CHAPTER 11

Radar Receivers

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11.1 INTRODUCTION

The receiver is an integral part of the radar system. As shown in Figure 11-1, it provides the necessary downconversion of the receive signal from the antenna and the inputs required to the signal and data processors.

In a typical pulsed radar system (see Figure 11-2), the transmit signal is coupled to the antenna via some form of a duplexer (transmit/receive [T/R] switch). The function of the duplexer is to couple the transmit signal into the antenna with low loss while providing high isolation during transmit time to prevent saturation of the receiver by the high-power transmit signal. Any radar signals present at the receiver during the transmit time are rejected by the duplexer. Following the completion of the transmit pulse, the received signals from the antenna are coupled into the receiver with low loss.

Modern receivers often have a low-noise amplifier (LNA) at the receiver input followed by a band-pass filter to reduce the noise figure of the receiver. The band-pass filter eliminates out-of-band frequency components from the mixer input. The mixer then downconverts the received signal to an intermediate frequency (IF) signal using local oscillator (LO) signals. The mixer output is subsequently amplified and narrowband filtered to reject unwanted intermodulation components. After this, detection of the intermediate frequency signal provides the output or outputs of the receiver subsystem.

The final receiver outputs are sometimes called the radar video signal. In radar, a video signal is the received signal after all carrier or intermediate frequencies are removed

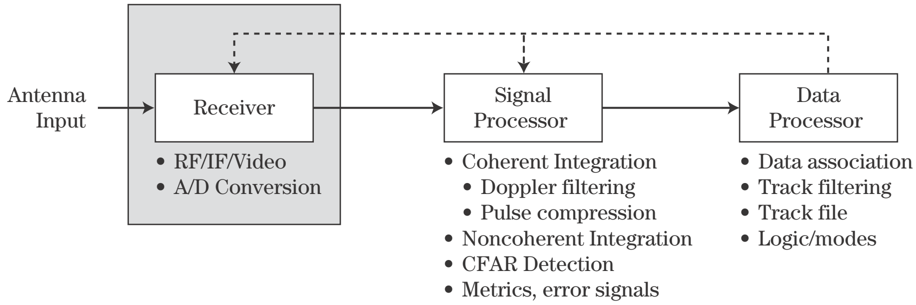


FIGURE 11-1 ■ General receiver functions.

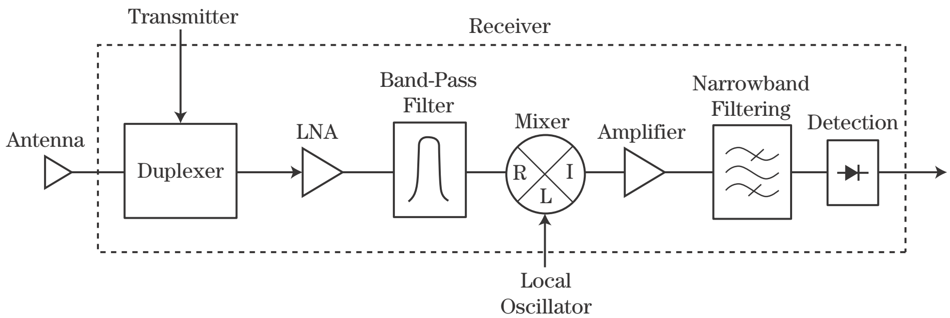


FIGURE 11-2 ■ Major receiver elements.

by demodulation to baseband. Historically, this voltage would be used to drive the radar operator's video display. In modern systems, the voltage is more likely to be digitized and subjected to digital signal processing.

The advantages and disadvantages of component placement and configurations will be addressed in this chapter. Continuing advances in analog-to-digital converters (ADCs) and digital signal processing (DSP) technology are driving receiver development. As converters improve in speed and resolution, the digitization moves closer to the antenna. Improvements in DSP resolution, speed, and cost are pushing traditional analog receiver functions into the digital domain. While these will be detailed in Chapter 14, some of the main digitization components affecting receivers will be discussed in this chapter.

Modern radar receivers are often required to perform a variety of tasks including change of frequency, bandwidth, and gain functions to support the radar modes. These more complex receivers often include digital control networks to select the appropriate receiver depending on the particular radar mode. In addition, these complex receivers often include built-in-test (BIT) functions to enable automated detection of receiver faults.

11.2 SUMMARY OF RECEIVER TYPES

There are several basic types of receiver configurations, including crystal video, superregenerative, homodyne, and superheterodyne. Most modern radars primarily use the latter, but there are other applications for which simpler receiver architectures are better suited. There has been increasing interest in digital receiver configurations for radar applications. Other types of receivers—those used primarily for electronic warfare (EW) receiver applications rather than for radar—include instantaneous frequency measurement (IFM) receivers and channelized receivers.

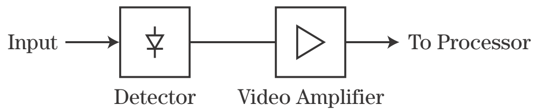


FIGURE 11-3 ■ Crystal video receiver.

11.2.1 Crystal Video Receiver

The crystal video receiver, shown in the block diagram of Figure 11-3, is inherently the simplest type of receiver configuration. The input signal to the receiver is coupled to a detector to convert the radiofrequency (RF) signal directly to video. The video output from the detector is then amplified before providing the output signal to the processor. One disadvantage of this type of receiver is that its sensitivity is 30 dB to 40 dB less than that of a typical superheterodyne receiver because the detector also processes the broadband noise at the detector input. The other disadvantage with this type of receiver is that all the amplification, typically 110 dB or more, is performed by the video amplifier. As a result, the received pulse shape is normally distorted. Because of these limitations, the use of this type of receiver is generally limited to short-range systems. Automotive collision avoidance radar might be a possible application for this type of receiver.

An alternate form of the crystal video receiver, sometimes called a tuned radio frequency (TRF) receiver, is shown in Figure 11-4. The RF input is amplified before detection. The sensitivity and associated noise figure of the receiver is improved by the selectivity and the gain of the RF amplifier, thus resulting in a higher signal to noise at the detector output. Another advantage of this configuration is that the video amplifier gain required is reduced because of the gain of the RF amplifier. Disadvantages of this type of receiver include the added cost of the RF amplifier and the considerably reduced sensitivity compared to that achievable with a superheterodyne receiver.

11.2.2 Superregenerative Receivers

Superregenerative receivers use a principle of positive feedback to cause them to oscillate periodically at the desired RF [1]. The self-quenching oscillator consists of a single tube or transistor circuit and can be used both as a transmit source as well as a receiver. The main feature of this type of receiver is the extremely high gain achieved by the single stage, thus making it attractive for applications requiring simple implementation. The drawbacks of this type of receiver include poor sensitivity compared with a superheterodyne receiver as well as inferior selectivity and gain stability, lack of frequency stability since it is not phase locked, and inherent reradiation. One of the main problems with this type of receiver is that the output has high noise content due to the regeneration noise. The noise is predominantly at the superregeneration oscillation frequency. Since the regeneration frequency is higher than the received bandwidth, the resulting bandwidth is wider due to sum and difference mixing between the input frequency and the superregeneration frequency. Some beacon implementations use this architecture, and it can also be employed for noncritical low-cost applications.

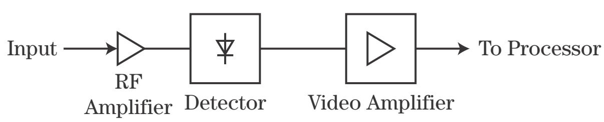


FIGURE 11-4 ■ Modified crystal video receiver.

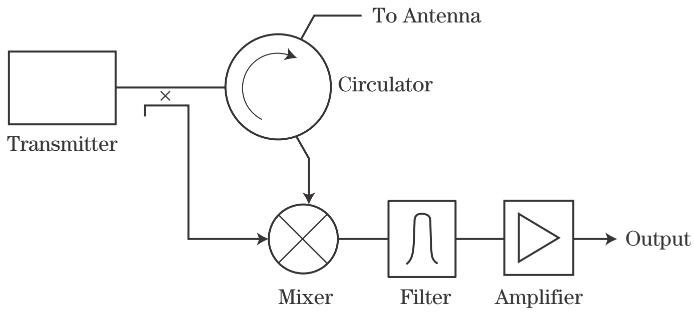


FIGURE 11-5 ■ Homodyne receiver.

11.2.3 Homodyne Receivers

An example of a homodyne receiver is shown in the block diagram of Figure 11-5. For this type of receiver, a portion of the transmit signal is coupled from the transmitter and is used as the local oscillator input to the receiver mixer. A circulator, which is a ferromagnetic device, couples the transmit signal to the antenna while isolating the transmit signal from the receiver. The received signal at the antenna is coupled to the receiver while providing isolation to the transmit port of the circulator. It thus functions as a duplexer for this radar. For this type of receiver to work, the transmitter must still be transmitting when the signals from the target are received. This type of receiver is rather simple to implement in that no local oscillator is required yet is still capable of providing coherent receive signals to the radar processor. Examples of the types of radars that can use a homodyne receiver are low-cost continuous wave (CW) radars (e.g., police speed radars) and radars using frequency modulated CW (FMCW) waveforms. Because of the long waveforms, this type of radar often uses relatively low-power solid-state transmitters, and the sensitivity is considerably improved compared with that of a crystal video receiver.

