USN

Internal Assessment Test 2 – April 2023 (QP-1)

Sub:	**RADAR ENGINEERING**				Sub Code:	**18EC823**	Branch:	ECE
Date:	15-04-2023	Duration:	90 minutes	Max Marks: 50	Sem/Sec:	8^{th} (A,B,C,D)		OBE

	ANSWER ANY 5 FULL QUESTIONS	MARKS	CO	RBT
1	Derive the modified RADAR equation in terms of signal-to-noise ratio. Discuss the various types of signal losses in Radar.	5 + 5	CO4	L3
2	Illustrate the concepts of pulse-repetition frequency and range ambiguities in the case of RADAR with suitable diagrams and equations. Discuss the relation between False Alarm Time and Probability of False Alarm.	7 + 3	CO1	L2
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
4	Explain the principle and working of MTI Radar with neat block diagram.	10	CO2	L1
5	Explain the operation of Digital MTI Doppler Signal Processor with neat diagram.	10	CO2	L2
6	With neat block diagram, explain original Moving Target Detector(MTD) processor.	10	CO2	L2
7	Explain(i) clutter attenuation,(ii) improvement factor,(iii)Doppler shift. What is blind speed? How can we eradicate it?	6 + 4	CO2	L2
8	Explain single Delay line canceller with neat block diagram. Derivean expression for frequency response of single DLC. Also obtain the expression for blind speeds.	10	CO2	L2

Course Instructor **Chief Course Instructor** **HOD**

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Q1a) Solution.

Even if we assume ideal conditions –
The Radar operates in a perfectly noise free environment (no external sources of noise present with Target Signal).
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

$$B_n = \frac{\int_0^\infty |H(f)|^2\, df}{|H(f_0)|^2} \qquad [2.3]$$

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 Bn.
Noise power in practical receivers is greater than that from Thermal Noise alone.
Noise Figure (Fn) –
The measure of the noise out of a real receiver (or network) to that from the ideal receiver with only Thermal Noise.

$$F_n = \frac{\text{noise out of practical receiver}}{\text{noise out of ideal receiver at std temp } T_0} = \frac{N_{out}}{kT_0BG_a} \qquad [2.4]$$

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,

$$F_n = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} \qquad [2.5]$$

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

Rearranging Eq. (2.5), the input signal is

$$S_{in} = \frac{kT_0BF_nS_{out}}{N_{out}} \tag{2.6}$$

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

$$S_{min} = kT_0BF_n\left(\frac{S_{out}}{N_{out}}\right)_{min} \tag{2.7}$$

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

$$R^4_{max} = \frac{P_tGA_e\sigma}{(4\pi)^2kT_0BF_n(S/N)_{min}} \tag{2.8}$$

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.

Q1b) Solution :

System loss, Ls is a number greater than 1.

Ls is inserted in the denominator of the radar equation.

