CHAPTER 3
RECEIVERS

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3.1 THE CONFIGURATION OF A RADAR RECEIVER

The function of a radar receiver is to amplify the echoes of the radar transmission and to filter them in a manner that will provide the maximum discrimination between desired echoes and undesired interference. The interference comprises not only the noise generated in the radar receiver but also energy received from galactic sources, neighboring radars and communication equipment, and possibly jammers. The portion of the radar's own radiated energy that is scattered by undesired targets (such as rain, snow, birds, insects, atmospheric perturbations, and chaff) may also be classed as interference. Where airborne radars are used for altimeters or mapping, other aircraft are undesired targets, and the ground is the desired target. More commonly, radars are intended for detection of aircraft, ships, surface vehicles, or personnel, and the reflection from sea or ground is classified as clutter interference.

The boundaries of the radar receiver must be defined arbitrarily. This chapter will consider those elements shown in Fig. 3.1 as the receiver. The input signal is derived from the duplexer, which permits a single antenna to be shared between transmitter and receiver. Some radar antennas include low-noise amplifiers prior to forming the receive beams; although these are generally considered to be antenna rather than receiver elements, they will be discussed in this chapter.

The receiver filters the signal to separate desired echoes from interference in many ways, but some elements of such processing are covered by other chapters because of the depth of treatment required:

Chapter 14 describes continuous-wave (CW) and FM-CW radars; the discussion here will be confined to receivers for pulse radars, the dominant form. Low-pulse-repetition-frequency (PRF) pulse radars transmit a burst of energy and listen for echoes between transmissions. Their outstanding advantage is that neither leakage from the transmitter nor the very strong echoes from close-range clutter occur at the same instant as reception of weak echoes from long-range targets. The delay of the pulsed echo also provides an instantaneous measurement of range.

Chapters 15 to 17 relate to discrimination of desired targets from interference on the basis of velocity or the change in phase from one pulse to the next; the

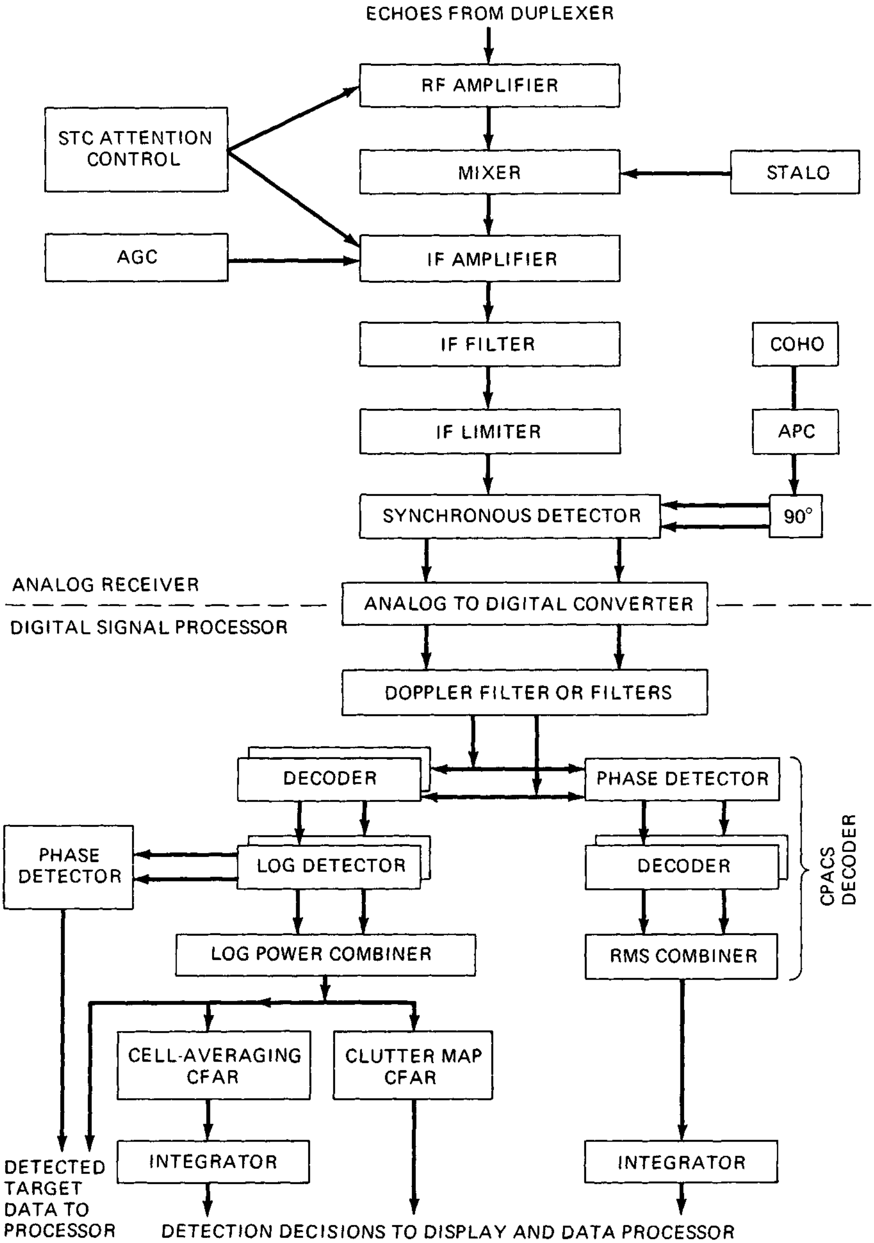


FIG. 3.1 General configuration of a radar receiver.

receiver discussed here serves only to provide the individual pulse signals in proper form for such doppler filtering. Section 3.12 will discuss data distortions in the synchronous detector or analog-to-digital (A/D) converter which affect these doppler filters.

Chapter 10 deals with pulse compression, and only brief mention of its ability to aid in the discrimination process will be included here. Physically, decoding such echoes may occur as part of the intermediate-frequency (IF) filter, typically with surface acoustic-wave (SAW) devices or with digital correlators, either preceding or following doppler filtering. In Fig. 3.1, pulse compression is provided by the cascade combination of IF filter and digital decoders (correlators) after doppler filtering.

Similarly, Chaps. 18 and 20 describe tracking radars and height finding radars, but the peculiarities of the receivers required to perform these functions will be mentioned briefly.

The purpose of Fig. 3.1 is to illustrate the usual sequence of processing functions that may occur in any radar receiver and the variety of possible outputs, although no radar receiver will include all these functions or provide all these outputs.

Virtually all radar receivers operate on the superheterodyne principle shown in Fig. 3.1. The echo, after modest amplification, is shifted to an intermediate frequency by mixing with a local-oscillator (LO) frequency. More than one conversion step may be necessary to reach the final IF, generally between 0.1 and 100 MHz, without encountering serious image- or spurious-frequency problems in the mixing process. Not only is amplification at IF less costly and more stable than at microwave frequency, but the wider percentage bandwidth occupied by the desired echo simplifies the filtering operation. In addition, the superheterodyne receiver can vary the LO frequency to follow any desired tuning variation of the transmitter without disturbing the filtering at IF. These advantages have been sufficiently powerful that competitive forms of receivers have virtually disappeared; only the superheterodyne receiver will be discussed in any detail.

Other receiver types include the superregenerative, crystal video, and tuned radio frequency (TRF). The superregenerative receiver is sometimes employed in radar-beacon applications because a single tube may function as both transmitter and receiver and because simplicity and compactness are more important than superior sensitivity. The crystal video receiver also is simple but of poor sensitivity. The TRF receiver uses only RF and video amplification; although its noise temperature may be low, its sensitivity is poor because optimum-bandwidth filtering of the usual radar echo spectrum is generally impractical to achieve. Only for radars that radiate a relatively wide-percentage-bandwidth signal is filtering practical.

3.2 NOISE AND DYNAMIC-RANGE CONSIDERATIONS

Receivers generate internal noise which masks weak echoes being received from the radar transmissions. This noise is one of the fundamental limitations on the radar range, and for this reason the subject has been treated in Sec. 2.5. The analysis of radar sensitivity is facilitated if the noise contribution of each element of the system is expressed as a *noise temperature* rather than as a *noise factor* or *noise figure*; these terms are defined and interrelated in Sec. 2.5.

The noise temperature of the radar receiver has been reduced to the point that it no longer represents a dominant influence in choosing between available alternatives. It is a paradox that a noise parameter is usually the first characteristic specified for a radar receiver, yet few radars employ the lowest-noise receiver available because such a choice represents too great a sacrifice in some other characteristic.

Cost is rarely a consideration in rejecting a lower-noise alternative. A reduction of requirements for antenna gain or transmitter power invariably produces cost savings far in excess of any added cost of a lower-noise receiver. More vital performance characteristics generally dictate the choice of receiver front end:

1. Dynamic range and susceptibility to overload
2. Instantaneous bandwidth and tuning range
3. Phase and amplitude stability
4. Cooling requirements

A direct compromise must be made between the noise temperature and the dynamic range of a receiver. The introduction of an RF amplifier in front of the mixer necessarily involves raising the system noise level at the mixer to make the noise contribution of the mixer itself insignificant. Even if the RF amplifier itself has more than adequate dynamic range, the mixer dynamic range has been compromised, as indicated below:

Ratio of front-end noise to mixer noise	6 dB	10 dB	13.3 dB
Sacrifice in mixer dynamic range	7 dB	10.4 dB	13.5 dB
Degradation of system noise temperature by mixer noise	1 dB	0.4 dB	0.2 dB

Definitions. Section 2.5 defined the noise parameters of a receiver in simple terms. Dynamic range, which represents the range of signal strength over which the receiver will perform as expected, is more difficult to define. It requires the specification of three parameters:

1. *Minimum signal of interest:* This is usually defined as the input signal that produces unity signal-to-noise ratio (*SNR*) at the receiver output. Occasionally, a minimum-detectable-signal definition is employed.

2. *Allowable deviation from expected characteristic:* The maximum signal is one that will cause some deviation from expected performance. Linear receivers usually specify a 1 dB decrease in incremental gain (the slope of the output-versus-input curve). Limiting or logarithmic receivers must define an allowable error in their outputs. Gain-controlled receivers must distinguish between instantaneous dynamic range and that achieved partly as a result of programmed gain variation.

3. *Type of signal:* Three types of signals are of general interest in determining dynamic-range requirements: distributed targets, point targets, and wideband-noise jamming. If the radar employs a phase-coded signal, the elements of the receiver preceding the decoder will not restrict the dynamic range of a point target as severely as they will distributed clutter; the bandwidth-time product of the coded pulse indicates the added dynamic range that the decoder will extract from point targets. Conversely, if the radar incorporates an excessively wide-

bandwidth RF amplifier, its dynamic range to wideband-noise interference may be severely restricted.