11.2.4 Superheterodyne Receivers

The superheterodyne receiver, shown in the block diagram of Figure 11-6, was originally developed as a radio receiver and used a tunable local oscillator signal to mix the signal down to a common intermediate frequency. A heterodyne receiver is essentially the same as a superheterodyne receiver, except that the LO frequency is fixed. Most current radar systems have the capability of changing the RF frequency to enable frequency diversity or specialized processing waveforms, and thus the LO frequency is then tuned to follow the RF frequency. These same features are applicable to radar receivers, except that in most cases the local oscillator is at a fixed frequency, offset from the transmit frequency

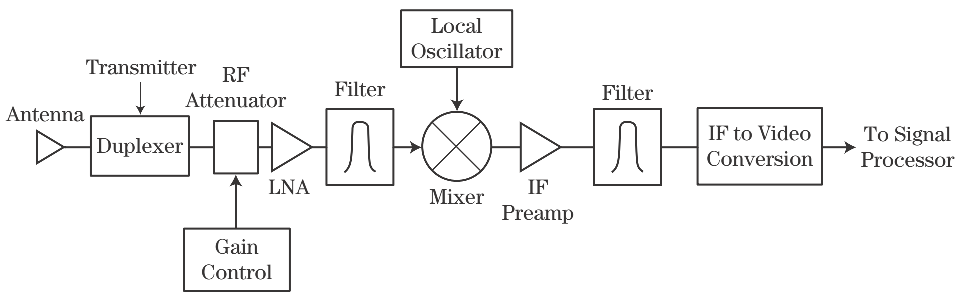


FIGURE 11-6 ■ Generic superheterodyne receiver configuration.

by a fixed IF. The receive signal from the antenna is coupled through the duplexer to the mixer. The receive signal may be fed through an attenuator to provide increased dynamic range and to protect the receiver from saturation from close-in targets. This is called Sensitivity Time Control (STC). In many current applications the receive RF signals are amplified by an LNA prior to the mixer. Recent developments of LNAs with low-noise figures enable improved sensitivity and, with the help of a band-pass filter following the LNA, can overcome the effects of the double sideband noise figure of mixers. For radar receivers requiring extreme sensitivity, cryogenic receiver front ends are used in place of the LNA. The mixer downconverts the input RF signal, using the local oscillator signal, to provide the received IF signal. The mixer output signal is commonly amplified, and in many cases the amplifier is included as part of the mixer assembly. Band-pass filtering is then used to limit the IF signal to that of the desired receive signal while providing rejection of unwanted mixer products and out-of-band signals. Following preamplification and filtering, the received signals are generally downconverted from IF to video and are detected. Some applications, such as dual polarized or monopulse radars, use multiple receiver channels, all connected to a common LO frequency.

11.2.5 Digital Receivers

The advances in high-speed A/D converters have led to their increased use in radar receiver applications. Fully digital receivers that directly digitize the RF signal are rarely used for current radar applications, although an example of such a radar receiver is included in the discussion of spurious-free dynamic range [2]. However, most modern radar receivers include A/D converters somewhere in the receiver chain to provide digital signals to the radar processor. Digitization at IF is increasingly being used for modern radar receivers, especially for coherent systems. Digital processing is also being employed to supplant analog functions in the receiver, for example, replacing analog filter functions with finite impulse response (FIR) digital filters. Software control of receiver functions via the use of embedded field programmable gate arrays (FPGAs) allow the receiver the flexibility to adapt to different radar modes of operation.

11.2.6 Instantaneous Frequency Measurement Receivers

IFM receivers can be used to determine the frequency of an incoming radar signal. Pace [3] describes the implementation of an IFM receiver, in which the incoming signal is split into N channels, with each of the N channels providing a direct path and a delayed (in time) path. The frequency of the signal is determined by measuring the phase between the direct and delayed paths. The implementation is wideband and can provide near instantaneous measurement of the receive signal frequency. However, this implementation works only if a single RF frequency is present, and it yields flawed measurements if two or more RF signals are simultaneously present.

11.2.7 Channelized Receivers

A channelized receiver separates the incoming signals into a multitude of superheterodyne receiver channels, channelized in frequency or time [4]. A frequency channelized receiver uses multiple receiver channels, each provided with separate LO frequencies. Note that many radars have multiple receiver channels; however, the term *channelized receivers* implies bands of frequency selective filtering with parallel receiver chains. The band-pass

filters on the RF input signals would be selected to limit the incoming bandwidth of the individual receive channel, but, depending on the number of channels, the overall receiver bandwidth could be extremely large. The advantage of this type of receiver is that it can cover an extremely wide bandwidth while providing good receiver sensitivity. It also has the capability of separating out the returns from different radars. The disadvantage of this implementation is the increased hardware required.

11.3 MAJOR RECEIVER FUNCTIONS

11.3.1 Receiver Protection

The sensitive components of a receiver must be protected from high-power RF signals that can leak in from the radar transmitter or come from high-power interfering signals in the environment. During transmit time, the transmit energy must be isolated from the receiver, since the relatively high transmit power could either burn out the sensitive receive components or cause the receiver to saturate. The recovery time from receiver saturation could prevent the radar from detecting close-in targets to the radar. For a high-power system, the duplexer may be a waveguide device with a radioactive gas-discharge T/R switch activated by the transmit pulse to short out the signal to the receiver and reflect the transmit energy to the antenna. For lower-power radar systems, the duplexer may be a circulator that directs the transmit energy to the antenna port while the signal received by the antenna is coupled into the receiver components. In certain cases, the isolation provided by the circulator itself is insufficient to prevent receiver saturation, so a receiver protector switch may be connected in series with the circulator to short out the input to the receiver during transmit time.

Typical radar duplexers employ one or a combination of the following: pre-T/R tubes; T/R tubes; ferrite limiters; and diode limiters. Pre-T/R tubes and T/R tubes have radioactive material to enable fast operation at the onset of the transmitter pulse, but they also have limited life. Ferrite limiters have moderate power-handling capability, fast turn-off, and inherently long life. Diode limiters provide moderate power-handling capability and fast turn-off. Multipactors have high power-handling capability and fast turn-off. While an all solid-state approach to receiver protection is being sought, it currently has limitations, and catastrophic failure of a solid-state protector could result in damage to the receiver.

At the present time, the most economical approach for high- and moderate-power radar systems is a combination of pre-T/R, T/R switch, and diode limiter [5]. For moderate- or low-power radars, a ferrite circulator can be used as receiver protector. Generally speaking, a four-port circulator is normally used to prevent transmit power from being reflected into the transmitter. This can be done with either two three-port ferrite circulators or a differential phase shift circulator, which is inherently a four-port device. Diode limiters or PIN switches must often be used in addition to circulators in the receiver to provide additional isolation during transmit time.

11.3.2 RF Preselection

RF preselection is required in most receiver designs to minimize or eliminate interfering signals. In a benign background, the most important reason for RF preselection is to reduce the exposure of the receiver to spurious signals. However, most radar returns are competing with the radar signals reflected not only from targets of interest but also from

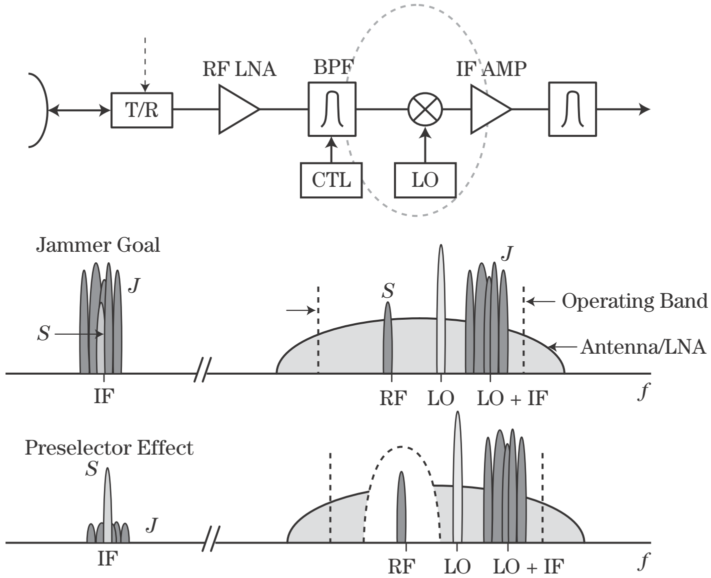


FIGURE 11-7 ■ Effects of preselection on rejection of jammer signals.

other frequency sources. These other frequency sources may be unintentional returns from other radars or communication sources or from jammers deliberately trying to jam the radar to avoid target detections. For barrage noise-type jammers, the jammer signal is spread in frequency so that only a portion of the jammer signal is in the bandwidth of the desired radar signal.