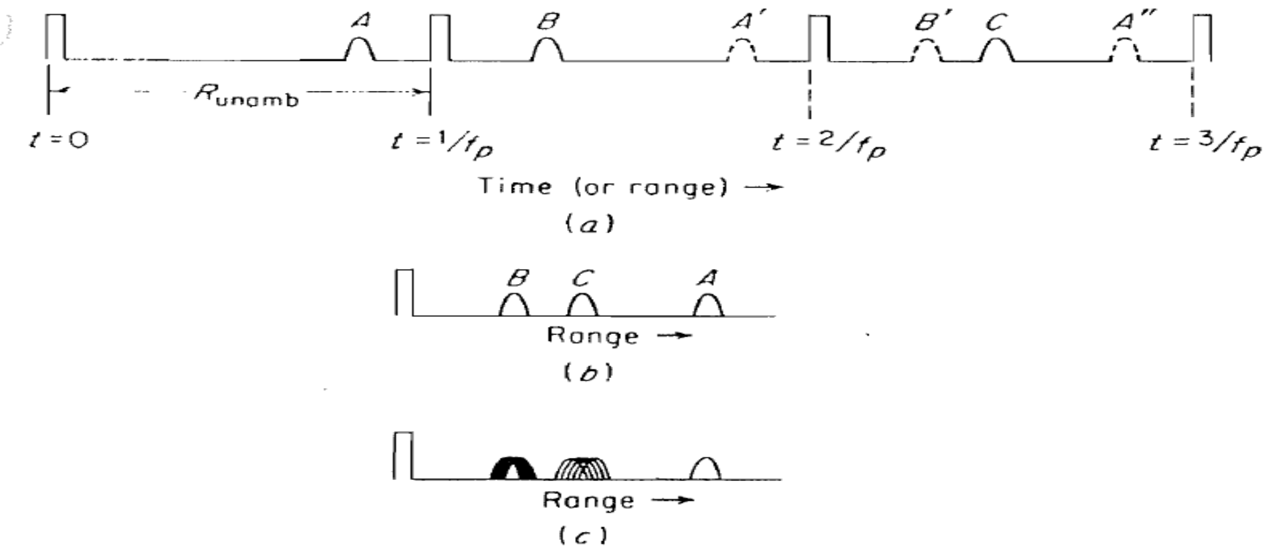
It is the reciprocal of efficiency (a number less than 1).

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.

Q2a) 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.



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.

Instead of modulating the prf, other schemes that might be employed to "mark" successive pulse so as to identify multiple-time-around echoes include changing:
the pulse amplitude
pulse width
frequency
Phase
polarization of transmission from pulse to pulse.

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.

Q2b) Solution :

Noting that Pfa is the ratio of the time the envelope(R) is actually above the threshold (sum of all tk from k=1 to N), to the total time it could have been above the threshold(Tfa) :

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

tk and Tk are shown in Fig.2.4
B = bandwidth of the IF amplifier and approx. equals reciprocal of “sum of all tk from k=1 to N”.

Q3) 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.

Sphere:
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 (2πa/λ, where a is the radius of the sphere and λ I is wavelength).

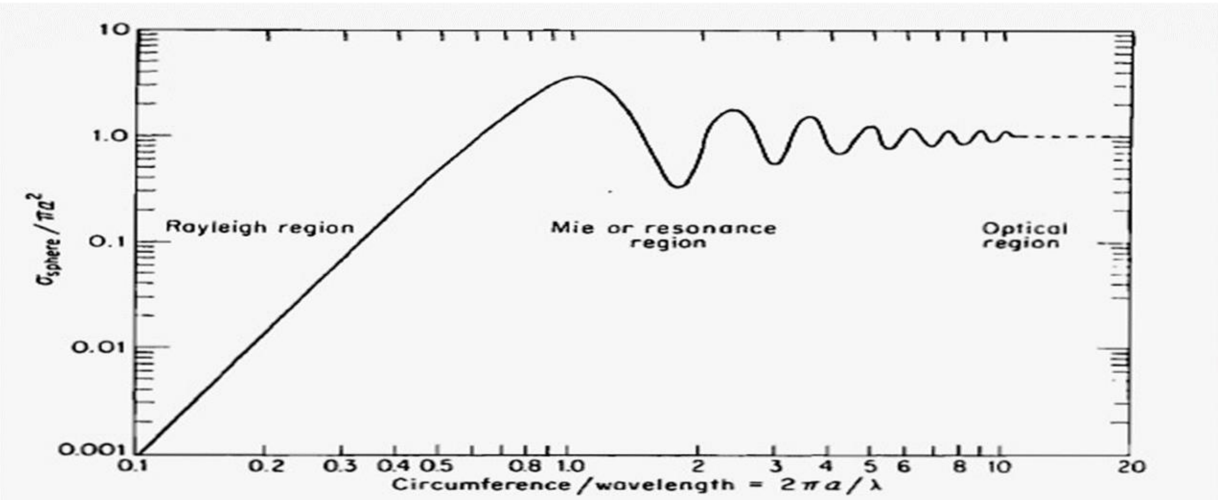


Figure 2.9 Radar cross section of the sphere. a = radius; λ = wavelength.