When low-noise amplifiers (LNAs) are included in the antenna, prior to forming the receive beams, the sidelobe levels achieved are dependent upon the degree to which gain and phase characteristics are similar in all LNAs. Dynamic range has an exaggerated importance in such configurations because matching nonlinear characteristics is impractical. The effect of strong interference [mountain clutter, other radar pulses, or electronic countermeasures (ECM)], entering through the sidelobes, will be exaggerated if it exceeds the dynamic range of the LNAs because sidelobes will be degraded. The LNAs are wideband devices, vulnerable to interference over the entire radar operating band and often outside this band; although off-frequency interference is filtered in subsequent stages of the receiver, strong interference signals can cause clutter echoes in the LNA to be distorted, degrading the effectiveness of doppler filtering and creating false alarms. This phenomenon is difficult to isolate as the cause of false alarms in such radars owing to the nonrepetitive character of many sources of interference.

Evaluation. A thorough evaluation of all elements of the receiver is necessary to prevent unanticipated degradation of noise temperature or dynamic range. Inadequate dynamic range makes the radar receiver vulnerable to interference, which can cause saturation or overload, masking or hiding the desired echoes. A tabular format for such a computation (a typical example of which is shown as Table 3.1) will permit those components that contribute significant noise or restrict the dynamic range to be quickly identified. "Typical" values are included in the table for purposes of illustration.

One caution is required in using Table 3.1. The dynamic range of each component is computed by comparing the maximum signal and system noise levels at the output of each component. The assumption inherent in this method is that all filtering (bandwidth reduction and decoding) by this component is accomplished prior to any saturation. It is important to treat those stages of the receiver that provide significant filtering as separate elements; if multiple stages are lumped into a single filter, this assumption may be grossly in error.

3.3 BANDWIDTH CONSIDERATIONS

Definitions. The instantaneous bandwidth of a component is the frequency band over which the component can simultaneously amplify two or more signals to within a specified gain (and sometimes phase) tolerance. The tuning range is the frequency band over which the component may operate without degrading the specified performance if suitable electrical or mechanical controls are adjusted.

Important Characteristics. The environment in which a radar must operate includes many sources of electromagnetic radiation, which can mask the relatively weak echoes from its own transmission. The susceptibility to such interference is determined by the ability of the receiver to suppress the interfering frequencies if the sources have narrow bandwidth or to recover quickly if they are more like impulses in character. One must be concerned with the response of the receiver in both frequency and time domains.

Generally, the critical response is determined in the IF portion of the receiver; this will be discussed in Sec. 3.7. However, one cannot ignore the RF portion of the receiver merely by making it have wide bandwidth. Section 3.2 discussed

TABLE 3.1 Noise and Dynamic-Range Characteristics

		Antenna	Trans-line	RF amplifier	Mixer	Filter	Log detector
Noise temperature of component	K			520	1300	300	24K
Gain of component *	dB		-1.0	25	-6	15	
Total gain to input	dB			-1.0	24	18	33
Noise-temperature contribution referred to antenna	System 838 K	80	75	660	6	5	12
	29.3 dB K						
Overall RX bandwidth	63.0 dBHz						
	92.3						
Boltzmann's constant	-198.6						
Narrowband noise level †	-106.3 dBm	-106	-107	-82	-88	73	(-73)
Maximum signal capability †	dBm			-5	-16	+5	(+7)
Dynamic range to distributed targets †	dB			77	72	78	(80)
Bandwidth x time of point target †	dB	11	11	11	11	0	0
Dynamic range to point target †	dB			88	83	78	(80)
Bandwidth of receiver †	MHz			200	100	2	2
Ratio to overall receiver bandwidth †				100	50	1	1
Wideband-noise vulnerability †	dB			20	17	0	0
Dynamic range to wideband noise †	dB			57	55	78	(80)

* CW output-CW input on the center frequency, not coded pulse.

† At the output terminal of the designated component except where indicated by parentheses (at the input terminal of a nonlinear device).

how excessively wide bandwidth can penalize dynamic range if the interference is wideband noise. Even more likely is an out-of-band source of strong interference (TV station or microwave communication link) which, if allowed to reach this point, can either overload the mixer or be converted to IF by one of the spurious responses of the mixer.

Ideal mixers in a superheterodyne receiver act as multipliers, producing an output proportional to the product of the two input signals. Except for the effect of nonlinearities and unbalance, these mixers produce only two output frequencies, equal to the sum and the difference of the two input frequencies. Product mixers, although common at intermediate frequencies, are not generally available for RF conversion down to IF, and diode mixers are most commonly employed. The frequency-conversion properties of the diode are produced by its nonlinear characteristics. If its characteristic is defined by a power series, only the square-law term produces the desired conversion. The other terms produce spurious products, which represent an unwanted ability to convert off-frequency signals to the IF of the receiver. The efficiency of conversion of these unwanted frequencies, except for the image frequency, is sufficiently poor that the system noise temperature is not significantly degraded, but the mixer is vulnerable to strong out-of-band interference. The best radar receiver is one with the narrowest RF instantaneous bandwidth commensurate with the radiated spectrum and hardware limitations, and with good frequency and impulse responses.

A wide tuning range provides a flexibility to escape interference, but if the interference is intentional (jamming), change in frequency on a pulse-to-pulse basis may be required. Such frequency agility can be achieved by using switchable microwave filters or electronically tuned yttrium iron garnet (YIG) filters to restrict the instantaneous bandwidth. Each involves some insertion loss, another sacrifice in noise temperature to achieve more vital objectives.

3.4 RECEIVER FRONT END

Configuration. The radar *front end* consists of a bandpass filter or bandpass amplifier followed by a downconverter. The radar frequency is downconverted to an intermediate frequency, where filters with suitable bandpass characteristics are physically realizable. The mixer itself and the preceding circuits are generally relatively broadband. Tuning of the receiver, between the limits set by the preselector or mixer bandwidth, is accomplished by changing the LO frequency.

Effect of Characteristics on Performance. Noncoherent pulse radar performance is affected by front-end characteristics in three ways. Noise introduced by the front end restricts the maximum range. Front-end saturation on strong signals may limit the minimum range of the system or the ability to handle strong interference. Finally, the front-end spurious characteristic affects the susceptibility of off-frequency interference.

Coherent radar performance is even more affected by spurious mixer characteristics. Range and velocity accuracy is degraded in the pulse doppler radar; stationary-target cancellation is impaired in MTI (moving-target indication) radar; and range sidelobes are raised in high-resolution pulse compression systems.

Spurious Distortion of Radiated Spectrum. It is a surprise to many radar engineers that components of the radar receiver can cause degradation of the radiated transmitter spectrum, generating harmonics of the carrier frequency or spurious doppler spectra, both of which are often required to be 50 dB or more below the carrier. Harmonics can create interference in other electronic equipment, and their maximum levels are specified by the National Telecommunications and Information Administration (NTIA) and MIL-STD-469. Spurious doppler spectra levels are dictated by requirements to suppress clutter interference through doppler filtering.

Harmonics are generated by any component which is nonlinear at the power level created by the transmitter and which passes those harmonics to the antenna. Gaseous or diode receiver-protectors are designed to be nonlinear during the transmitted pulse and reflect the incident energy back toward the antenna. Isolators or circulators are often employed to absorb most of the reflected fundamental, but they are generally much less effective at the harmonics. Moreover, these ferrite devices are nonlinear in themselves and can generate harmonics.

Harmonic filters are included in most radars but often are improperly located to perform adequately. It is useless to locate the harmonic filter between the transmitter and the duplexer if the latter generates unacceptable harmonic levels itself; the filter must be located between the antenna and the duplexer.

Spurious doppler spectra are created by any process which does not reoccur precisely on each transmitted pulse. Gaseous receiver-protectors ionize under transmitter power levels, but there is some small statistical variation in the initiation of ionization on the leading edge of the pulse and in its subsequent development. In radars demanding high clutter suppression (in excess of 50 dB), it has sometimes been found necessary to prevent this variable reflected power from being radiated by use of both a circulator and an isolator in the receive path.

Spurious Responses of Mixers

Mathematical Mixer Model. The power-series representation of the mixer is perhaps the most useful in predicting the various spurious effects that are often noted. The current i flowing in a nonlinear resistance may be represented by a power series in the voltage V across the resistor terminals:

$$i = a_0 + a_1 V + a_2 V^2 + a_3 V^3 + \cdots + a_n V^n \tag{3.1}$$

The voltage applied to the mixer is the sum of the LO voltage $V_1 e^{j\omega_1 t}$ and the signal voltage $V_2 e^{j\omega_2 t}$:

$$V = V_1 e^{j\omega_1 t} + v_2 e^{j\omega_2 t} \tag{3.2}$$

When V from Eq. (3.2) is substituted into Eq. (3.1) and the indicated operations performed, the spectral characteristics are predicted.

Mixer Spurious-Effects Chart. The results of these calculations have been tabulated in several forms to show the system designer at a glance which combinations of input frequencies and bandwidths are free of strong low-order spurious components. The most useful form of the mixer chart[1] is shown in Fig. 3.2. The heavy line shows the variation of normalized output frequency $(H - L)/H$ with normalized input frequency L/H. This response is caused by the first-order mixer product $H - L$, which originates mainly from the square-law term in the power-series representation. All other lines on the chart define spurious effects arising from the cubic and higher-order terms in the power series. To simplify use of the chart, the higher input frequency is designated by H and the lower input frequency by L.