As shown in Figure 11-7, the desired radar signal (S) and the jammer signals (J) both lie within the receiver passband. So when mixed with the LO, both signals end up at the IF of the receiver due to mixer intermodulation products (see Section 11.3.3 for a discussion of mixer intermodulation components); thus, the J signal falls on top of the S signal at IF. By installing a band-pass filter on the RF signal prior to the mixer, the mixer input sees only the input signals that are present in the radar's receiver band. The RF low-noise amplifier prior to the IF improves the sensitivity of the receiver and overcomes the losses in the filter prior to the mixer. By mixing the RF signal with the LO frequency and filtering the mixer IF output to the frequency bandwidth of the transmit pulse, a substantial portion of the jammer returns not within the desired signal bandwidth are rejected by the filter. This has the effect of pulling the target returns not coincident with the desired signal bandwidth out of the jammer background. Not only does the LNA and filter combination help remove broadband jammer signals and other unintentional interference, but it also eliminates the noise and jammer signals at the image frequency from the mixer input. Normally, the LNA is located as closely as possible to the antenna (following the duplexer) so that it helps to considerably reduce additions to the noise figure due to transmission line, filter, switch, and attenuator losses prior to the mixer.

11.3.3 Frequency Downconversion and Mixers

Most radars must downconvert the RF signals to a lower intermediate frequency for detection and processing. Figure 11-8 shows an example of a single downconversion from a 10 GHz RF to a 200 MHz IF. The final IF frequency must be low enough for the input frequency to the logarithmic amplifier for downconversion to video or to enable conversion

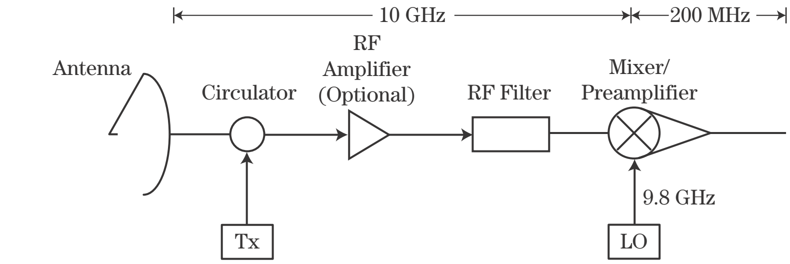


FIGURE 11-8 ■ Single downconversion process.

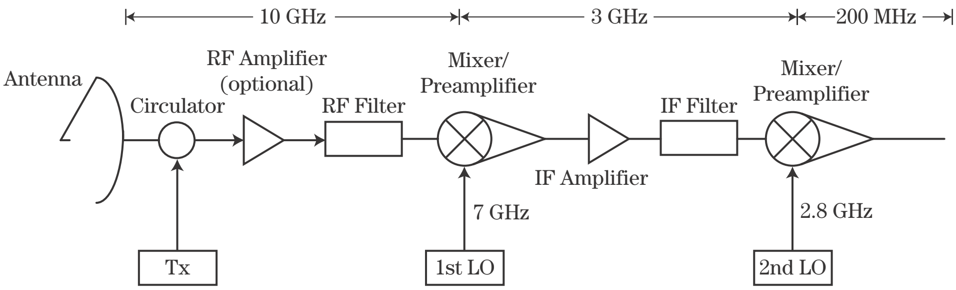


FIGURE 11-9 ■ Double downconversion process.

of the IF signal to digital format. For a radar with a single downconversion to IF, the LO frequency has to be fairly close to that of the RF frequency. Mixer intermodulation components are fairly close to the desired signal, which makes possible introducing spurious frequency components into the IF passband. It is generally difficult to incorporate an RF filter with a narrow enough bandwidth at RF to filter out the unwanted spurious frequencies due to the small offset of the LO from the transmit frequency. In addition, the image frequency component can also be in the receiver band, and, while it can be suppressed by a single-sided mixer, its level may still be in excess of that required.

Most microwave and millimeter wave radar receivers use a double downconversion process. Figure 11-9 shows this process, from a 10 GHz RF to a 3 GHz first IF and then from the 3 GHz first IF down to a 200 MHz second IF. With the first downconversion, the mixer intermodulation components are further separated from the RF, making it easier to filter them out at RF. For both, the mixer is at the heart of the downconversion process.

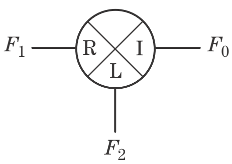


FIGURE 11-10 ■ Mixer model.

The mixer, shown in the model of Figure 11-10, is a nonlinear device used in receivers to convert signals at one frequency to a second frequency. In a receiver the mixer is normally used to mix the RF signal, f_1, with the LO signal, f_2, to a lower IF signal, f_0. The nonlinear mixer products result not only in the desired difference frequency but also in products of the harmonics of the frequencies mixing with each other resulting in undesired mixer outputs. These undesired mixer outputs are spurious signals, commonly referred to as *spurs*. These spurs have the disadvantage that they effectively reduce the spurious-free dynamic range of the receiver.

The outputs from the nonlinear mixing process can be described by the following equation [6]:

$$I_0 = F(V) = a_0 + a_1 V + a_2 V^2 + a_3 V^3 + \cdots + a_n V^n + \cdots \tag{11.1}$$

where I is the device current, and V is the voltage. For a mixer the voltage, V, is a combination of the RF voltage, $V_1 \sin\omega_1 t$, and the LO voltage, $V_2 \sin\omega_2 t$. Thus, V is of

the form

$$V(t) = V_1 \sin \omega_1 t + V_2 \sin \omega_2 t \tag{11.2}$$

where ω_1 is the RF angular frequency, and ω_2 is the LO angular frequency. This results in an infinite Taylor power series:

$$\begin{aligned} I_0 = a_0 &+ a_1(V_1 \sin \omega_1 t + V_2 \sin \omega_2 t) + a_2(V_1 \sin \omega_1 t + V_2 \sin \omega_2 t)^2 \\ &+ a_3(V_1 \sin \omega_1 t + V_2 \sin \omega_2 t)^3 + a_n(V_1 \sin \omega_1 t + V_2 \sin \omega_2 t)^n + \cdots \end{aligned} \tag{11.3}$$

The desired mixing product (normally $f_1 - f_2$) results from the second-order term, and the remainder of the mixing products results in undesired spurious signals. The derivation is intended to reflect the current response of the sum of two voltages across the nonlinearity of the diode $V - F$ characteristic. Most modern radars use bridge rectifiers, where one input is of sufficient level to cause the bridge to act as double-pole, double-throw to the other signal changing the mixing function. This derivation illustrates the nonlinear mixing recognizing that in practice a different function would usually be used. The desired signal can be either $(f_1 + f_2)$ or $(f_1 - f_2)$, but for downconversion in a receiver the difference component $(f_1 - f_2)$ is generally preferred. Intermodulation products are usually considered to be the resulting products from two or more signals at the input that are each converted by the LO rather that higher-ordered products of one signal and the LO [7]. Thus, given two input frequencies f_1 and f_2 into the input of the mixer, with f_L being the local oscillator frequency, the spurs of f_1 would be of the form $mf_1 + nf_2$ where sign(m) = −sign(n) for in- or near-band products.

11.3.4 Selection of LO and IF Frequencies

Selection of LO and IF frequencies are driven by the need to provide the required IF and video bandwidths and to minimize the effect of spurious products due to the downconversion process. Software analysis tools are available on the Internet to enable rapid determination of spurious (intermodulation) products and thus to facilitate selection of the LO and IF frequencies with the lowest spurious residuals. The levels of spurious components depend on the specific mixers and should be available from respective mixer manufacturers. In addition to avoiding or minimizing spurs, careful attention should be given to avoiding the presence of harmonics or LO frequencies in the receiver passband.

Mixer specifications normally list the double sideband noise figure for mixers. For an RF input signal f_{RF1}, the mixer output of interest is the intermediate frequency f_{IF} $(= f_{RF1} - f_{LO})$, assuming that the RF frequency is above the LO. The noise output from the mixer will contain contributions from both the RF frequencies $f_{RF1} = f_{LO} + f_{IF}$ and $f_{RF2} = f_{LO} - f_{IF}$. In addition, unwanted signals at f_{RF2} would also be downconverted to IF unless they were filtered out prior to the mixer. Low-noise amplifiers are currently available with low noise figures and are increasingly being used to amplify the signal prior to the mixer. If the LNA output is filtered to reject the unwanted f_{RF2} noise signal component, the noise out of the mixer is due only to that at f_{RF1}. If the noise figure of the LNA is equivalent to that of the mixer double sideband noise figure and if the gain of the LNA is sufficient, the effective signal-to-noise ratio out of the mixer is increased by up to 3 dB. As a bonus, the unwanted signals at f_{RF2} are also filtered out prior to the mixer, and the filter losses do not degrade the noise figure, assuming sufficient LNA gain. Generally speaking, the LNA is situated close to the antenna to minimize transmission line (waveguide or

coaxial) losses and further improves the receiver noise figure since it cancels the effect of post-LNA losses into the receiver, especially if the receiver is located some distance from the antenna. Modern radar receivers generally incorporate double or triple balance stripline mixer modules to provide superior LO rejection and to decrease the effect of spurs.

11.4 DEMODULATION

Conversion of the signal to video (baseband) and detection are normally required prior to radar signals in the signal processor being used for target acquisition, tracking, and radar displays. This is true for both coherent and noncoherent systems, although some coherent receivers directly sample the IF signal and perform the required detection and processing in the digital signal processor.

11.4.1 Noncoherent Demodulation

Noncoherent detectors provide the conversion of IF signals to baseband to provide the video signal format required for radar displays and subsequent processing. This video signal is the amplitude-only envelope of the IF signal. It is called *noncoherent* because it does not preserve the phase information of the in-phase (I) signal. Noncoherent detection is used for the simplest radar configurations because it does not require complex coherent transmitter and receiver implementations.