Rayleigh scattering is the predominantly elastic scattering of light or other electromagnetic radiation by particles much smaller than the wavelength of the radiation. It does not change the state of material. ($2\pi a/\lambda << 1$)
Rayleigh scattering is strongly dependent upon the size of the particle and the wavelengths.
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 $\pi a2$.
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).

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.

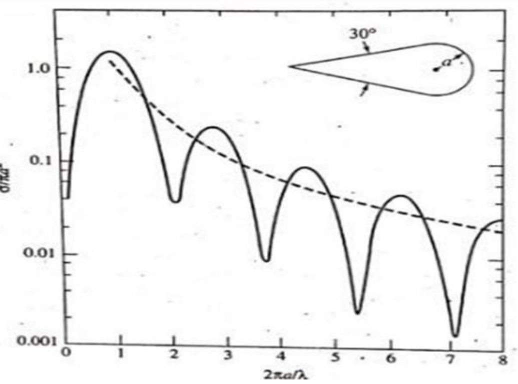
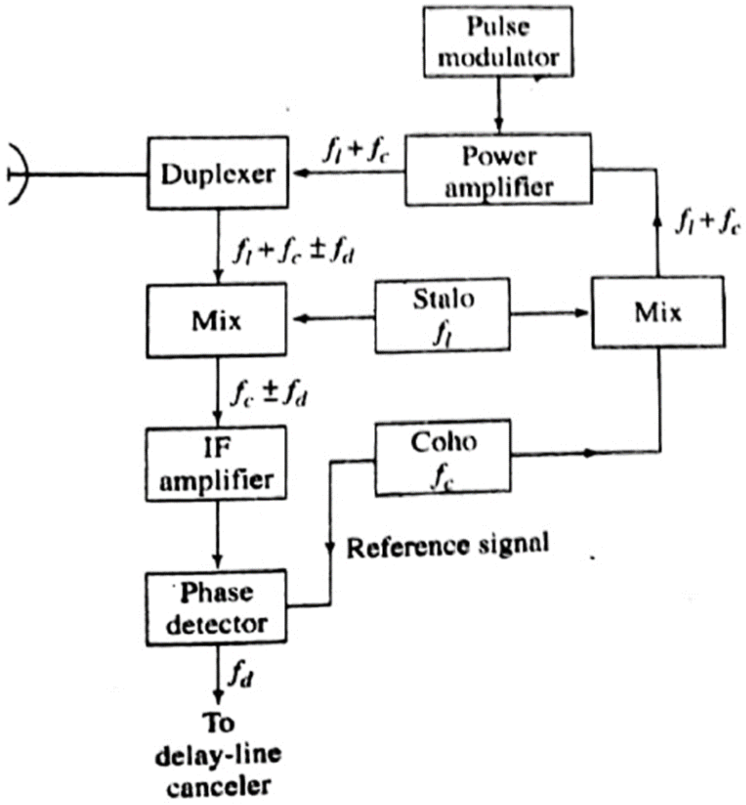


Figure 2.12 Theoretical normalized nose-on radar cross section of a cone-sphere based on an approximate impulse analysis; 15° half cone-angle (30° included cone-angle), a = radius of the sphere, and λ = wavelength. The dashed curve represents the approximation

Q4) Solution :

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



A slightly more elaborate block diagram of an MTI radar employing a power amplifier as the transmitter is shown in Fig.3.7.
The local oscillator (LO) of an MTI radar's superheterodyne receiver must be more stable than the LO for a radar that does not employ doppler.
If the phase of the local oscillator were to change significantly between pulses, an uncancelled clutter residue can result.
This will appear at the output of the delay-line canceler which might be 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 called the stalo, which stands for stable local oscillator.
The IF stage is designed as a matched filter, as is usually the case in radar.
Instead of an amplitude detector, there is a phase detector following the IF stage.
This is a mixer-like device that combines the received signal (at IF) & the reference signal from the coho.
This is done such that the difference between frequencies of received signal & the reference signal is produced.