Seven particularly useful regions have been outlined on the chart. Use of the chart is illustrated by means of the region marked A, which represents the widest available spurious-free bandwidth centered at $L/H = 0.63$. The available RF passband is from 0.61 to 0.65, and the corresponding IF passband is from 0.35 to 0.39. However, spurious IF frequencies of 0.34 ($4H - 6L$) and 0.4 ($3H - 4L$) are generated at the extremes of the RF passband. Any extension of the instantaneous RF bandwidth will produce overlapping IF frequencies, a condition that is not corrected by IF filtering. The $4H - 6L$ and $3H - 4L$ spurious frequencies,

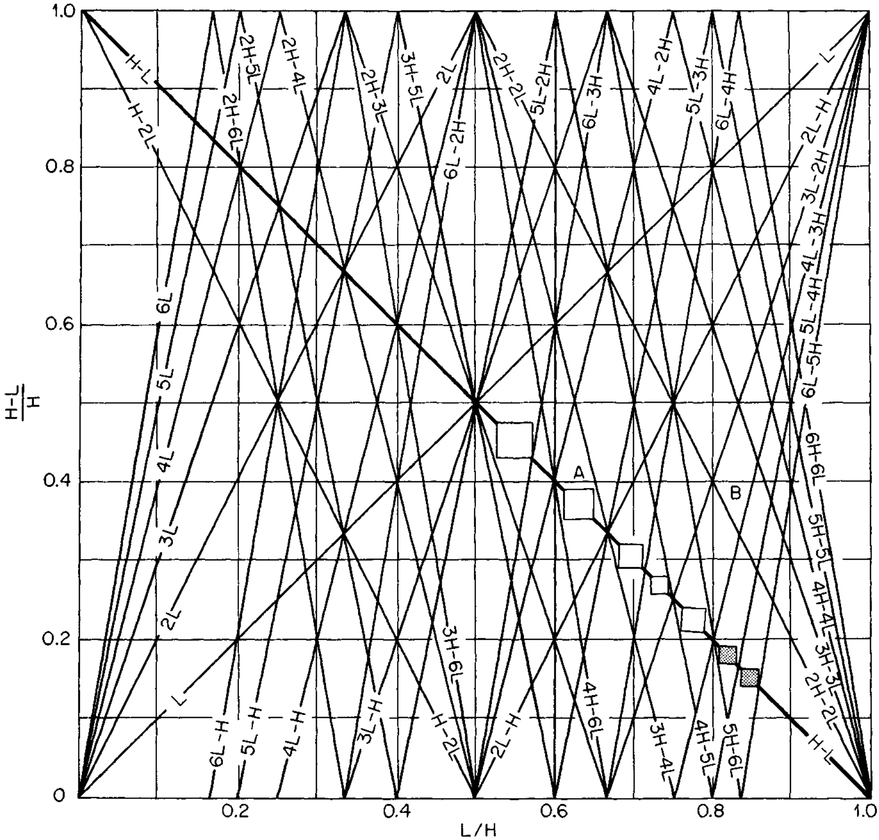


FIG. 3.2 Downconverter spurious-effects chart. H = high input frequency; L = low input frequency.

like all spurious IF frequencies, arise from cubic or higher-order terms in the power-series model of the mixer.

The available spurious-free bandwidth in any of the designated regions is roughly 10 percent of the center frequency or $(H - L)/10H$. Thus receivers requiring a wide bandwidth should use a high IF frequency centered in one of these regions. For IF frequencies below $(H - L)/H = 0.14$ the spurious frequencies originate from extremely high-order terms in the power-series model and are consequently so low in amplitude that they can usually be ignored. For this reason, single-conversion receivers generally provide better suppression of spurious responses than double-conversion receivers. The rationale for a choice of double conversion should always be validated.

The spurious-effects chart also demonstrates spurious input responses. One of the stronger of these occurs at point B, where the $2H - 2L$ product causes a mixer output in the IF passband with an input frequency at 0.815. All the products of the form $N(H - L)$ produce potentially troublesome spurious responses. These frequencies must be filtered at RF to prevent their reaching the mixer.

A spurious input response not predicted by the chart occurs when two or more off-frequency input signals produce by intermodulation a third frequency that lies within the RF passband. This effect is caused by quartic and higher-order even terms in the series. Its effect will be noted, for example, when

$$\frac{2H - L_1 - L_2}{H} = \frac{H - L}{H} \tag{3.3}$$

Intermodulation is reduced in some mixer designs by forward-biasing the mixer diodes to reduce the higher-order curvature.

The Balanced Mixer. The mixer model and the spurious-effects chart predict the spectral characteristics of a single-ended mixer. In the balanced-mixer configuration these characteristics are modified by symmetry. The two most common forms of balanced-mixer configuration are shown by Fig. 3.3*a* and *b*.

The configuration of Fig. 3.3*a* suppresses all spurious IF frequencies and spurious RF responses derived from even harmonics of the *signal* frequency. For the case where the subtraction is not obtained by a time delay, the LO frequency and *all* its harmonics are suppressed at the signal input port. Also of importance, noise sidebands of the LO which are converted to IF frequency are suppressed at the mixer IF port.

The configuration of Fig. 3.3*b* suppresses all spurious IF frequencies and spurious responses derived from even harmonics of the LO frequency. For the case where the RF phase shift is not obtained by a time delay, the LO frequency and

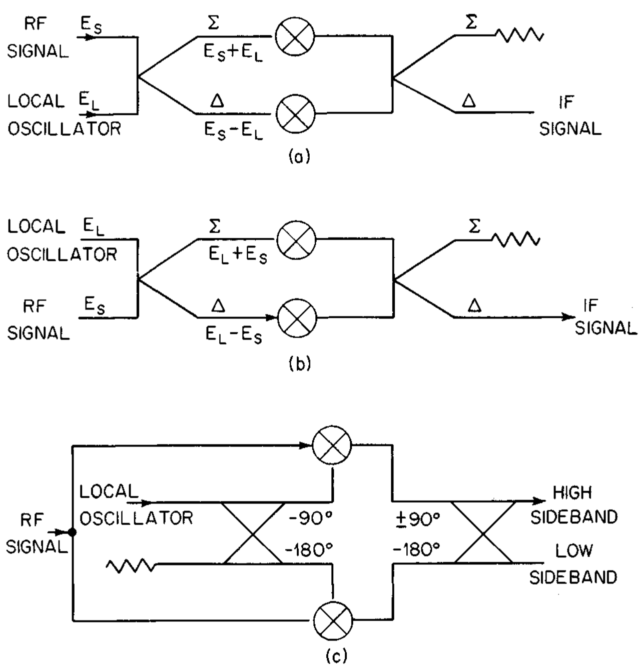


FIG. 3.3 (*a*) Balanced mixer with an inverted signal. (*b*) Balanced mixer with an inverted LO. (*c*) Image-reject mixer.

its *odd* harmonics are suppressed at the signal input port. Noise sidebands of the LO converted to IF *are not* rejected by this configuration, however.

Image-Reject Mixer. The single-ended mixer has two input responses which are derived from the square-law term in the power series. The responses occur at points above and below the LO frequency where the frequency separation equals the IF. The unused response, known as the image, is suppressed by the image-reject or single-sideband mixer shown in Fig. 3.3*c*. The RF hybrid produces a 90° phase differential between the LO inputs to the two mixers (which may be balanced mixers). The effect of this phase differential on the IF outputs of the mixers is a +90° shift in one sideband and a −90° shift in the other. The IF hybrid, adding or subtracting another 90° differential, causes the high-sideband signals to add at one output port and to subtract at the other. Where wide bandwidths are involved, the IF hybrid is of the all-pass type.

Characteristics of Amplifiers and Mixers

Noise Temperature. The most frequently cited figure of merit for a mixer or amplifier is its noise figure. However, the concept of noise temperature has proved more useful. Chapter 2 defines the usage of these parameters in determining the detectability of signals in a noise background.

Dynamic Range. A second useful figure of merit of the front-end device is the dynamic range from rms noise to the signal level that causes 1 dB compression in dynamic gain. Since the rms noise is dependent on the IF bandwidth, the effective dynamic range decreases with increasing IF bandwidth.

The balanced-diode mixer exhibits the largest dynamic range for a given IF bandwidth. However, the dynamic range of the mixer preceded by a low-noise amplifier will be reduced in proportion to the gain of the amplifier. Thus noise performance and dynamic range cannot be simultaneously optimized. A solution to this problem may come in the form of an active converter.[2,3]

3.5 LOCAL OSCILLATORS

Functions of the Local Oscillator. The superheterodyne receiver utilizes one or more local oscillators and mixers to convert the echo to an intermediate frequency that is convenient for filtering and processing operations. The receiver can be tuned by changing the first LO frequency without disturbing the IF section of the receiver. Subsequent shifts in intermediate frequency are often accomplished within the receiver by additional LOs, generally of fixed frequency.

Pulse-amplifier transmitters also use these same LOs to generate the radar carrier with the required offset from the first local oscillator. Pulsed oscillator transmitters, with their independent "carrier" frequency, use automatic frequency control (AFC) to maintain the correct frequency separation between the carrier and first LO frequencies.

In many early radars, the only function of the local oscillators was conversion of the echo frequency to the correct intermediate frequency. The majority of modern radar systems, however, coherently process a series of echoes from a target. The local oscillators act essentially as a timing standard by which the echo delay is measured to extract range information, accurate to within a small fraction of a wavelength. The processing demands a high degree of phase stability throughout the radar. Although these processing techniques are described elsewhere (Chaps. 15 to 17 and 21), they determine the basic stability requirements of the receiver.

The first local oscillator, generally referred to as a stable local oscillator (stalo), has a greater effect on processing performance than the transmitter. The final local oscillator, generally referred to as a coherent local oscillator (coho), is often utilized for introducing phase corrections which compensate for radar platform motion or transmitter phase variations.

Stalo Instability. The stability requirements of the stalo are generally defined in terms of a tolerable phase-modulation spectrum. Sources of unwanted modulation are mechanical or acoustic vibration from fans and motors, power supply ripple, and spurious frequencies and noise generated in the stalo. In general, the tolerable phase deviation decreases with increasing modulation frequency because the doppler filter is less efficient in suppressing the effects. In a radar having two-pulse MTI, there is a linear relationship between the tolerable phase deviation and the period of the modulation. Their ratio is the allowable FM (frequency modulation) or *short-term frequency stability* sometimes encountered in the literature. This parameter does not adequately define the phase-stability requirements for pulse doppler or MTI radars where more than two pulses are coherently processed.

The phase-modulation spectrum of the stalo may be measured and converted into the MTI improvement factor limitation, which is dependent on range to the clutter and the characteristics of the two cascaded filters in the radar receiver. This conversion process involves three steps, described below.

It should be noted that some spectrum analyzers do not distinguish between frequencies below the desired stalo frequency and those above; their response is the sum of the power in the two sidebands at each designated modulation frequency. This is of no consequence in MTI radars which have equal response to positive and negative doppler frequencies. In radars using doppler filters unsymmetrical about zero doppler, it is necessary to assume that the stalo spectrum measured is symmetrical, generally a valid assumption. The examples shown subsequently employ measured data from this type of double-sideband (DSB) spectrum analyzer. If a single-sideband (SSB) spectrum analyzer is available, positive and negative modulation-frequency components can be measured separately and analyzed without any assumption of symmetry. It is essential that the measured data be defined as SSB or DSB, since there is a 3 dB difference in the two forms of data.

