stability than if the functions were performed with 2 different oscillators.
- The output of the phase detector is the input to the delay-line cancele Fig.3.6.
- The delay-line canceler acts as a high-pass filter.
- This is because it is used to separate the doppler-shifted echo sig moving targets from the unwanted echoes of stationary clutter.

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



b) A Doppler radar set operates at 10GHz and is used for traffic speed measurements. What is the Doppler frequency for the speed of 40 kmph?
Solution :

Fd= (2*Vr*Fo)/c = 740.740 Hz.

6 a) Explain single Delay line canceller with neat block diagram. Derive an expression for frequency response of single DLC. CO2 L2

Solution : [10 =6+4]

- The simple MTI delay-line canceler (DLC) of Fig.3.6 is an eg. of a time-domain filter that rejects stationary clutter at zero frequency.
- It has a frequency response function H(f) that can be derived from the time-domain representation of the signals.
The signal from a target at range Ro at the output of the phase detector can be written

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

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

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

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

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

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

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

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

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

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

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

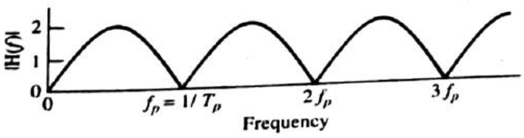


Figure 3.8 Magnitude of the frequency response $|H(f)|$ of a single delay-line canceler as given by Eq. (3.10), where f_p = pulse repetition frequency and $T_p = 1/f_p$.

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

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

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

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

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

b) What is blind speed? How can we eradicate it? Also obtain the expression for blind speeds.
Solution:

- Blind speeds can be a serious limitation in MTI radar.
- This is because they cause some desired moving targets to be canceled along with the undesired clutter at zero frequency.
- Based on Eq.(3.13), there are 4 methods for reducing the detrimental effects of blind speeds :

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

- Combinations of 2 or more of the above are also possible to further alleviate the effect of blind speeds.

- Each of these 4 methods has particular advantages as well as limitations.
- So, there is not always a clear choice as to which to use in any particular application.

- In some circumstances, it might be desirable to tolerate the blind speeds rather than accept the limitations of the above methods.
- As in many aspects of engineering, there is no one single solution best for all cases.
- The engineer has to decide which of the above limitations can be accepted in any particular application.
- Blind speeds occur because of the sampled nature of the pulse radar waveform.
- Thus, it is sampling that is the cause of ambiguities, or aliasing, in the measurement of the doppler frequency.
- This is similar to sampling in a pulse radar (at the prf) can give rise to ambiguities in the range measurement.

7 With neat block diagram, explain original Moving Target Detector (MTD) processor. [10] CO2 L2

Solution:

- The Moving Target Detector (MTD) is an example of an MTI processing system.
- It takes advantage of the various capabilities offered by digital techniques.
- These techniques are used to produce improved detection of moving targets in clutter.
- It was originally developed by the MIT Lincoln Laboratory for the airport-surveillance radar (ASR).
- ASR is a 60-nmi radar found at major airports for control of local air traffic.
- The introduction of the MTD represented an innovative & significant advance in radar detection of aircraft in the presence of clutter.

The original MTD concept was designed for a radar similar to the FAA's ASR-8.

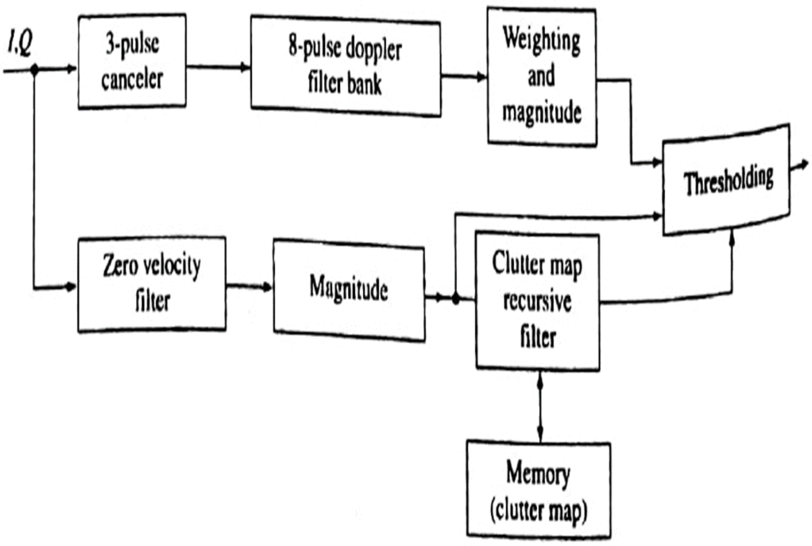
The ASR-8 employed 4 staggered prfs; but staggering of the prfs was not used with the MTD.

The original MTD included the following :

- An *eight-pulse FFT digital filter bank* with eight filters, preceded by a three-pulse delay-line canceler. The three-pulse canceler reduced the dynamic range of the signals which the doppler filter bank had to handle, and it compensated for the lack of adequate cancellation of stationary clutter in the doppler filters. The doppler filter bank separated moving targets from moving weather clutter if they appeared in different doppler filters.
- *Frequency-domain weighting* to reduce the doppler-filter sidelobes for better clutter attenuation.
- *Alternate prfs* to eliminate blind speeds and to unmask aircraft echoes from weather clutter.
- *Adaptive thresholds* to take advantage of the nonuniform nature of clutter.
- *Clutter map* to detect crossing targets with zero radial velocity that would otherwise be canceled by an ordinary MTI.
- *Centroiding* of multiple reports from the same target for more accurate location measurements.