The diode detector, such as that shown in Figure 11-11, provides the simplest form of detection. The diode is accompanied by a low-pass filter, normally consisting of a resistor–capacitor combination, to remove the IF signal component. This type of detector is known as an envelope detector or peak detector. Because of the nonlinear nature of diodes, care must be exercised in the selection of the resistor–capacitor combination to enable the video output to effectively follow the envelope of the IF signal. Another form of video detection is the square law detector, shown in Figure 11-12. The signal is split and mixed with itself to form the square law video output, and the low-pass filter is used to remove the IF signal components from the mixer outputs. This type of video detector provides a true square law relationship between the IF signal input and the video output. Figure 11-12 is just one example of a low-pass filter implementation. Selection of the

FIGURE 11-11 ■ Basic diode detector.

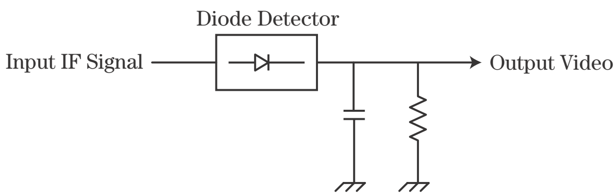
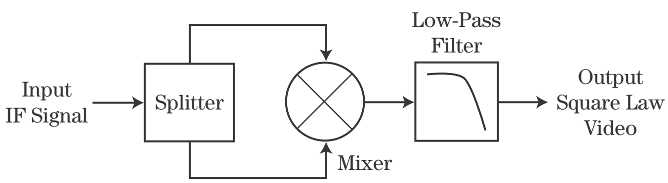


FIGURE 11-12 ■ Square law detector.



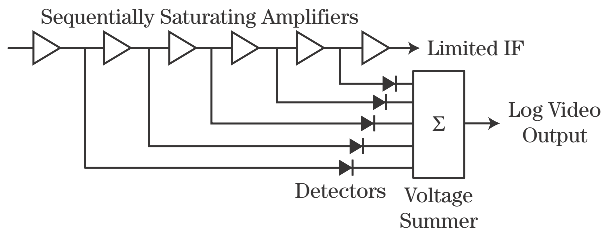


FIGURE 11-13 ■ Log amplifier block diagram.

low-pass filter resistor–capacitor combination is based on the overall circuit impedance, including the diode impedance and the characteristic impedance of the preceding circuit, the highest video frequency required, and the IF. Generally, the IF is much higher than that of the highest video frequency, so the selection is usually not critical. Often, the diode detector is followed by a video amplifier or logarithmic amplifier whose bandwidth effectively acts as a low-pass filter.

Logarithmic amplifiers are commonly used with noncoherent radars to provide amplification with wide dynamic range and subsequent video detection. Log amplifiers, shown in the block diagram of Figure 11-13, normally consist of series of limiting amplifiers, which form the logarithmic signal in the amplifier chain. The detected outputs from the limiters are summed to form the output signal, whose voltage output is proportional to the logarithm of the input IF signal level. Log amplifiers typically maintain the log relationship within approximately ± 1 dB over a 70–80 dB dynamic range, as shown in the example of Figure 11-14. This makes them ideally suited for applications such as plan position indicators (PPIs) radar displays, which require good sensitivity to resolve weak targets while not overloading on stronger targets.

Tangential signal sensitivity (TSS) is one measure of sensitivity of a receiver. The received power observed as TSS is often considered the receiver's minimum detectable signal prior to processing. To measure TSS, a pulse signal is inserted into the receiver while observing the detected signal output. The input signal level is increased until the base of the detected signal-plus-noise waveform is at the peak of the detected noise-only signal. This measure is somewhat subjective, since it depends on the observer's estimate of the peak of a noise-like signal, and may vary several dB depending on the particular

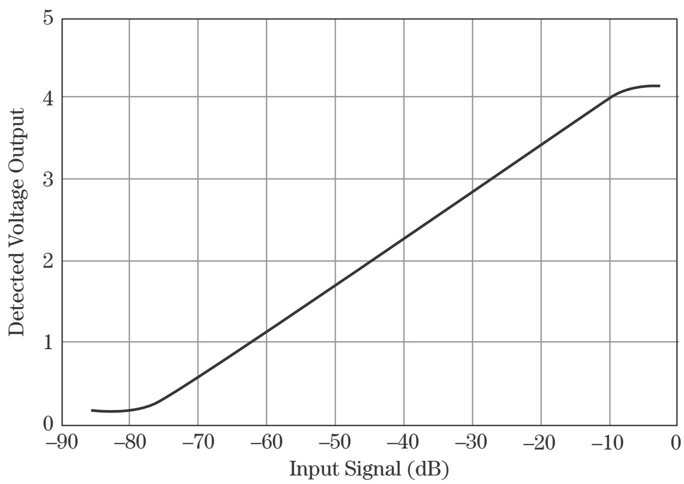


FIGURE 11-14 ■ Typical log amplifier output characteristic.

observer. Generally speaking, this measure of sensitivity is primarily used for noncoherent receiver systems.

11.4.2 Coherent Demodulation

In addition to providing amplitude information on the detected target signal, coherent detectors also provide information on the phase of the received target signal relative to the transmitter phase. The preservation of the phase of the received signal relative to that of the transmitter enables the radar processor to perform tasks such as moving target indication (MTI), Doppler processing (determination of target velocity relative to the radar), synthetic aperture radar (SAR) imaging and space-time adaptive processing (STAP). To enable coherent processing, the phase of the LO signals must be locked to that of the transmit signal. This is generally accomplished by using a highly stable frequency source to determine the frequencies of both the transmitter and the LO frequencies. Coherent-on-receive radars employ a noncoherent transmitter but sample the transmit frequency during the transmit time and use it to lock the phase of the receive signal to that of the transmit signal. In general, the performance of coherent-on-receive radars is somewhat inferior to that of fully coherent radars and is generally not adequate for the needs of modern coherent radar processing

The phase of the receive signal can be obtained in coherent radars by converting the IF signal to in-phase and quadrature (Q) phase video (Figure 11-15). The in-phase and quadrature phase signals can be represented as

$$I = A\cos\theta \tag{11.4}$$

and

$$Q = A\sin\theta \tag{11.5}$$

where A is the amplitude of the signal, and θ is the phase angle between the transmit and receive signal phases.

The amplitude and phase of the signal can be obtained from the I and Q signals, since

$$A = \sqrt{I^2 + Q^2} \tag{11.6}$$

and

$$\theta = \tan^{-1}\left(\frac{Q}{I}\right) \tag{11.7}$$

where the four-quadrant arctangent function is used.

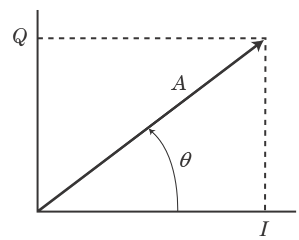


FIGURE 11-15 ■ Analog I and Q detection.

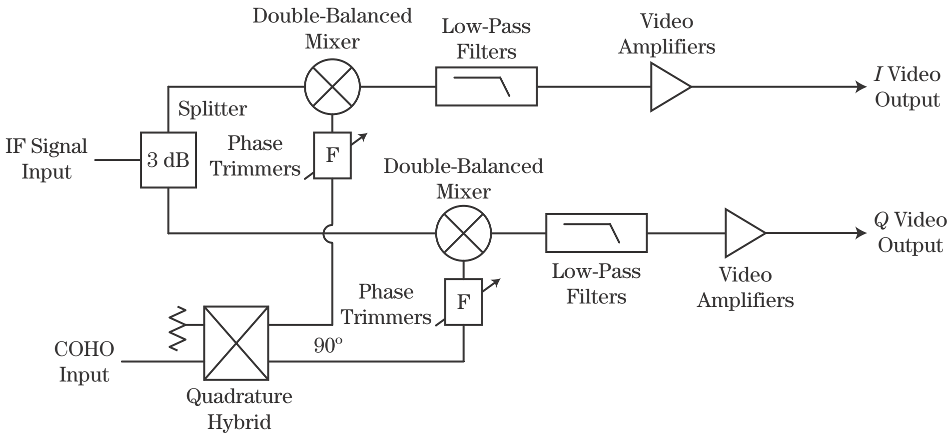


FIGURE 11-16 ■ Circuit for analog coherent I and Q processing.

11.4.3 Analog Coherent Detection Implementation and Mismatch Errors

A typical circuit for analog conversion of the IF signal to coherent I and Q signals is shown in Figure 11-16. The IF signal is split equally and is input to two identical double balanced mixers, while the local oscillator signal is normally provided to a quadrature hybrid that provides equal amplitude signals but with a phase difference of 90° to each other. The quadrature hybrid signals are then coupled to the mixers and the mixer outputs are then in phase quadrature to each other. The mixer outputs are low-pass filtered to pass the video components, which are normally amplified prior to further processing.