Range Dependence. Most modern radars use the stalo to generate the transmitted pulse as well as to shift the frequency of the received echoes. The transmitters are power amplifiers (traveling-wave tubes, klystrons, twystrons, crossed-field amplifiers, solid-state amplifiers, etc.) rather than oscillators (magnetron, etc.). It is this double use of the stalo that introduces a dependence on range of the clutter and exaggerates the effect of certain unintentional phase-modulation components by 6 dB, the critical frequencies being those which change phase by odd multiples of 180° during the time period between transmission and reception of the clutter echo from a specified range. At these critical frequencies, a maximum positive phase deviation on transmission changes to a maximum negative deviation at the time of reception, doubling the undesired phase modulation of the echo at IF.

Figure 3.4 shows this range-dependent filter characteristic, which may be expressed mathematically as

$$\text{dB} = 10 \log 4 \sin^2 (2\pi f_m R/c) = 10 \log 4 \sin^2 (\pi f_m t) \tag{3.4}$$

where f_m = modulation frequency, Hz
R = range, m
c = propagation velocity, 3×10^8 m/s
t = time delay = $2R/c$

A short time delay can tolerate much higher disturbance at low modulation frequencies, as illustrated by the two cases in Fig. 3.4. Consequently, stalo stability needs to be computed for several time delays.

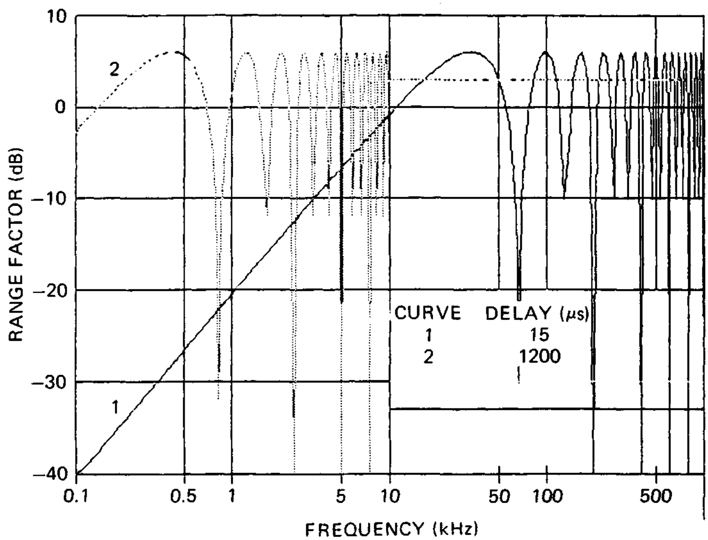


FIG. 3.4 Effect of range delay on clutter cancellation.

For example, stalo phase modulation caused by power supply ripple at 120 Hz creates nearly equivalent phase modulation of a clutter echo from nearly 100-nmi range (delay of 1200 μs resulting in 0 dB range factor). Phase modulation of a clutter echo with a range delay of 15 μs is about 38 dB less than the stalo phase modulation because the stalo phase has changed only slightly in this short time interval; the phase added by the stalo to the transmitted pulse is nearly the same as the phase subtracted from the received echo in the mixer.

Adding the decibel values of the measured stalo spectrum and the range-dependent effect at each modulation frequency provides the spectrum of undesired doppler modulation at the output of the mixer.

Receiver Filtering. Subsequent stages of the radar receiver have responses which are functions of the doppler modulation frequency; so the output residue spectrum can be obtained by adding the decibel responses of these filters to the preceding spectrum at the mixer.

The receiver contains two cascaded filters: an optimum-bandwidth filter at IF and a doppler filter, generally implemented digitally in modern radars. The example illustrated in Fig. 3.5 includes a gaussian filter at IF with a 3 dB bandwidth of 1.6 MHz and a four-pulse MTI with variable interpulse periods and time-

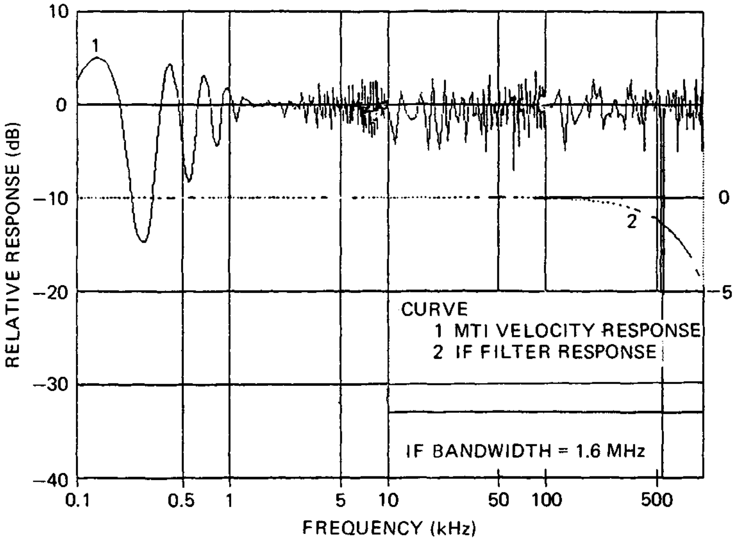


FIG. 3.5 Frequency responses of radar receiver.

varying weights. This MTI exaggerates certain stalo modulation frequencies by as much as 5 dB more than average. Note that for this analysis to be correct the MTI velocity response must be scaled to have 0 dB gain to noise, not to an optimum doppler frequency.

Integration of Residue Power. The MTI improvement factor limitation of the stalo may be expressed as the ratio of the stalo power to the total power of the echo modulation spectrum it creates at the output of the cascaded filters (Figs. 3.4 and 3.5).

Figure 3.6 shows an example of the measured modulation spectrum of a stalo (curve 1) and the effect of 15-μs delay on the clutter residue (curve 2). A computer program can be utilized to alter the measured spectral data, using the filters of Fig. 3.4 and 3.5, and to integrate the total power in the doppler residue spectrum, with the exception of modulation frequencies below 100 Hz, which cannot be measured. The result (51.8 dB) is the MTI improvement factor limitation due to stalo instability.

Figure 3.7 shows the same measured stalo modulation spectrum, subjected to a range delay of 1200 μs. The clutter residue spectrum contains more power at the lower modulation frequencies than Fig. 3.6, but the MTI residue is increased by only 1 dB. Long-range clutter is suppressed to nearly the same degree as short-range clutter.

If the radar utilizes more than one doppler filter, the effect of stalo instability should be calculated for each individually. If an individual filter's doppler response is unsymmetrical, residues from positive and negative doppler bands must be computed separately and added in power.

It should be noted that many textbooks analyze only a simple two-pulse MTI, and the resulting equations for the limitation on MTI improvement factor cannot be employed for more sophisticated doppler filters. This important fact is generally overlooked by the casual reader.

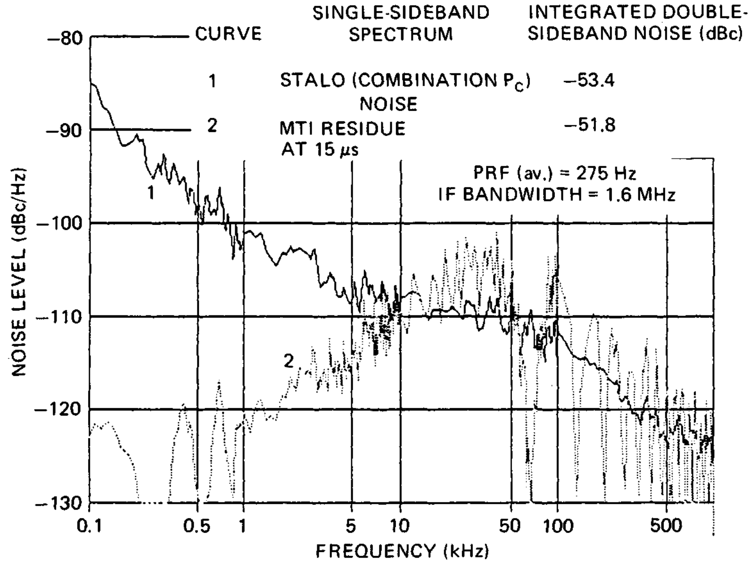


FIG. 3.6 Effect of 15-μs range delay on stalo MTI limitation.

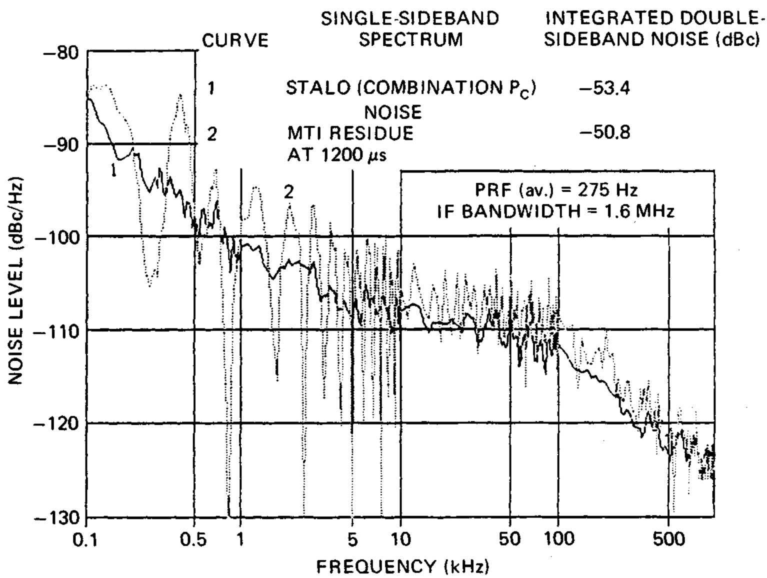


FIG. 3.7 Effect of 1200-μs range delay on stalo MTI limitation.

It should be noted that most textbook analyses assume that the stalo instability is due either to a single modulation frequency or to a combination of white gaussian-noise modulations. Rarely are these assumptions valid for real stalos; so a different method of analysis must be employed.

The computer analysis described measures the colored modulation spectrum of the stalo, modifies it in conformance with the range-dependent effect and receiver filters of the radar, and integrates the output residue power.

It is a valid procedure for determining stalo stability that requires no assumptions, but the resulting value must be considered merely a figure of merit to compare *different* stalos for a *given* application. The same stalo would have a different stability value in another application with different receiver filters.