This difference is the doppler frequency.
The name coho stands for coherent oscillator to signify that it is the reference signal that has the phase of the transmitter signal.
Coherency with the transmitted signal is obtained by using the sum of the coho & the stalo signals as the input signal to the power amplifier.
Thus the transmitter frequency is the sum of the stalo frequency fl & the coho frequency fc.
This is accomplished in the mixer shown on the upper right side of Fig.3.7.
The combination of the stalo & coho is sometimes called the receiver-exciter portion of the MTI radar.

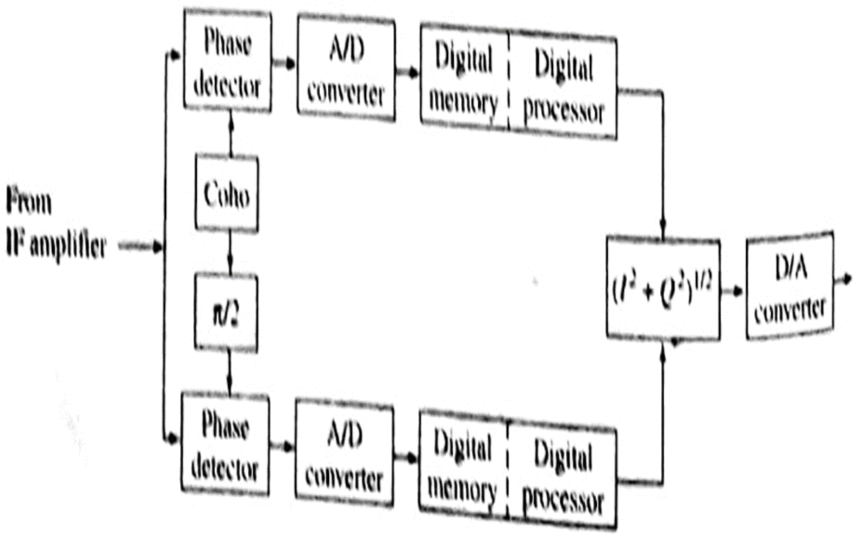
Using the receiver stalo & coho to also generate the transmitter signal insures better stability than if the functions were performed with 2 different sets of oscillators.
The output of the phase detector is the input to the delay-line canceler, as in Fig.3.6.
The delay-line canceler acts as a high-pass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

Q5) 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 $Ad \cos (2\pi fdt + \varphi 0)$ & 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.



Q6) Solution :

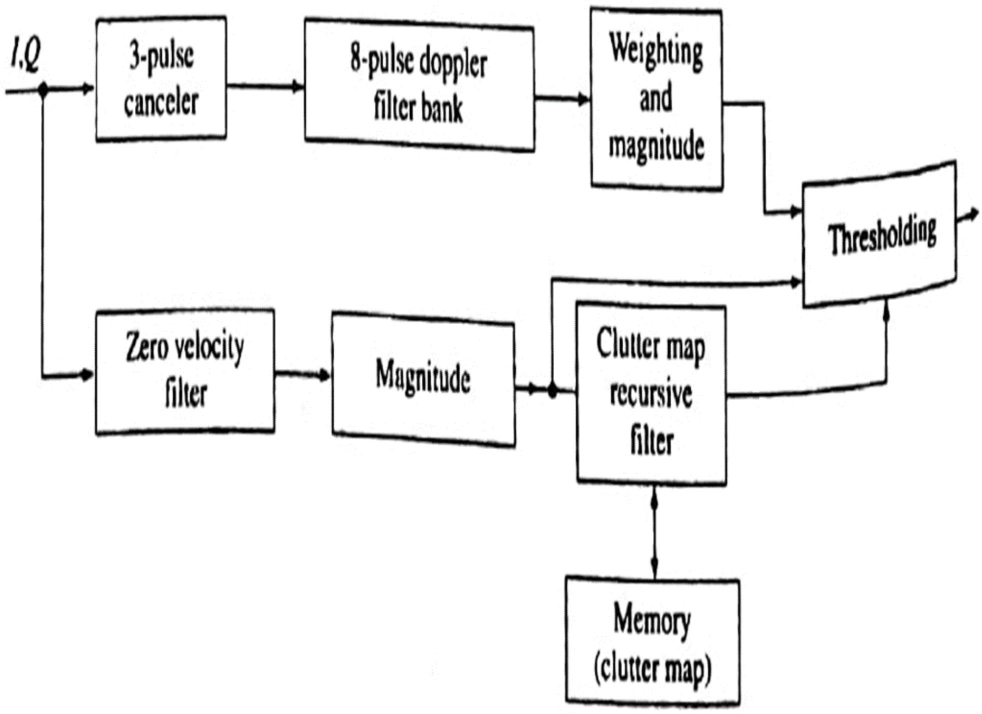
The Moving Target Detector (MTD) is an example of an MTI processing system.
It takes advantage of the various capabilities offered by digital techniques.
These techniques are used to produce improved detection of moving targets in clutter.
It was originally developed by the MIT Lincoln Laboratory for 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.