In performing the analog I and Q conversion, it is vital that the amplitude balance and the 90° phase difference is maintained by the circuit. Gain and phase mismatches in analog coherent processors have the potential to introduce false Doppler images (ghosts). These ghost images are apparent as false targets, showing up at a Doppler frequency that is the negative of the true target's Doppler frequency. The effect of amplitude and phase mismatches in generating undesired ghost images can be determined from Figure 11-17. In a radar with analog coherent detection, such as that shown in Figure 11-16, careful

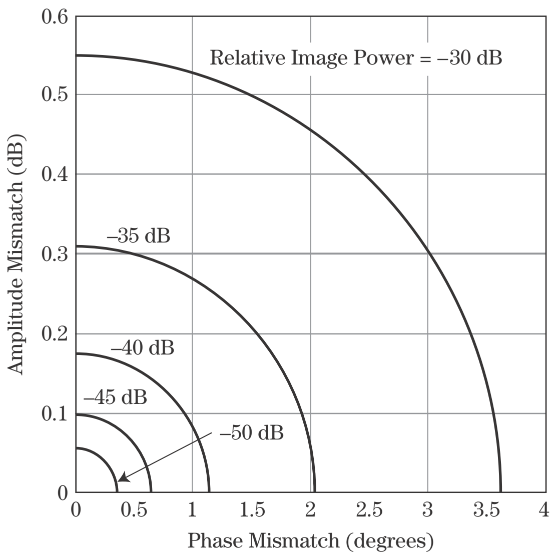


FIGURE 11-17 ■ Effect of gain and phase error mismatches.

calibration of the phase and amplitude trimmers is required to reduce the effect of phase and amplitude errors.

Direct digital coherent detection is increasingly being used to eliminate not only the physical components associated with analog I and Q detection but, more importantly, also the phase and amplitude errors and associated calibration requirements associated with analog coherent detection. A discussion of direct digital coherent detection is given in Section 11.7.2.

11.5 RECEIVER NOISE POWER

Thermal noise is inherent in receivers and limits the ability to amplify the RF spectrum. The inherent noise in receiver components is Gaussian white noise and limits the ability of a radar to detect low-level targets. The ability of a radar to detect targets of a particular radar cross section (RCS, σ) is a function of the transmitter power, the antenna, receiver parameters, and signal processing. The radar equation was developed for a point target in Chapter 2:

$$SNR = \frac{P_t G A_e \sigma}{(4\pi)^2 k T_0 B_n F R^4} \tag{11.8}$$

A modified form of the radar equation substituting for the antenna aperture area, $A_e = G/4\pi$, and including losses, L_s, results in a modified form of the radar equation:

$$SNR = \frac{P_t G^2 \sigma}{(4\pi)^3 k T_0 B_n F L_s R^4} \tag{11.9}$$

where T_0 is the temperature of the receiver (290 K nominally used), B_n is the receiver bandwidth, F is the receiver noise figure, L_s is the system losses, k is Boltzmann's constant (1.38×11^{-23} Joules/K), and R is range (m).

The receiver bandwidth, B_n, is generally taken as the final IF bandwidth of the receiver since this is normally the narrowest bandwidth in the receiver and therefore sets its noise bandwidth. In some cases video filtering is used either instead of or in addition to IF filtering. Video filtering is not as effective as IF filtering in reducing the noise bandwidth due to the double-sided nature of the filtering, so care must be exercised in evaluating its effect on the receiver noise bandwidth.

The system losses, L_s, associated with the radar equation are transmitter, receiver, and signal-processing losses. The losses associated with the receiver, L_r, are losses up to the receiver input and are normally the transmission line losses from the antenna to the receiver.

The *noise figure*, F_n, for the n-th receiver stage is defined as

$$F_n = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{1}{G_n}\frac{N_{out}}{N_{in}} \tag{11.10}$$

where S is the signal power, N is the noise power, and G_n is the gain of the n-th stage. The overall noise figure for a receiver is then

$$F = F_1 + \frac{F_2}{G_1} + \frac{F_3}{G_1 G_2} + \cdots \tag{11.11}$$

The noise figure of the first stage is determined by the input signal level to the receiver and is primarily dominated by the noise of the LNA or mixer (if an LNA is not present). However, attenuation of the signal by components prior to the LNA or mixer reduces the signal into that stage, thus increasing the noise figure F_1. Therefore, the noise figure F_1 is

$$F_1 = L_c + F_{s_1} \tag{11.12}$$

where F_{s_1} is the noise figure of the first stage, excluding components prior to the LNA or mixer, and L_c is the sum of the component attenuations prior to the LNA or mixer.

For a mixer, the noise figure stated in the mixer specifications is normally the double-sided noise figure, assuming both the signal and noise are present in both RF for the upper sideband ($f_{RF1} = f_{LO} + f_{IF}$) and RF for the lower sideband ($f_{RF2} = f_{LO} - f_{IF}$). However, for a superheterodyne receiver, this noise figure must be increased by 3 dB, since the signal is present only in one of these RF sidebands whereas the noise is present in the RF for both sidebands. Also, in most cases, the mixer is immediately followed by a mixer preamplifier (and in many cases in the same package) so that the overall noise figure is normally the mixer–preamplifier combination.

Following the first stage, the contribution of noise from the second stage is reduced due to the gain of the first stage, and the contribution of noise from the third stage is reduced even further by the combination of the gains from the first and second stages. As a result, the first stage of amplification (or mixing) is generally the predominant contribution to the noise figure. With the development of LNAs at microwave (and even millimeter wave) frequencies, most new radars use LNAs to improve the noise figure of the receiver. However, care must be exercised with component placement before or after the LNA. A band-pass filter is often placed before the LNA to reject out-of-band signals, but as depicted in Figure 11-18 the noise out of the amplifier will still contain components at both f_{RF1} and f_{RF2}. However, if the band-pass filter is placed after the LNA, as shown in Figure 11-19, the noise component out of the LNA at f_{RF2} is rejected prior to the mixer input, thus resulting in about a 3 dB lower noise figure for the receiver.

Instead of using noise figure to define receiver noise, an alternate method of characterizing receiver thermal noise is to use the receiver noise temperature. The noise power at the receiver output, P_n, is then expressed as

$$P_n = kT_s B_n \tag{11.13}$$

where T_s is the system noise temperature. The system noise temperature is given as ([8], pp. 26–31)

$$T_S = T_a + T_{tr}(L_r - 1) + L_r T_0(F - 1) \tag{11.14}$$

where T_a is the effective antenna temperature, T_{tr} is the transmission line thermal temperature (K), T_0 is the receiver thermal temperature (K), and L_r is the receiver transmission line loss factor (signal power at the antenna terminal divided by signal power at the receiver input).

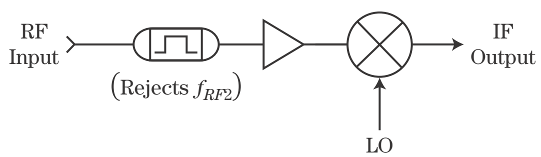
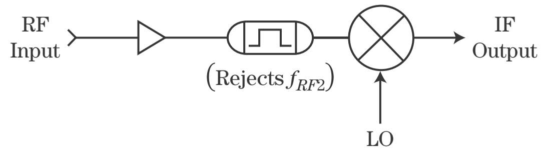


FIGURE 11-18 ■ Filter situated before LNA.

FIGURE 11-19 ■ Filter situated after LNA.



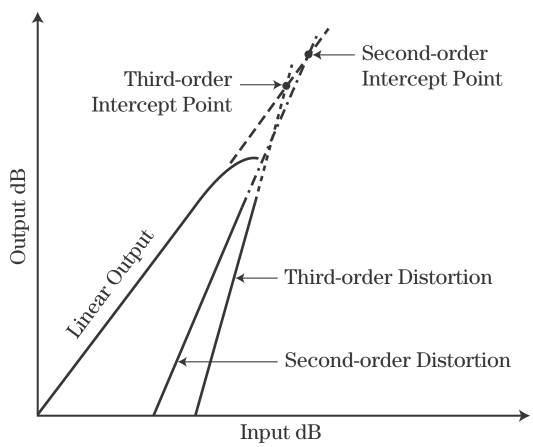
The thermal noise at the receiver input is characterized by a Gaussian probability density function with a white power spectral density. The thermal noise at the receiver output is then characterized by a Gaussian probability density function with a power spectral density of kT_s W/Hz, band limited to B_n Hz.

11.6 RECEIVER DYNAMIC RANGE

Radar returns from targets can vary over a wide dynamic range due to differences in RCS between large and small targets and due to $1/R^4$ decrease in returns as a function of range. The receiver dynamic range is limited by the components in the system, the analog components, as well as the A/D converters. For the analog portions of the receiver, the most notable contributors to limiting the dynamic range include the mixers and the amplifiers. The lower limit on the dynamic range is the noise floor, while the upper end is limited by the saturation of the amplifiers, mixers, or limiters. For linear amplifiers, the upper end of the usable range is generally the 1 dB compression point. That is, in the linear range the amplifier output will increase by a constant dB increment for a given dB increase in the input, thus forming a line of constant slope such as that shown in Figure 11-20. The 1 dB compression point for an amplifier is the point at which the output signal level departs from that of the linear slope by 1 dB.