Coho and Timing Instability. In modern radars using pulsed amplifier transmitters, the coho is rarely a significant contributor to receiver instability. However, in older radars using pulsed oscillator transmitters, the coho must compensate for the random phase of each individual transmitter pulse, and imperfect compensation results in clutter residue at the output of the doppler filters. Compensation is possible only for echoes from the most recent transmission; echoes from prior transmissions (multiple-time-around clutter) cannot be suppressed by doppler filtering in pulsed oscillator radars, and for this reason alone such radars are no longer popular. Those readers interested in the methods employed to compensate the coho in these older radars are referred to Sec. 5.5 of the 1970 edition of this handbook.

When the radar is on a moving platform or when the clutter is moving rain or sea, the frequency of the coho is sometimes varied to try to compensate for this motion, shifting the clutter spectrum back to zero doppler. The servo which accomplishes this task, if properly designed, will introduce insignificant instability under ideal environmental conditions (solely clutter echoes and receiver noise typical of laboratory tests), but the effect of strong moving targets and pulsed interference from other radars can sometimes be serious, shifting the coho frequency from the proper compensation value.

Timing signals for the transmitter and A/D converter are usually generated from the coho, and timing jitter can cause clutter attenuation to be degraded. However, the effect of timing jitter is too complex to predict accurately; so it is rarely measured separately.

Total Radar Instability. The primary sources of radar instability are usually the stalo and the transmitter. If the doppler spectra of these two components are available, either through measurements or through predictions based on similar devices, the convolution of the two-way stalo spectrum (modified by the range-dependent effect) and the transmitter doppler spectrum provides an estimate of the spectrum of echoes from stable clutter, which is then modified by the two receiver filters and integrated to obtain the residue power caused by these two contributors. This power can be larger than the sum of the residues created by each contributor alone. These procedures are employed to diagnose the source of radar instability in an existing radar or to predict the performance of a radar in the design stage.

Measurement of total radar instability can be conducted with the radar antenna searchlighting a stable point clutter reflector which produces an echo close to (but below) the dynamic-range limit of the receiver and doppler filter. Suitable clutter sources are difficult to find at many radar sites, and interruption of rotation of the antenna to conduct such a test may be unacceptable at others; in this case, a microwave delay line can be employed to feed a delayed sample of the transmitter pulse into the receiver. All sources of instability are included in this single measurement except for any contributors outside the delay-line loop. It is important to recognize that timing jitter does not produce equal impact on all parts of the echo pulse and generally has minimal effect on the center of the pulse; so it is essential to collect data samples at a multiplicity of points across the echo, including leading and trailing edges. The total radar instability is the ratio of the sum of the multiplicity of residue powers at the output of the doppler filter to the sum of the powers at its input, divided by the ratio of receiver noise at these locations. Stability is the inverse of this ratio; both are generally expressed in decibels.

In radars with phase-coded transmission and pulse compression receivers, residue may be significant in the range sidelobe region as well as in the compressed pulse, caused by phase modulation during the long transmitted pulse rather than solely from pulse to pulse. Measurement of stability of such radars must employ a very large number of data points to obtain an answer valid for clutter distributed in range.

Radar instability produces predominantly phase modulation of the echoes, and the scanning antenna produces predominantly amplitude modulation; so the combined effect is the sum of the residue powers produced by each individually.

3.6 GAIN-CONTROLLED AMPLIFIERS

Sensitivity Time Control (STC). The search radar detects echoes of widely differing amplitudes, typically so great that the dynamic range of any fixed-gain receiver will be exceeded. Differences in echo strength are caused by differences in radar cross sections, in meteorological conditions, and in range. The effect of range on radar echo strength overshadows the other causes, however.

The radar echo power received from a reflective object varies inversely with the fourth power of the range or propagation time of the radar energy. The effect of range on signal strength impairs the measurement of target size. Yet determination of target size is needed to discriminate against radar echoes from insects, atmospheric anomalies, or birds (which in some cases have radar cross sections only slightly less than that of a jet fighter). Also, many radar receivers exhibit objectionable characteristics when signals exceed the available dynamic range. These effects are prevented by a technique known as sensitivity time control, which causes the radar receiver sensitivity to vary with time in such a way that the amplified radar echo strength is independent of range.

Search radars often employ a cosecant-squared antenna pattern whose gain diminishes with increasing elevation angle. The pattern restricts the power at high elevation angles, since an aircraft at a high elevation angle is necessarily at close range and little power is required for detection. At the high elevation angles, however, the echo power becomes independent of range and varies instead with the inverse fourth power of the altitude. An STC characteristic that is correct for the low-angle radar echoes restricts high-angle coverage. This incompatibility of the STC requirements at the elevation extremes severely limits the usefulness of STC.

The restriction on STC imposed by the cosecant-squared antenna pattern can be reduced by a more realistic radar design philosophy. It is recognized that the antenna must radiate more energy at the high angles than is provided by the cosecant-squared pattern. There are two reasons for this. First, high-angle coverage is limited by clutter from the stronger low-altitude section of the beam rather than from system noise. Second, ECM reduces both the maximum range and the altitude coverage of the radar. Of these, the loss of altitude coverage is the more serious. Both factors have caused the cosecant-squared pattern to be abandoned in favor of one that directs more energy upward.

The advent of stacked-beam radars, which achieve their coverage pattern by use of multiple beams, has liberated STC from the restrictions of the antenna pattern. In these systems, there is one receiver channel for each beam, and STC may be applied to the receiver channels independently. Consequently, the upper-beam receivers may be allowed to reach maximum sensitivity at short ranges, whereas the lower-beam receiver reaches maximum sensitivity only at long range.

Most modern radars generate STC waveforms digitally. The digital commands may be used directly by digital attenuators or converted to voltage or current for control of diode attenuators or variable-gain amplifiers. Digital control permits calibration of each attenuation to determine the difference between the actual attenuation and the command, by injecting a test pulse during *dead time*. This is essential in monopulse receivers which compare the echo amplitudes received in two or more beams simultaneously to accurately determine the target's position in azimuth or elevation. Accurate measurements depend on compensation for any difference in gains of the monopulse receivers.

Readers interested in the methods of generating the analog STC waveforms used in older radars may find descriptions of various methods in Sec. 5.6 of the 1970 edition of this handbook.

Clutter Map Automatic Gain Control. In some radars, mountain clutter can create echoes which would exceed the dynamic range of the subsequent stages of the receiver (A/D converter, etc.) if the STC attenuation at that range allows detection of small aircraft. The spatial area occupied by such clutter is typically a very small fraction of the radar coverage; so AGC is sometimes considered as an alternative to either boosting the STC curve (a performance penalty affecting detectability of small aircraft in areas of weaker clutter or no clutter) or increasing the number of bits of the A/D converter and subsequent processing (an economic penalty).

Clutter map AGC is controlled by a digital map which measures the mean amplitude of the strongest clutter in each map cell of many scans and adds attenuation where necessary to keep the mean amplitude well below saturation. One disadvantage of clutter map AGC is that it degrades detectability of small aircraft over clutter which, in the absence of AGC, would be well below saturation. The scan-to-scan fluctuation of clutter requires a 6 to 10 dB safety margin between the maximum mean level controlled and saturation. Another problem is the vulnerability of the map to pulsed interference from other radars.

Clutter map AGC can serious degrade other critical signal-processing functions, and the following fundamental incompatibilities prevent its successful application to many types of radars:

- Suppression of clutter by doppler filtering is degraded by change of attenuation from one interpulse period to the next.
- Control of false alarms in distributed clutter (rain, sea) can be degraded by change of attenuation from one range sample to the next (see Sec. 3.13).
- Time sidelobes of compressed pulses in radars which transmit coded waveforms are degraded by attenuation variation in range prior to compression. Gradual STC variations can be tolerated, but not large step changes.

Automatic Noise-level Control. AGC is widely employed to maintain a desired level of receiver noise at the A/D converter. As will be described in Sec. 3.11, too little noise relative to the quantization increment of the A/D converter causes a loss in sensitivity; too much noise means a sacrifice of dynamic range. Samples of noise are taken at long range, often beyond the instrumented range of the radar (in dead time), to control the gain by means of a slow-reaction servo. If the radar has RF STC prior to any amplification, it can achieve meaningful dead time by switching in full attenuation; this minimizes external interference with minimal (and predictable) effect on system noise temperature. Most radars employ amplifiers prior to STC; so they cannot attenuate external interference without affecting the noise level which they desire to sense, and the servo must be designed to tolerate pulses from other radars and echoes from rainstorms or mountains at extreme range. This interference occasionally can be of high amplitude but generally has a low duty cycle during a 360° scan; so the preferred servo is one which increments a counter when any sample is below the desired median noise level and decrements the counter when the sample is above that level, independently of how great the deviation is. The most significant bits of the counter control the gain, and the number of bits of lesser significance in the counter dictates the sluggishness of the servo.

3.7 FILTERING

Filtering of the Entire Radar System. The filter provides the principal means by which the receiver discriminates between desired echoes and interference of many types. It may approximate either of two forms: a matched filter, which is a passive network whose frequency response is the complex conjugate of the transmitted spectrum, or a correlation mixer, an active device which compares the received signals with a delayed replica of the transmitted signal. Receiver filters are assumed to have no memory from one transmission to the next; their response is to a single transmission.

Actually, most radars direct a multiplicity of pulse transmissions at a target before the antenna beam is moved to a different direction, and the multiplicity of echoes received is combined in some fashion. The echoes may be processed by an integrator, which is analogous to a matched filter in that ideally its impulse response should match the echo modulation produced by the scanning antenna. The echoes may be applied directly to a PPI, with the viewer visually integrating the dots in an arc which he or she associates with the antenna beamwidth. Various doppler processes (including MTI) may be applied to separate desired from undesired targets. From the radar system standpoint, these are all filtering functions, but they are treated in other chapters of this handbook. The receiver filtering to be discussed here is that associated with separating a single pulse from interference, although the subsequent problem of filtering the train of echoes from a single target dictates the stability of the receiver filter.

At some point in the radar receiver, a detector produces an output voltage which is some function of the envelope of the IF signal. If it provides a linear function, it is termed an envelope detector; logarithmic detectors will be described in Sec. 3.8. The response of a linear detector to weak signals which do not greatly exceed noise level has been extensively analyzed.[4] Various pairs of frequency components of input noise, which may be far removed from the spectrum of the desired echo, can intermodulate to produce a beat-frequency component at the detector output that is within the desired band. Similarly, the noise intermodulation smears some of the signal energy outside the desired band. As a consequence, filtering after envelope detection is less efficient than filtering prior to detection. All postdetection circuitry should have several times the bandwidth of the echo, and predetection filtering should be optimized, as will be described.