- The range coverage of this processor totaled 47.5 nmi.
- The MTI processor was preceded by a large dynamic range receiver to avoid the reduction in improvement factor caused by limiting.
- The output of the receiver IF amplifier was fed to I and Q phase detectors.
- From there, the A/D converters changed the analog signals to 10-bit digital words.
- Figure 3.30 is a block diagram of the MTD.

Figure 3.30 Block diagram of the original Moving Target Detector (MTD) signal processor.



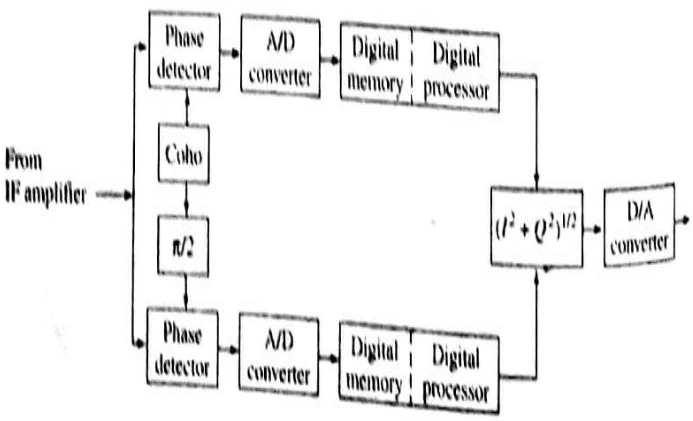
8	a) Explain the operation of Digital MTI Doppler Signal Processor with a neat block diagram.	[10 = 6+4]	CO2	L2

Solution :

- The block diagram of a digital MTI signal processor with I and Q channels is shown in Fig.3.29.
- The signal from the IF amplifier is split into 2 channels.
- The phase detectors in each channel extract the doppler-shifted signal.
- In the I channel, the doppler signal is represented as $A_d \cos (2\pi f_d t + \varphi_0)$ & the Q channel it is the same except that the sine replaces the cosine.

The signals are then digitized by the analog-to-digital (A/D) converter.
A sample & hold circuit usually is needed ahead of the A/D converter for more effective digitizing.
Sample & hold is often on the same chip as the A/D converter.
Some A/D converters, such as the flash type, do not require a sample & hold.
The digital words are stored in a digital memory for the required delay time(s).
They are then processed with suitable algorithm to provide the desired doppler filtering.
The magnitude of the doppler signal is obtained by taking the square root of I2 + Q2 .
If required, the combined unipolar output can be converted to an analog signal by a digital-to-analog (D/A) converter for display.
Otherwise, the digital output might be subject to further processing.

Figure 3.29 Block diagram of a digital MTI doppler signal processor.



b) Explain (i) Doppler shift (ii) Improvement factor.

Solution :

i)

- The doppler effect used in radar is the same phenomenon used to describe the changing pitch of an audible siren from an emergency vehicle as it travels toward or away from the listener.
- We are interested in the doppler effect that changes the frequency of the electromagnetic signal propagating from the radar to a moving target & back to the radar.
- If the range to the target is R, then the total number of wavelengths λ in the 2 way path from radar to target & return is 2R/λ.
- Each wavelength corresponds to a phase change of 2π radians.
- The total phase change in the 2-way propagation path is then

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

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

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

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

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

ii)

- The clutter attenuation is a useful measure of the performance of an MTI radar in canceling clutter.
- But, it has an inherent weakness if one is not careful.
- The clutter attenuation can be made infinite by turning off the radar receiver!
- This would not be done knowingly, since it also eliminates the desired moving-target echo signals.

- To avoid the problem of increasing clutter attenuation at the expense of desired signals, the IEEE defined a measure of performance.
- This is known as the *MTI Improvement Factor* which includes the signal gain as well as the clutter attenuation.
- It is defined as "The signal-to-clutter ratio at the output of the clutter filter divided by the signal-to-clutter ratio at the input of the clutter filter, averaged uniformly over all target radial velocities of interest."
- It is expressed as -

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

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

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

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

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

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

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

***NOTE* : THE QUESTIONS SHOULD BE NEATLY WRITTEN & ANSWERED IN STUDENT'S OWN HANDWRITING. ON TOP OF EACH PAGE, WRITE YOUR NAME, USN & PAGE NO. BEFORE MAKING A PDF AND UPLOADING THE PDF IN GOOGLE CLASSROOM. TOTAL TIME TAKEN SHOULD NOT EXCEED 2 HOURS FOR BOTH ANSWERING & UPLOADING THE PDF (1.5 HOUR FOR ANSWERING + 0.5 HOUR FOR UPLOADING PDF). PDF SUBMITTED AFTER 2 HOURS OR NOT AS PER THE ABOVE INSTRUCTIONS WILL NOT BE VALUATED AND MARKS ALLOTED WILL BE ZERO FOR THE TEST.**

ALL THE BEST