Unwanted signal components can be generated by nonlinearity in the receiver. Second- and third-order intercept points, as shown in Figure 11-20, are measures of receiver linearity, and these distortions become dominant at the upper end of the receiver curve. The second-order intercept point is due to second-order distortions, while the third-order intercept point is due to third-order distortions. Generally, the third-order distortions are the most dominant, so most amplifiers provide information on the third-order distortions. The distortions that occur due to nonlinearity of the receiver create additional signal

FIGURE 11-20 ■ Receiver distortion versus input power intercept point.



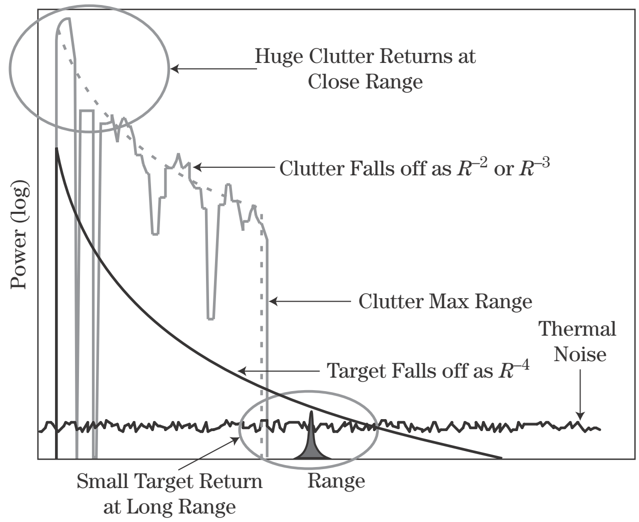


FIGURE 11-21 ■ Return signal amplitudes from targets and clutter.

components for different frequencies, such as $(2f_2 - f_1)$ and $(2f_1 - f_2)$. The intercept points are determined by inserting two different frequency signals of equal input levels that are linearly combined. The distortion products on the receiver output are measured and compared with the original signal inputs. The distortion products are generally measured at relatively low signal levels and are extrapolated to intersect at the extension of the linear gain line.

In designing a radar receiver, each stage of mixing and amplification must be given careful attention regarding degradation of the noise temperature and dynamic range of the receiver. Taylor [9] suggest using a tabular format to enable the designer to keep track of the noise temperature and dynamic range throughout the receiver. Also, the RF design guide by Vizmuller provides valuable insight into the aspects of receiver design [10]. For modern receiver design, software using spreadsheets is currently available to assist in keeping track of the gains and dynamic range throughout the receiver.

Often a radar receiver cannot achieve as large an instantaneous dynamic range needed by the system for all operational situations, and gain control components must be incorporated to achieve system performance goals. Manual gain control, automatic gain control, and sensitivity time control are tools that enable radars to achieve desired overall system dynamic range. The received signals into a radar can have a dynamic range of 120 dB or more. For a constant radar cross section target, the receive signal falls off by R^{-4} as a function of range, so the target returns decrease by 12 dB each time the range to a target increases by a factor of 2. In addition, for surface radars, returns from near-in clutter such as that shown in Figure 11-21 can exceed the target returns at similar range by tens of dB. The returns from surface clutter tend to fall off by only R^{-2} or R^{-3} (see Chapter 5) until the clutter returns become masked due to the radar horizon or other nearer obstructions. As such, the wide range of input signal levels generally exceeds the linear range of most receivers, and even receivers with log amplifiers.

11.6.1 Sensitivity Time Control

One method to increase the dynamic range of receivers is to decrease the sensitivity of the radar for near-range returns. Sensitivity time control (STC) is normally accomplished by

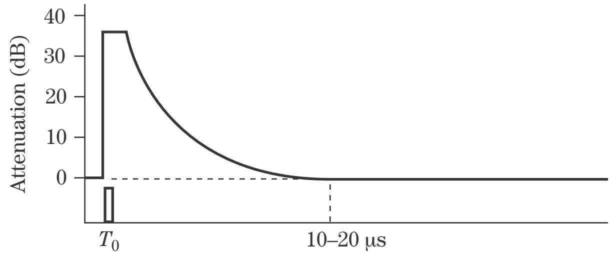


FIGURE 11-22 ■ Sensitivity time control attenuation.

inserting an attenuator between the duplexer and the receiver front end or LNA. Typically around 30 dB, the maximum attenuation (Figure 11-22) is switched in either at or just prior to transmit time. A short time following pulse transmission, the attenuation is decreased as the range increases. Decreasing the attenuation as a factor of R^{-4} would maintain constant amplitude for a given target RCS, whereas decreasing it as a function of R^{-3} would hold it constant for pulse-limited clutter RCS. Reducing the attenuation following an $R^{-3.5}$ rule provides a compromise between both. The STC attenuation present at short ranges helps prevent receiver saturation from strong targets or clutter, and then as the range is increased the attenuation is reduced to enable detection of small RCS targets.

11.6.2 Gain Control

Gain control can be accomplished either with a manual or automated gain control (AGC). Linear amplifiers have limited dynamic range compared with log amplifiers. Gain control is normally required to enable the detection of smaller targets. Typically, gain control in radar is accomplished using gain-controlled IF amplifiers. Some applications control the gain using feedback from the detected video signal.

Manual gain control allows an operator to set the overall gain of the receiver. It is not normally used for adjusting the gain based on target returns since this would require frequent adjustments.

Automated, or automatic, gain control enables the radar to control the gain based on the strength of the target returns. One form of AGC employed when tracking targets samples the detected returns from the radar and continually adjusts the gain to provide an almost constant detected output. This is commonly referred to as *slow AGC* in that the gain adjustment time constant is several radar pulses in duration. This type of gain control is normally used in monopulse radar receivers for tracking single targets. For example, the output from the monopulse sum channel is detected and used to set the gain in both the sum and the angle difference channels. In this manner the amplitude of the angle error signals in the difference channels maintains a constant relationship with the off-boresight angle. Currently, most automated gain control circuits use analog feedback to adjust the gain control voltage into the IF amplifier. Another form of AGC is to have a switchable attenuator on the RF input to prevent receiver saturation and to increase overall receiver dynamic range. This is usually a rather coarse adjustment, for example, to switch in 20 or 30 dB of attenuation on the receiver front end.

Instantaneous AGC can be used, for example, in a monopulse receiver to maintain constant angle indication independent of target amplitude returns. In an instantaneous AGC receiver, the gain is adjusted separately for each target separated in range. With this type of gain control, the detected video signal is used in an analog feedback to the gain control input of the IF amplifier. A delay line is normally incorporated to provide

the receiver time to adjust the IF gain for each individual detected target. Care must be taken in this type of receiver to prevent the receiver from adjusting the gain on the basis of noise.

11.6.3 Coupling Issues

A number of radar receivers use multiple receiver channels. One example is dual polarized radars, in which two receiver channels process the orthogonal polarized signals. Inadvertent cross-coupling of the signal from one polarization channel into the orthogonal polarization channel would contaminate the polarization determination process, especially if the detected polarization ratio was being used for target discrimination purposes. The cross-coupling of the receiver channels must be kept below that of the polarization isolation of the antenna to minimize polarization contamination.

Another example is monopulse radar systems, which commonly use two or three channels for sum (Σ) and difference (Δ) angle sensing (see Chapters 9 and 19). Also, certain instrumentation radars include beacon channels, which could add an additional two or three channels to the receiver. With a monopulse receiver, the Σ signal is maximum and the Δ signal is nulled when the tracked target is on boresight. Any cross-coupling of the Σ signal into the Δ channel could result in a boresight bias and thus an angle tracking error. Isolation of at least 60 dB is generally required for monopulse receivers. Finally, channelized receivers, where the receiver channels may be signals at different RF frequencies, also suffer from cross-coupling effects.

Another form of coupling to be avoided is feedback from alternating current (AC) and direct current (DC) power lines. This is normally prevented by decoupling the receiver circuits using a combination of capacitors and resistors connected to the DC power lines feeding the circuits. The introduction of digital circuits into modern receivers makes them susceptible to transient overvoltage spikes. These transient spikes can occur when a length of cable between the power supply and the circuit acts as an inductor resulting in a voltage spike at the circuit due to instantaneous load changes.

11.7 ANALOG-TO-DIGITAL DATA CONVERSION

Even though the radar signals are analog, most radar systems today use digital signal processing to perform radar detection, tracking, and target display functions. To convert the analog radar signal to a digital representation of that signal, ADCs are used. In the past, the signals were normally downconverted to video prior to analog-to-digital conversion, but many of the newer radar systems directly convert the IF signals to digital, reducing the amount of analog components.

Selection of the ADC sampling speed is determined based on the IF and IF bandwidth as well as on requirements to avoid aliasing of unwanted signal and noise components. Lyons ([11], Chapter 1) shows that, for a sinusoidal input to an ADC, the sampled output data obeys

$$x[n] = \sin(2\pi(f_0 + kf_S)nT_s) \tag{11.15}$$

where f_0 is the baseband signal frequency, f_S is the sampling frequency, n is the sample number, T_s is the sample interval ($1/f_S$), and k is any positive or negative integer.

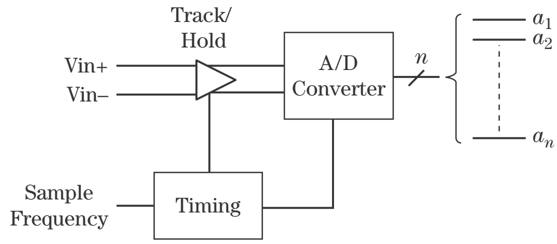


FIGURE 11-23 Typical ADC configuration.