Definitions. The reader is cautioned that there are no universally accepted definitions of the terms *pulse duration* and *spectral bandwidth* of the transmitted signal, *impulse response* and *bandwidth* of filters, or the equivalent antenna parameters, *beamwidth* and *spectral bandwidth* caused by scanning. These terms should always be used with clarifying adjectives to define their meaning.

Energy Definitions. For detection of radar echoes against a noise background, the only fundamental parameters are the energy content of the transmitted signal, of the receiver noise, and of the echoes received as the antenna scans past the target. These energy parameters define the width of a rectangular function that has the same peak response and same energy content as the real function. Their only purpose is to relate the peak value of the function to the more vital energy content.

Of these parameters, only the energy of the pulse is easy to measure (average power/PRF), and this may be employed in the radar range equation directly, without distinguishing peak power and "energy" pulse width. The noise or energy bandwidth of a receiver is often employed in theoretical analyses but rarely stated in the tabulation of radar parameters; bandwidth need not even be included in the radar range equation if the receiver approximates a matched filter.

3 dB Definitions. In the interest of making possible direct measurement of parameters from oscilloscope waveforms or pen recordings of these functions, it has been customary to utilize widths measured either at half-power (3 dB) or half-voltage (6 dB) points. For functions that resemble a gaussian pattern, the 3 dB width is a close approximation to the energy width; the receiver bandpass generally fulfills this condition sufficiently to make the 3 dB bandwidth meaningful. Transmitter pulse shapes and spectra generally deviate significantly from gaussian.

6 dB Definitions. Although antenna beamwidths (and number of echoes received) are often specified between the 3 dB points, this actually represents a 6 dB definition of the echo response as the radar antenna scans past the target; in those radars whose transmitting and receiving beams are not identical, the 6 dB points of the two-way pattern are usually specified. Most definitions of pulse shapes include voltage parameters, with rise and fall times being represented by 10 and 90 percent points and pulse duration by 50 percent (6 dB) points. Likewise, filter bandpass characteristics are often defined by their widths at the 6 and 60 dB points. The 6 dB definitions will be the dominant definitions employed in this chapter.

Entirely aside from custom, there are several valid arguments favoring the use of 6 dB parameters. As indicated in Table 3.2, the optimum bandwidth–time product for detection of a pulse in white gaussian noise, with each defined at the 6 dB points, does not deviate significantly from unity for most practical functions. The 3 dB or energy definitions yield widely variable optimum bandwidths, dependent upon the shape of the pulse and the bandpass of the filter; there can be no quick estimation of optimum bandwidth if these parameters are utilized.

Distributed clutter, rain or chaff, is often a more serious interference with target detection than noise. In passing through an optimum-bandwidth filter and optimum-bandwidth integrator, the echo is stretched in both range and angle; the clutter spectrum, being the product of the transmitted spectrum and the bandpass of the receiver, is narrower than either. As a result, the 6 dB two-way beamwidth and the 6 dB pulse width closely approximate the extent of the radar cell from which the "optimum" receiver accumulates clutter energy.

To summarize the general utility of 6 dB definitions: (1) The range equation for detection in noise need not include peak power, pulse width, or receiver bandwidth; only the efficiency of the integrator requires a definition of the number of pulses being received, and a 6 dB echo definition is universally employed. (2) The optimum bandwidth is close to the inverse of the echo duration if both are 6 dB definitions; this applies to both the receiver filter and the integrator. (3) The energy of the interference from clutter, rain or chaff, that is accepted by an approximately matched receiver is well defined by the 6 dB pulse duration and 6 dB two-way beamwidth.

Approximations to Matched Filters. The most efficient filter for discriminating between white gaussian noise and the desired echoes is a matched filter, a passive network whose frequency response is the complex conjugate of the transmitted spectrum. It can process echoes from all ranges. The correlation mixer, an active device which compares the received signals with a delayed replica of the transmitted signal, is mathematically equivalent to a matched filter, but it is responsive only to echoes from one specific range; consequently its use in radar systems is more limited.

Table 3.2 illustrates the degree of sacrifice in detectability that results in approximating a matched filter, either to simplify the hardware or to achieve better filtering of other forms of interference. The optimum bandwidths of these filters

TABLE 3.2 Approximations to Matched Filters

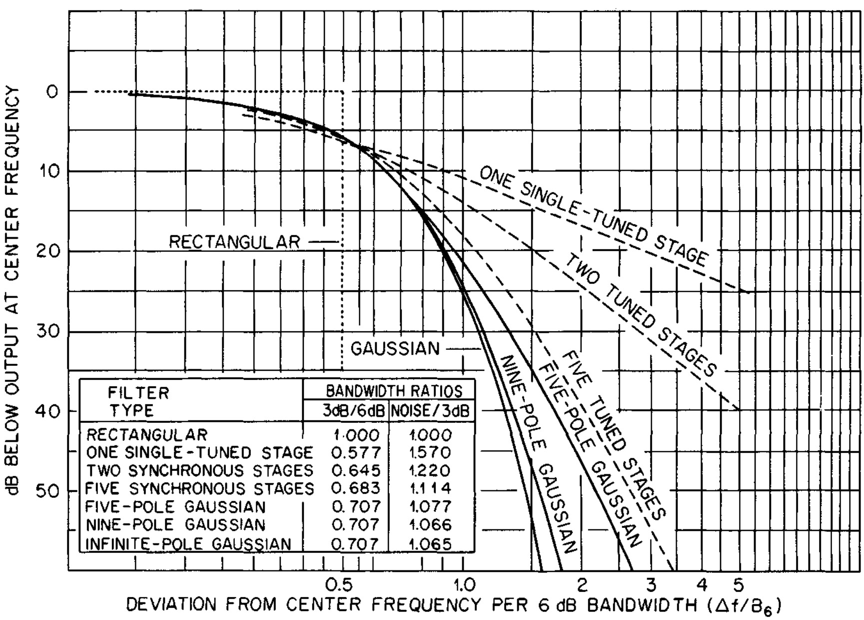
Transmitted pulse shape	Receiver filter	Optimum bandwidth–time product			Mismatch loss, dB
		6 dB	3 dB	Energy	
Gaussian	Gaussian bandpass	0.88	0.44	0.50	0
Gaussian	Rectangular bandpass	1.05	0.74	0.79	0.51
Rectangular	Gaussian bandpass	1.05	0.74	0.7	0.51
Rectangular	5 synchronously tuned stages	0.97	0.67	0.76	0.51
Rectangular	2 synchronously tuned stages	0.95	0.61	0.75	0.56
Rectangular	Single-pole filter	0.70	0.40	0.63	0.88
Rectangular	Rectangular bandpass	1.37	1.37	1.37	0.85
Phase-coded					
Biphase	Gaussian	1.05	0.74	0.79	0.51
Quadriphase	Gaussian	1.01	0.53	0.5	0.09

also are tabulated in terms of product of the filter bandwidth and pulse duration. Typically, the bandwidth may deviate 30 to 50 percent from the optimum value before the detectability is degraded by an additional 0.5 dB. This rather broad "optimum" region is centered near unity bandwidth-time product for virtually all practical filters if one uses the 6 dB definitions.

Sometimes the bandwidth of a radar receiver is in excess of the optimum to allow for some offset between the echo spectrum and the filter bandpass, caused by target velocity and receiver tuning tolerances. Although this makes the radar more susceptible to off-frequency narrowband interference (Fig. 3.8), it reduces the time required to recover from impulse interference (Fig. 3.9). These figures also illustrate that, to provide good suppression of both forms of interference, the shape of the filter bandpass characteristic is even more important than its bandwidth. Rectangular bandpass or impulse responses should be avoided; the closer one approximates a gaussian filter, the better the skirts in both frequency and time domains.

In the case of phase-coded transmissions, the duration of the subpulses is the time parameter. This is equivalent to the spacing of the subpulses with one exception: the quadriphase code[5,6] employs half-cosine subpulses with 6 dB width equal to four-thirds of subpulse spacing. One of the virtues of the quadriphase code is the unusually low mismatch loss, owing to the fact that the impulse response of the gaussian filter is an excellent approximation to the subpulse shape. A digital correlator in a later stage of the receiver completes the matched filter.

Filtering Problems Associated with Mixer Spurious Responses. The approximation of a matched filter is generally most easily accomplished at some



FILTER TYPE	BANDWIDTH RATIOS 3dB/6dB	NOISE/3dB
RECTANGULAR	1.000	1.000
ONE SINGLE-TUNED STAGE	0.577	1.570
TWO SYNCHRONOUS STAGES	0.645	1.220
FIVE SYNCHRONOUS STAGES	0.683	1.114
FIVE-POLE GAUSSIAN	0.707	1.077
NINE-POLE GAUSSIAN	0.707	1.066
INFINITE-POLE GAUSSIAN	0.707	1.065

FIG. 3.8 Bandpass characteristics of filters.

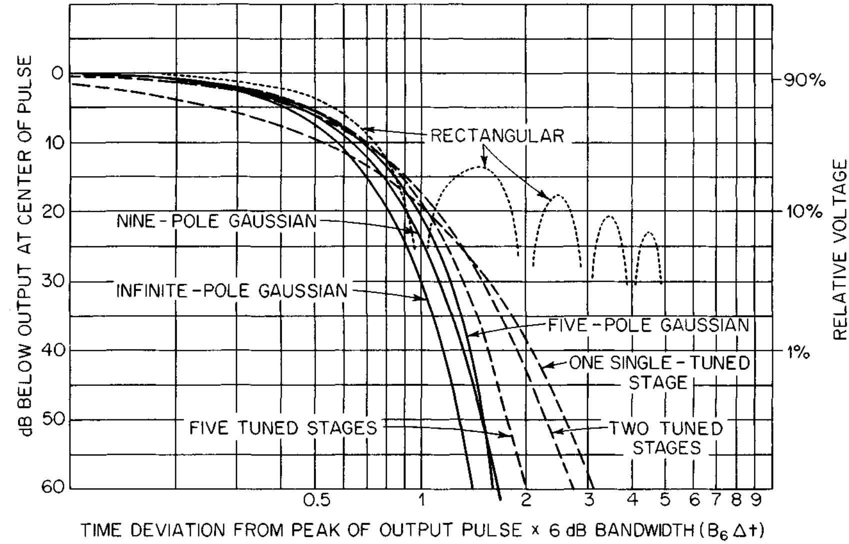


FIG. 3.9 Impulse characteristics of filters.

frequency other than that radiated by the radar. The optimum filtering frequency is a function of the bandwidth of the echo and the characteristics of the filter components. Consequently, it is necessary for the radar receiver to translate the frequency of the echo to that of the filter, in one or more steps, using local oscillators and mixers.