Q7) Solution :

Clutter Attenuation :
The other limitation of the single delay-line canceler is insufficient attenuation of clutter.
This results from the finite width of the clutter spectrum.
The single delay-line canceler does what it is supposed to do, which is to cancel stationary clutter with zero doppler shift.
In the "real world," however, the clutter spectrum has a finite width due to such things as –
The internal motions of the clutter.
Instabilities of the stalo & coho oscillators.
Other imperfections of the radar & its signal processor.
The finite signal duration.

The consequences of a finite-width clutter spectrum can be seen from Fig.3.10.
The frequency response of the single delay-line canceler shown by the solid curve encompasses a portion of the clutter spectrum.
Therefore, clutter will appear in the output.
The greater the standard deviation, the greater the amount of clutter that will be passed by the filter to interfere with moving target detection.

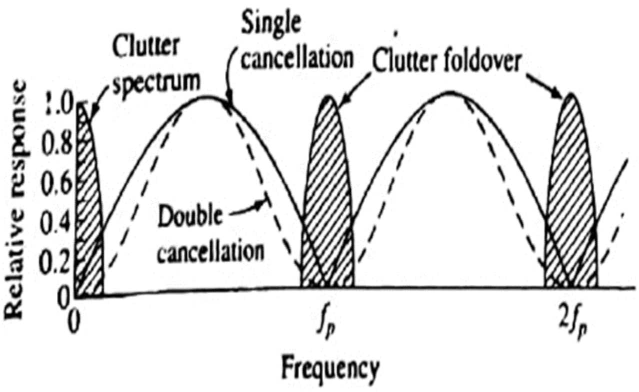


Figure 3.10 Relative frequency response of a single delay-line canceler (solid curve) and the double delay-line canceler (dashed curve), along with the frequency spectrum of the clutter (shaded area). Note the clutter spectrum is folded over at the prf and its harmonics because of the sampled nature of a pulse radar waveform.

Improvement Factor :

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]

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

Doppler Shift :

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.

Blind Speed & Its Eradication :

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,

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

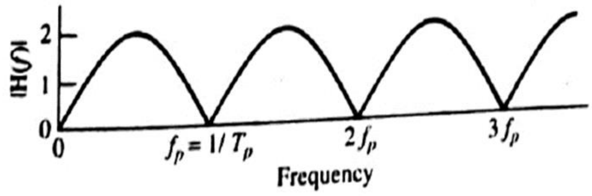


Figure 3.8 Magnitude of the frequency response $|H(f)|$ of a single delay-line canceler as given by Eq. (3.10), where f_p = pulse repetition frequency and $T_p = 1/f_p$.

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 :
Operate the radar at long wavelengths (low frequencies).
Operate with a high pulse repetition frequency.
Operate with more than one pulse repetition frequency.
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.