Equation (11.15) indicates that not only the baseband signal frequency f_0 but also frequencies at $f_0 + kf_S$ produce exactly the same sampled output $x[n]$. Thus, the output sampled waveform will be ambiguous unless the input frequencies are low-pass filtered with some cut-off frequency f_{co} and the sampling frequency is greater than $2f_{co}$. More generally, a non-baseband, band-limited IF signal can be sampled uniquely as long as the sampling rate of the IF signal is at least equal to twice the IF bandwidth B_n. Additional discussion of sampling and aliasing is given in Chapter 14.

The direct digital implementation is becoming increasingly popular due to the continuing advances in ADCs with high sampling speeds and increased bit resolutions [3]. Figure 11-23 shows a generic block diagram for a typical high-speed ADC. The IF or video signal, V_{IN}, is normally connected to differential input track/hold circuit (possibly through a buffer amplifier). The track/hold circuit samples the signal and holds the sampled signal constant until the analog-to-digital conversion is performed. The timing circuit is triggered by the sample frequency and samples the V_{IN} signal during the aperture time. The length of the aperture window is limited by the sampling speed and the number of bits in the ADC, and, for high-speed ADCs, the aperture window is generally less than a picosecond. The timing also signals the internal ADC after the aperture window is closed so that it can convert the sampled analog signal to digital format. The output digital signal is then provided to either complementary metal oxide semiconductor (CMOS) or low-voltage differential signaling (LVDS) drivers to provide the b-bit digital signal format.

The relationship between the analog and digital representations of a voltage V_a for an b-bit unipolar ADC is

$$V_a = V_{FS}\left(\sum_{i=1}^{b} a_i 2^{-i}\right) + q_e \tag{11.16}$$

where V_{FS} is the full-scale (saturation) voltage of the ADC, a_i is the value of the i-th bit of the digital representation (0 or 1), and q_e is the quantization error (also known as quantization noise). The digital bits are chosen so that the quantization error is no greater than the ADC quantization error ($\pm$ ½ *LSB* [least significant bit]):

$$\left| V_a - V_{FS}\sum_{i=1}^{b} a_i 2^{-i} \right| = |q_e| < \frac{1}{2}LSB \tag{11.17}$$

where *LSB* is the minimum voltage step size of the ADC and equals $V_{fs}/2^b$.

Digitizing IF signals require ADCs with high sampling speeds, particularly if direct coherent sampling is implemented. In addition, ADCs with a large number of effective bits are desired. Table 11-1 lists the results of a limited survey of currently available ADCs.

TABLE 11-1 ■ Sample of Analog-to-Digital Converters

Part No.	Manufacturer	Bits	Sampling Speed Msamples/sec	SFDR[a] dBc	SNR dB
ADC083000	National Semiconductor	8	3,000	57	45.3
MAX19692	Maxim	12	2,300	68@1.2 GHz	NS
AT84AS004	Atmel	11	2,000	55	51
ADC081500	National Semiconductor	8	1,500	56	47
Model 366	Red Rapids[b]	2/8	1,500	57	47
TS860111G2B	Atmel	11	1,200	63	49[c]
ADC10D1000	National Semiconductor	10	1,000	66	57
MAX5890	Maxim	14	600	84@16 MHz	NS
MAX5888	Maxim	16	500	76@40 MHz	NS
Model 365	Red Rapids[b]	2/14	400	84	70
ADS62P49	Texas Instrument	14	250	85	73
LTC2208	Linear Technology	16	130	83	78
AD9446	Analog Devices	16	110	90	81.6

Note: NS, not specified.
[a]Spurious-free dynamic range.
[b]Dual sampler.
[c]Noise power ratio.

Table 11-1 lists, in addition to sampling speeds, the spurious-free dynamic range (SFDR) and the signal-to-noise ratio (SNR) for the devices when available. Spurs associated with the downconversion process are undesirable in that they could appear as false targets in the receiver, so maintaining a high SFDR is essential to maintaining high sensitivity for low-level targets. While the number of bits is indicative of the resolution of the ADC, the SNR of the device is an important parameter since it indicates the effective number of bits and is generally one to three bits less than that obtained from 20 $\log_{10}(2^n)$ (see Chapter 14). As with all of the ADCs listed in the table, the manufacturer's product specification sheets need to be examined in detail to determine the input frequencies for which the SFDR and SNR are listed.

Random jitter in clocks sampling ADC inputs will induce additional noise on the output of the ADC, resulting in additional noise on the ADC output. For high-resolution ADCs very low clock jitter is required to prevent the effects of clock jitter from reducing the output SNR on the ADC output [12,13]. In many cases, the current trend is to sample the analog signal at the IF, and the presence of jitter can induce phase errors in the sampling process as well as affect the overall SNR [14]. This effect is increasingly evident with very high-speed sampling frequencies.

Dynamic range in a digital receiver (prior to processing gain in the signal processor) is limited by the dynamic range of the ADC. Figure 11-24 show the effective dynamic range for an ADC of 12 bits, although currently ADCs of 14 to 16 bits are readily available. Normally, the noise level into the receiver is set to be about 1 to 2 LSB. This is because the signals from small targets at far ranges are often buried in the noise, and unless gain is set so that the noise gets digitized, the follow-on signal processor will not be able to integrate the signal to improve the SNR of target returns. It is also important to keep the maximum received signal from exceeding the full-scale value of the ADC, and normally about 1 dB of headroom is maintained on clutter and targets to prevent ADC saturation effects.

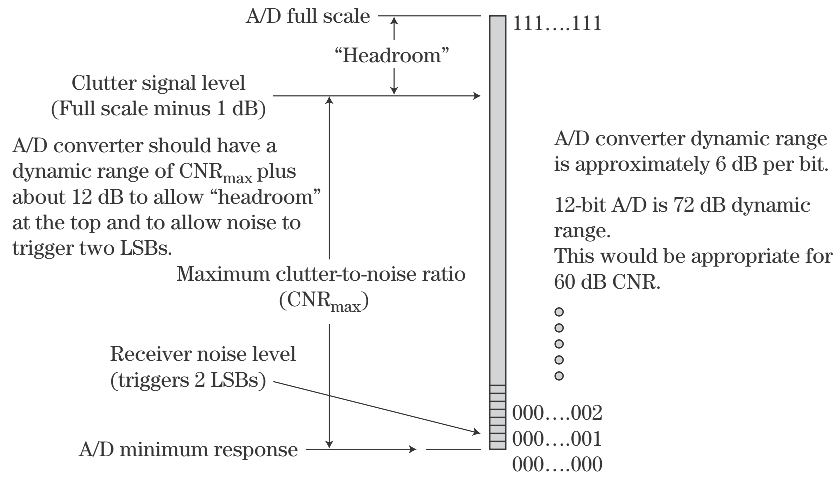


FIGURE 11-24 ■ Dynamic range of a signal-following ADC.

11.7.1 Spurious-Free Dynamic Range

The dynamic range of a receiver is limited not only by the dynamic range of the ADC and associated components but also by the spurious-free dynamic range. Spurs can cause signals in the receiver, which appear as false targets or can mask small targets. Spurs can be generated in the mixing (downconversion) process, the ADC, and other possible nonlinear sources in the receiver. It is necessary to suppress these spurs as much as possible to maintain a useable dynamic range for the receiver. Software is currently available to assist the designer in determining the overall spurious-free dynamic range of the receiver.

Figure 11-25 shows an example implementation of a direct digital receiver where the RF following band-pass filtering and amplification is directly digitized to enable the digital signal processor to form the I/Q signals [15]. Figure 11-26 shows the spurious signals associated with the 2 Gigasample/second ADC multichip module (MCM). This spurious response at twice the test signal frequency limits the dynamic range of the receiver to about 40 dB.

11.7.2 Direct Digital Coherent Detection Implementation

Direct digital in-phase/quadrature (I/Q) sampling is being used in most new radar implementations to obtain coherent video representation of the IF signal. This is primarily enabled because of advances in the sampling speeds of current analog-to-digital converters. The main advantage of direct digital sampling is that it eliminates the errors associated with phase and amplitude offsets associated with analog I/Q downconversion. Figure 11-27 shows a typical implementation of a direct digital sampler [11]. It is worth

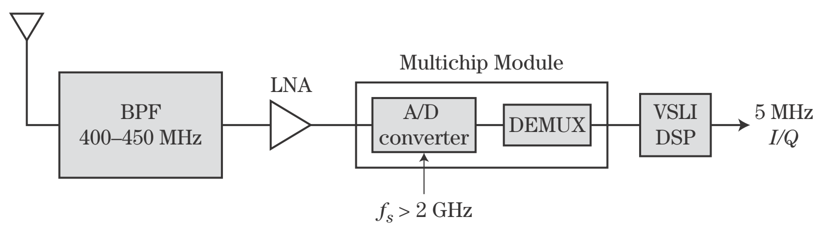


FIGURE 11-25 ■ Block diagram of a direct digital radar receiver. (From Thompson [15]. With permission.)