Section 3.4 described how spurious responses are generated in the mixing process. Unwanted interference signals can be translated to the desired intermediate frequency even though they are well separated from the echo frequency at the input to the mixer. The ability of the radar to suppress such unwanted interference is dependent upon the filtering preceding the mixer as well as on the quality of the mixer itself.

The image frequency is the most serious of the spurious-response bands, but an image-rejection mixer can readily suppress these signals by 20 dB. A filter can further attenuate image-frequency signals before they reach the mixer unless the ratio of input to output frequencies of the mixer exceeds the loaded Q of the available filters. This image-suppression problem is the reason why some receivers do not translate from the echo frequency directly to the final intermediate frequency in a single step.

The other spurious products of a mixer generally become more serious if the ratio of input to output frequencies of the downconverter is less than 10. The spurious-effects chart (Fig. 3.2) shows that there are certain choices of frequency ratio that provide spurious-free frequency bands, approximately 10 percent of the intermediate frequency in width. By the use of a high first IF, one can eliminate the image problem and provide a wide tuning band free of spurious effects. Filtering prior to the mixer remains important, however, because the neighboring spurious responses are of relatively low order and may produce strong outputs from the mixer.

In addition to external sources of interference, the radar designer must be concerned with internal signal sources. MTI and pulse doppler radars are particularly susceptible to any such internal oscillators that are not coherent, i.e., that do not have the same phase for each pulse transmission. The effect of the spurious signal then is different for each echo, and the ability to reject clutter is degraded.

A truly coherent radar generates all frequencies, including its interpulse periods, from a single stable oscillator. Not only all the desired frequencies but also all the internally generated spurious signals are coherent, and they do not affect the clutter rejection.

More commonly, MTI and pulse doppler radars are pseudo-coherent, as illustrated in Fig. 3.10. The coho is the master reference for the phase detector and may be the clock from which the interpulse periods are determined. The coho also is employed in generating the transmitter frequency, offset from a noncoherent local oscillator. Neither the local oscillator nor the transmitter is co-

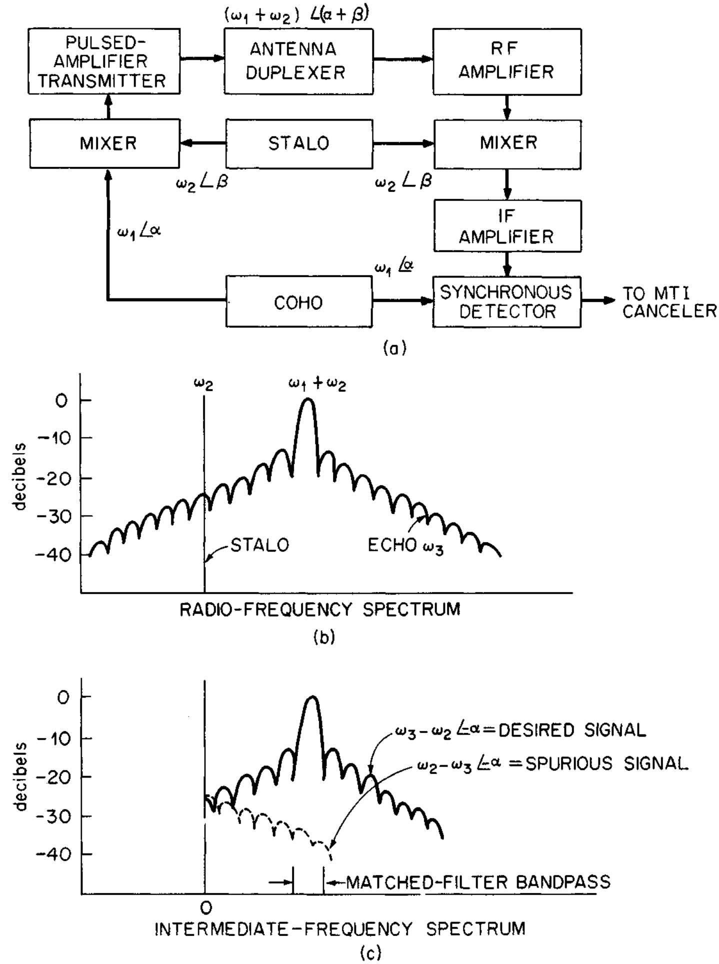


FIG. 3.10 Pseudo-coherent radar. (*a*) Block diagram. (*b*) RF spectrum of the echo. (*c*) IF spectrum of the echo.

herent; their phases are different for each pulse transmitted. Only in the IF portion of the receiver is the desired echo from a stationary target coherent (α = constant, where α is the phase of the echo when the transmitter is turned on); virtually all the spurious outputs of the mixer are noncoherent and produce a fluctuating signal at IF.

The only spurious component that may be coherent in the pseudo-coherent radar is the image frequency. As illustrated in Fig. 3.10, a pulsed spectrum can have fairly wide skirts and can overlap into the image band. The folded spectrum at intermediate frequency has two components within the bandpass of the matched filter, and if the percentage bandwidth at IF is large, the undesired image component can be significant. Only when it is coherent is a degradation of clutter cancellation impossible.

Consider a pseudo-coherent radar in which the interpulse periods are generated from an independent source. Now the phase of the coho (α) changes from pulse to pulse, and although the phase of the desired echo from a stationary target changes equally, the phase of the spurious image echo changes oppositely. The cancellation of the stationary-target echo is limited to the level of the spurious echo.

It is clear that the overall filtering capability of the radar, its ability to enhance the desired echoes and suppress undesired interference, may be degraded by the spurious responses of the various mixing stages. Particularly susceptible to degradation are MTI and pulse doppler radars, which may not provide the expected improvement in the rejection of clutter if the coho is not at the same phase condition each time that the transmitter pulses. All radars are susceptible to off-frequency interference, which, unless it is filtered before reaching the mixer, may create a detectable output in the desired IF band.

The ability of MTI or pulse doppler processing to suppress clutter may be degraded if the receiver filter is not perfectly stable. The receiver's transfer characteristic (gain, time delay, and bandpass or impulse response) must be constant so that its effect on each echo pulse is identical.

The mixer spurious responses just discussed and the stalo and coho problems of Sec. 3.5 represent only the most likely sources of instability to be encountered. Other elements of the receiver require attention to avoid instability problems. Vibration or power supply ripple can cause gain and phase modulation, particularly in RF amplifiers. Such modulation will degrade clutter attenuation unless the ripple frequency is a harmonic of the PRF.

3.8 LOGARITHMIC DEVICES

Characteristics

Accuracy. Logarithmic devices and IF amplifiers are devices whose output is proportional to the logarithm of the envelope of the IF input. They often approximate the logarithmic characteristic by multiplicity of linear segments. Normally linear segments of equal *length ratio* and varying slope are joined to give a best fit to a logarithmic curve. Each segment will be correct at two points and will have a maximum error at the ends and center. The magnitude of the error[7] increases with the length ratio of the segment. Figure 3.11 shows how this error changes with segment length ratio (also called gain per segment). In practice, the joints between the "linear" segments are not abrupt, and the best fit to the logarithmic curve may have less than the theoretical error.

Logarithmic detectors and amplifiers are frequently designed with adjustments in each stage. This allows for adjustment of the slope and/or length of the segment for a better fit. A precision exponentially decaying IF waveform from a test set is applied to the unit under test. The unit is adjusted for a linearly decaying output, which indicates the correct adjustment.

Dynamic Range. The dynamic range of a logarithmic detector or amplifier is dependent on the number of linear segments N and on the length ratio G of the segments:

$$\text{Dynamic range} = 20N \log G \tag{3.5}$$

A well-designed logarithmic detector may have a dynamic range of 80 dB derived from nine stages with an error as low as ±0.2 dB.

Bandwidth. The bandwidth of a logarithmic detector or amplifier generally varies with signal level. For this reason the logarithmic device is usually designed with excess bandwidth and is preceded by filters which establish the receiver bandwidth. However, the large signal bandwidth of the logarithmic device itself may be measured by using the method indicated in Fig. 3.12. The input voltage is increased from V_i to V_i + 3 or 6 dB, changing the operating point from A to B. The frequency is now changed in both directions to find those two frequencies that place the operating point at C. The same result also may be obtained by changing the frequency until the output is reduced by $3S$ or $6S$, where S is the volts-per-decibel characteristic. This places the operating point at C, which lies on the same curve as C. Because of the dependence of bandwidth on signal level, if possible the logarithmic amplifier should be aligned with signals below the threshold and its pulse response used as a criterion of performance.

The pulse response is measured with a pulsed IF signal having much faster rise and fall times than those of the logarithmic device being tested. The rise time is the time required for the output to rise from the $-20S$ to the $-S$ point, and the fall time is the time required for the output to fall from the $-S$ point to the $-20S$ point. Because of the logarithmic characteristic, the fall time tends to be a straight line and to exceed the rise time.

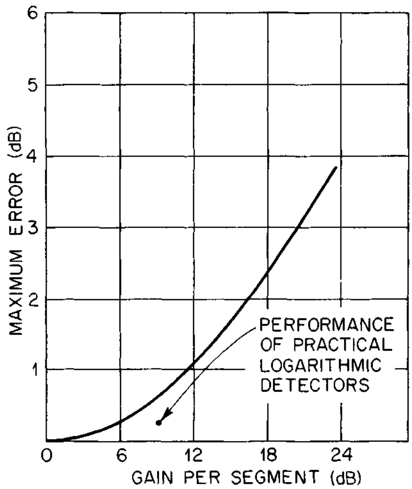
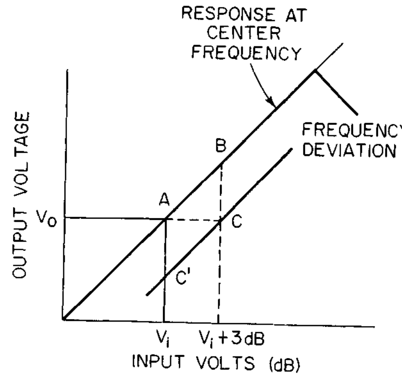


FIG. 3.11 Error in the approximation of a logarithmic curve by linear approximation.