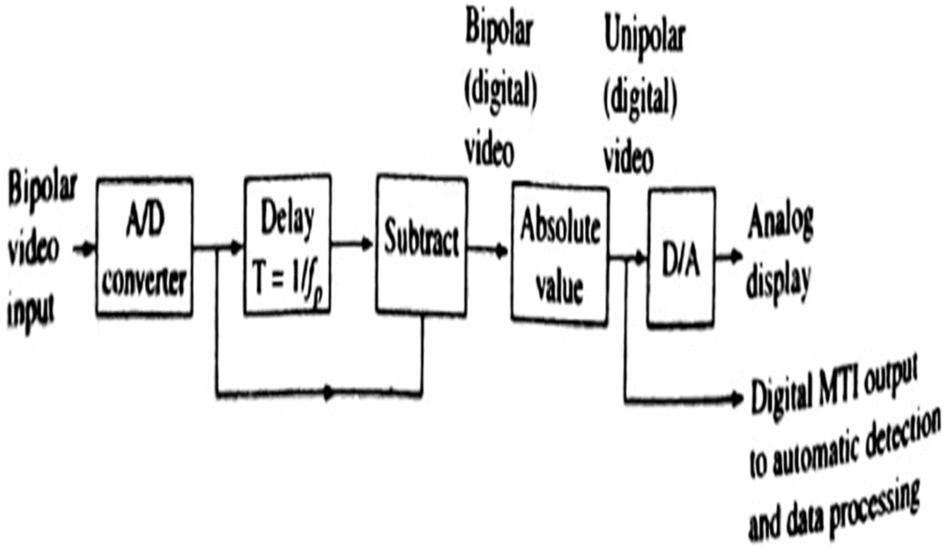
Q8) Solution :

Subtraction of the echoes from 2 successive sweeps is accomplished in a delay-line canceler.
This is indicated by the diagram of Fig.3.6.
The output of the MTI receiver is digitized & is the input to the delay line canceler.
This delay line canceler performs the role of a doppler filter.
The delay T is achieved by storing the radar output from one pulse transmission, or sweep.
This is done in a digital memory for a time equal to the pulse repetition period so that T=TP=1/fP.

The output obtained after subtraction of 2 successive sweeps is bipolar (digital) video.
This is because the clutter echoes in the output contain both positive & negative amplitudes [as can be seen from Eq.(3.6) when fd=0].
It is usually called video, even though it is a series of digital words rather than an analog video signal.
The absolute value of the bipolar video is taken, which is then unipolar video.

Unipolar video is needed if an analog display is used that requires positive signals only.
The unipolar digital video is then converted to an analog signal.
This is done by the digital-to-analog (D/A) converter if the processed signal is to be displayed on a PPI (plan position indicator).
Alternatively, the digital signals may be used for automatically making the detection decision & for further data processing, such as automatic tracking and/or target recognition.
The term delay-line canceler was originally applied when analog delay lines (usually acoustic) were used in the early MTI radars.
Though analog delay lines are replaced by digital memories, the name delay-line canceler is still used to describe the operation of Fig.3.6.

Figure 3.6 Block diagram of a single delay-line canceler.



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

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

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

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

$$H(f) = 2 \sin (\pi f_d T_p)$$ [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,

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

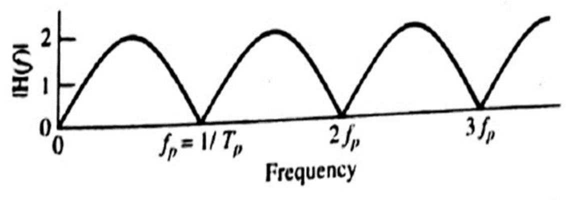


Figure 3.8 Magnitude of the frequency response $|H(f)|$ of a single delay-line canceler as given by Eq. (3.10), where f_p = pulse repetition frequency and $T_p = 1/f_p$.

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.

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END