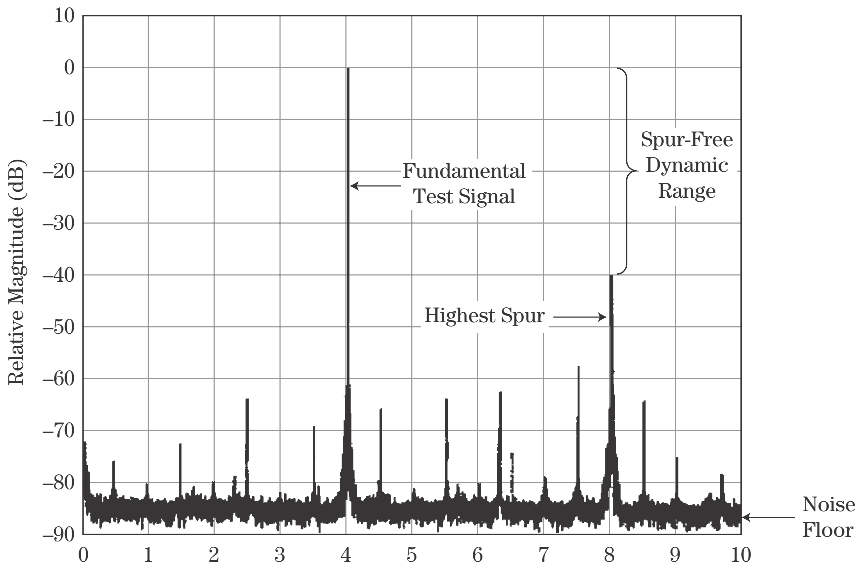


FIGURE 11-26 Spurious-free dynamic range associated with the ADC MCM. (From Thompson [15]. With permission.)

pointing out that the cos($2\pi n/4$) and sin($2\pi n/4$) multipliers in Figure 11-27 are simply sequences of 1's, 0's, and −1's, and the downstream complex processing rate is the IF bandwidth. Also, there are other implementations with 1's, –1's, and 0's multipliers that can reduce the ADC sampling requirement slightly over two times the IF bandwidth.

The sampling frequency f_S needs to be chosen to be four times the IF center frequency, f_0, but in most cases this would require an extremely high sample frequency. Lyons ([11], Chapter 8) points out that since the IF signal is normally band-limited, the sampling can be performed at a lower frequency. For example, suppose the IF center frequency is 60 MHz and the IF bandwidth is $\leq$ 12 MHz; instead of sampling the IF at 240 MHz, the sample frequency could be chosen as 48 MHz. The sampled IF signal replicates in the frequency domain with one of the sets of spectral components centered at 12 MHz. The associated sine and cosine signals shown in Figure 11-27 could then be at a frequency of 12 MHz, which satisfies the requirement that the sampling frequency be four times the frequency of the sine and cosine multipliers. The digital I and Q signals can be derived by low-pass filtering the in-phase ($x_I[n]$) and quadrature phase ($x_Q[n]$) signals.

Urkowitz [16] describes an alternate method for generating the real and imaginary components of a signal using a Hilbert transformer, such as that shown in Figure 11-28. A Hilbert transformer has the property of providing a phase shift of $\pi/2$ (90°) to the digital representation of a signal. The IF signal is sampled, as shown in Figure 11-28, by a high-speed ADC at a sample rate equal to or greater than two times the IF. The direct signal out of the ADC is then usually called the real component, while the Hilbert transformed

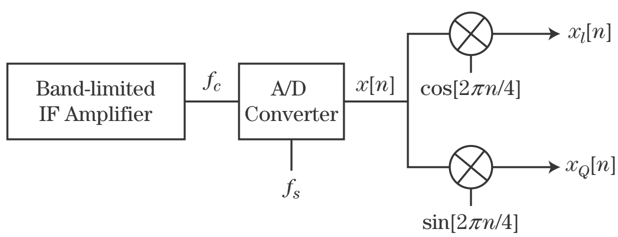


FIGURE 11-27 Direct digital I/Q converter.

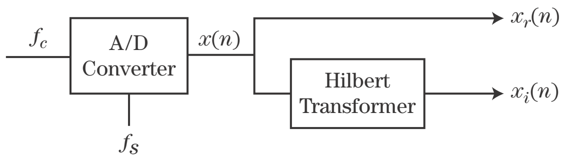


FIGURE 11-28 ■ Derivation of real and imaginary components using Hilbert transform.

output is the imaginary component. These signals can then be used to form the in-phase (I) and quadrature phase (Q) components. A detailed description of Hilbert transformer implementation is provided in ([11], chapter 9).

11.7.3 Digital Up/Down Frequency Conversion

In legacy systems, frequency up/down conversions at RF and IF were normally performed using analog signals with mixers and associated analog filters and amplifiers. Digital receiver and exciter technologies have advanced to the point that some of these frequency conversions can currently be performed digitally, thus reducing the cost and complexities associated with their analog equivalents. Oppenheim and Schafer [17, p. 798] describe an implementation for performing single-sideband frequency up/down conversions using Hilbert transformers.

11.8 FURTHER READING

The design of analog radar receivers has been fairly consistent over the past several decades, evolving primarily with improvement in analog components. This chapter is a general overview of radar receivers, and the reader is advised to investigate the references cited in the chapter to gain a more in-depth understanding of the particular aspects of receiver design and use. The effects of systematic error effects on receiver performance can be found in *Coherent Radar Performance Estimation* by Scheer and Kurtz [18].

With the rapid advances in signal processing components, more of the basic receiver functions are being implemented digitally. Although there is a tendency to consider anything past the ADC as signal processing, many of the functions have similar concerns as their analog counterparts. It is strongly recommended that engineers interested in receiver design become familiar with digital implementation of these functions. One strongly recommended book to help a designer become familiar with digital receiver design is R. G. Lyons's [11] *Understanding Digital Signal Processing*, which thoroughly explains the digital concepts in an easy-to-understand manner and without requiring advanced mathematics. This text also deals with many of the component items involved in radar receivers. After becoming familiar with his book, the advanced concepts in Oppenheim and Schafer's [17] *Discrete-Time Signal Processing* will become clearer to the reader.

11.9 REFERENCES

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11.10 PROBLEMS

1. Why is receiver protection necessary for radar receivers?

2. Consider the case of a pulsed radar at RF of 9.4 GHz that is being jammed by a broadband jammer covering the bandwidth from 9.0 to 10.0 GHz with uniform random noise. If the radar has a 20 MHz bandwidth preselect filter, what is the potential reduction in jammer power due to the preselect filter?

3. For a radar with an RF center frequency of 9.6 GHz and a desired first IF of 2.4 GHz. What is the LO frequency if it is below the RF? Determine if any of the intermodulation components fall into the IF for $n = 1$ to 3 and $m = 1$ to 4.

4. For Problem 4, consider changing the IF center frequency to 3.0 GHz. What is the new LO frequency (LO below the RF)? For $n = 1$ to 3, and $m = 1$ to 5, what is the closest intermodulation component to the IF center frequency? Will an IF with a 1 GHz bandwidth with a sharp cut-off at $\pm$ 500 MHz from the IF center frequency reject the intermodulation component?

5. For a radar with a logarithmic amplifier that provides a 0.05 volt/dB increase in the video output voltage of the logarithmic amplifier per dB increase in input signal power, what is the voltage increase for a 35 dB increase in input signal level?

6. A coherent radar for a particular target has video output voltage of $Q = +1.2$ volts and $I = -0.5$ volts. What is the magnitude and phase of the target signal?

7. Use the radar equation from Chapter 2 to solve the following problem. For a radar with a 100 kW peak power transmitter with an antenna gain of 30 dB, noise figure of 4 dB, losses of 6 dB, and receiver bandwidth of 1 MHz, find the maximum range a radar can provide a 13 dB SNR on a target with 10 m^2 radar cross section (assume that $kT_0 = -204$ dBW).

8. Consider a receiver with three stages: The first stage has a noise figure of 3.0 and a gain of 23 dB; the second stage has a noise figure of 10 and a gain of 30 dB; and the third stage has a noise figure of 15. Determine the overall noise figure for the receiver.

9. Why is it necessary to increase the noise figure stated in mixer specifications by 3 dB when used in superheterodyne receivers?

10. What is the purpose of sensitivity time control?

11. Assume that the effective antenna temperature is 50 °C (Note: K $\approx$ °C + 270), $T_{tr} = T_0 =$ 290 K. When the received signal at the antenna is 1 mW the signal at the receiver input is 0.9 mW and the receiver noise figure is 2.5 (4 dB). What is the effective system noise temperature, T_S?

12. Why is it necessary to set the input noise level to provide one to two least significant bits? Why not just set the noise level below the least significant bit to provide maximum dynamic range from the ADC?

13. Assume that the maximum signal into a 12-bit ADC is set to 1 dB below its saturation and the input noise level is set to about 2 LSBs. Assuming 6 dB/bit, what is the effective dynamic range from the ADC?

14. An ADC has a noise floor about 80 dB down relative to a peak signal near its maximum and spurious responses down below the peak signal by 70 dB, 65 dB, and 45 dB. What is the spurious-free dynamic range of the ADC?

15. Why does the sampling frequency for the in-phase and quadrature phase circuit in Figure 11-27 have to be four times the frequency of the sine and cosine multiplier functions?