3.12 Method of measuring logarithmic-amplifier bandwidth.

Analog Logarithmic Devices

Logarithmic Detector. A well-known form of the logarithmic detector uses successive detection,[6] wherein the detected outputs of N similar limiter stages are summed as shown by Fig. 3.13. If each stage has a small signal gain G and a limited output level E, the intersections of the approximating segments fall on a curve described by

$$E_0(M) = n\left[E\frac{\log E_i(M)G^{N+1}}{\log G} + E\left(\frac{1}{G^{M-1}} + \cdots + \frac{1}{G^2} + \frac{1}{G}\right)\right] \tag{3.6}$$

where n is the detector efficiency and $E_i(M)$ represents the particular input levels that correspond to the intersections of the line segments,

$$E_i(M) = \frac{E}{G^M} \tag{3.7}$$

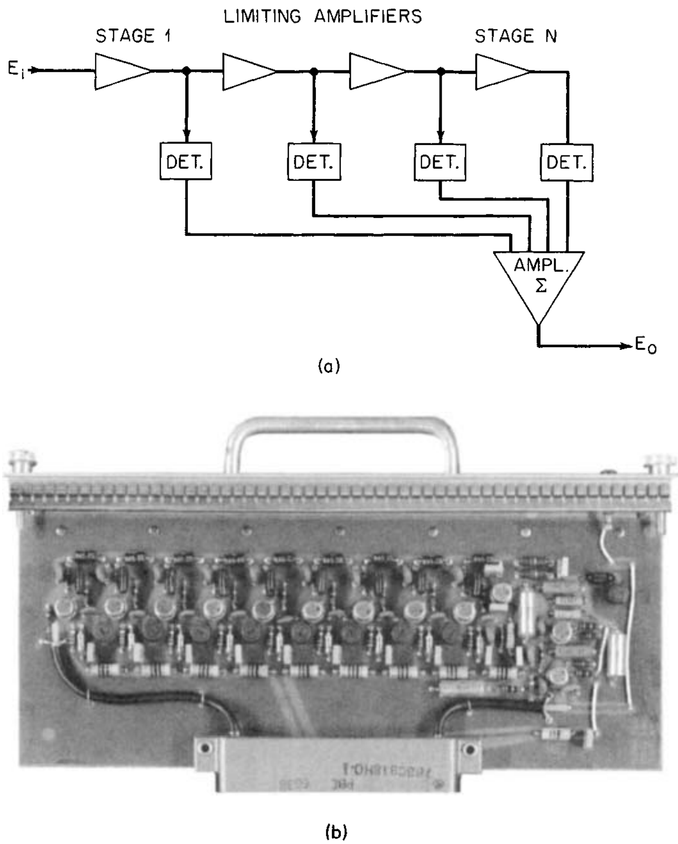


FIG. 3.13 (*a*) Logarithmic detector. (*b*) Nine-stage logarithmic detector. Gain adjustment for each stage is shown below the transistor.

The independent variable M is the order of the stage at incipient saturation and will take on only integer values from 1 to N.

A nonlogarithmic term in the form of a power series in $1/G$ has only a minor effect on logarithmic accuracy. The successive difference in this term as M is varied from 2 to N is E/G^{M-1}. The overall tendency of this term, therefore, is to produce an offset in the output with only a minor loss in logarithmic accuracy at the highest signal levels.

A typical logarithmic detector may have an accuracy of ±4 dB, a dynamic range of 80 dB, and a bandwidth of 5 to 10 MHz at 30 MHz, derived from nine stages (Fig. 3.13*b*). However, Rubin[9] describes a four-stage successive detection design having a bandwidth of 640 MHz centered at 800 MHz. Pulse response time of 2.5 ns is claimed.

Logarithmic Amplifier. A logarithmic IF amplifier may be implemented with N identical cascaded dual-gain stages. In this case a precise logarithmic response results if each amplifier has a threshold E_T below which the gain is a fixed value G and above which the incremental gain is unity. The intersections of the approximating segments fall on a curve described by

$$E_0(M) = E_T\, G \frac{\log E_i(M)G^N}{\log G} \tag{3.8}$$

where $E_i(M)$ represents the input levels corresponding to the intersection of the line segments:

$$E_i(M) = E_T\, G^{1-M} \tag{3.9}$$

The independent variable M is the order of the stage that is at incipient saturation; M will have integer values between 1 and N. In the case of the dual-gain logarithmic amplifier, all intersections of the line segments fall on a logarithmic curve.

A typical logarithmic amplifier may have a dynamic range of 80 dB, derived from nine stages, with an overall bandwidth of 5 MHz or more. Typical accuracy is ±¼ dB over 70 dB dynamic range and ±1 dB over the full dynamic range.

Details of a typical stage are shown in Fig. 3.14, and the voltage characteristics in Fig. 3.15. The stage consists of an attenuator bridged by a series-diode limiter and followed by an amplifier. In the absence of an input voltage, the current divides equally between the limiter diodes. The thresholds are reached when diode $D2$ carries either all or none of the current. If the voltage drops across the diodes are neglected, this occurs when

$$E_i = E_T = \pm \frac{V}{2}\left(\frac{R}{R + R_s}\right) \tag{3.10}$$

Signals above the threshold in magnitude are attenuated by a factor $R/(R + R_f)$, the reciprocal of the amplifier gain. The incremental gain in this region is therefore unity. The blocking capacitors are used because of the dc offset voltage that is characteristic of this form of limiter.

Digital Logarithm. The trend toward digital processing requires mention of a piecewise linear digital approximation which may be accomplished after analog-to-digital conversion and digital doppler filtering to suppress clutter

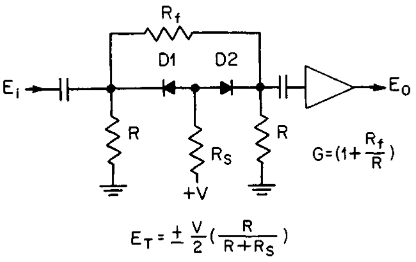


FIG. 3.14 Dual-gain IF stage.

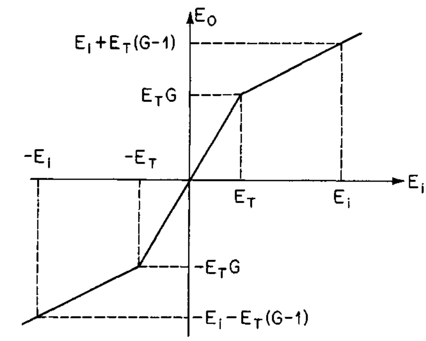


FIG. 3.15 Dual-gain-stage voltage characteristic.

interference prior to any nonlinear operation. A digital word in the power-of-2 binary format may be written

$$E = 2^{N-M}\left(b_{N+1-M} + \frac{b_{N-M}}{2} + \frac{b_{N-1-M}}{4} + \cdots + \frac{b_0}{2^{N-1}}\right) \tag{3.11}$$

where M is the place beyond which all coefficients to the left are zero. Note that M has essentially the same significance as in previous sections. The logarithm of E to the base 2 is

$$\log_2 E = N - M + \log_2\left(1 + \frac{b_{N-M}}{2} + \frac{b_{N-1-M}}{4} + \cdots + \frac{b_0}{2^{N-1}}\right) \tag{3.12}$$

$$\log_2 E \approx N - M + \left(\frac{b_{N-M}}{2} + \frac{b_{N-1-M}}{4} + \cdots + \frac{b_0}{2^{N-1}}\right) \tag{3.13}$$

The whole number $(N - M)$ becomes the characteristic, and the series, a fractional number, is a line approximation of the mantissa. The approximation is accurate to ± 0.25 dB if the mantissa contains at least 4 bits. The accuracy may be improved to any desired degree by use of a programmable read-only memory (PROM) to convert the linear approximation to the true mantissa.

Digital Log Power Combiner. In log format fewer bits of data need to be stored or manipulated, and many arithmetic computations are simplified. For example, the rms combination of the voltages (I and Q) is complicated in linear format, and approximations are generally employed which introduce some error. In log format the process is simple and more accurate:

$$\log_2 I^2 = 2 \log_2 |I| \qquad \log_2 Q^2 = 2 \log |Q| \tag{3.14}$$

$$\log_2 (I^2 + Q^2) = \log_2 I^2 + \log_2 (1 + Q^2/I^2) \tag{3.15}$$

The latter term of Eq. (3.15) is the output of a PROM, using $\log_2 I^2 - \log_2 Q^2$ as the address.

The size of the PROM can be substantially reduced by comparing the two variables to determine the larger (L) and smaller (S):

$$\log_2 (I^2 + Q^2) = \log_2 L + \log_2 (1 + S^2/L^2) \tag{3.16}$$

The latter term of Eq. (3.16) requires fewer bits than that of Eq. (3.15) because its maximum value is 3 dB. The PROM address also is unipolar and may be limited at a ratio where the PROM output drops to zero.

The log power combiner makes power (square-law) integration feasible. A number of variables may be weighted by addition of logarithmic scaling factors and successively accumulated, using the log power combiner to combine each with the prior partial power summation. Square-law integration of multiple echoes from a target provides better sensitivity than prior methods, but this was impossible with analog signal processing and costly when using conventional digital processing.

3.9 IF LIMITERS

Applications. When signals are received that saturate some stage of the radar receiver which is not expressly designed to cope with such a situation, the distortions of operating conditions can persist for some time after the signal disappears. Video stages are most vulnerable and take longer to recover than IF stages; so it is customary to include a limiter in the last IF stage, designed to quickly regain normal operating conditions immediately following the disappearance of a limiting signal. The limiter may be set either to prevent saturation of any subsequent stage or to allow saturation of the A/D converter, a device which is usually designed to cope with modest overload conditions.

The IF phase detector described in Sec. 3.10 requires a limiter to create an output dependent on phase and independent of amplitude. It is employed in phase-lock servos and phase-monopulse receivers.

An IF limiter is sometimes employed prior to doppler filtering to control the false-alarm rate when the clutter echo is stronger than the filter can suppress below noise level. This was widely used in early two-pulse MTI, but it has drastic impact on the performance of the more complex doppler filters of modern radars. It is only compatible with phase-discrimination constant false-alarm rate (CFAR; Sec. 3.13), but it serves a useful purpose in radars utilizing this CFAR process after doppler filtering.

Characteristics. The limiter is a circuit or combination of like circuits whose output is constant over a wide range of input signal amplitudes. The output waveform from a bandpass limiter is sinusoidal, whereas the output waveform from a broadband limiter approaches a square wave.

There are three basic characteristics of limiters whose relative importance depends upon the application. They are performance in the presence of noise, amplitude uniformity, and phase uniformity. When the input signal varies over a sufficiently wide range, all these characteristics become significant. Amplitude uniformity and phase uniformity are dependent largely on the design of the limiter and are a direct measure of its quality.

Noise. Limiter performance in the presence of noise is characterized by a failure to limit signals buried in noise and by an output signal-to-noise ratio that