Chapter 6
Radar Receivers*

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6.1 THE CONFIGURATION OF A RADAR RECEIVER

The function of a radar receiver is to amplify, filter, downconvert, and digitize the echoes of the radar transmission in a manner that will provide the maximum discrimination between desired echo signals and undesired interference. The interference comprises not only the self noise generated in the radar receiver but also the 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 classified as interference and is commonly categorized as clutter. Where airborne radars are used for altimeters or mapping, other aircraft are undesired targets, and the ground is the desired target. In the case of weather radars, ground, buildings, and aircraft are clutter, and rain or snow is the desired target. More commonly, radars are intended for detection of aircraft, missiles, ships, surface vehicles, or personnel, and the reflection from weather, sea, or ground is classified as clutter interference.

Although the boundaries of the radar receiver are somewhat arbitrary, this chapter will consider those elements identified in Figure 6.1 as the receiver. The radar exciter generates the transmit waveforms as well as local oscillator (LO), clock, and timing signals. Since this function is usually tightly coupled to a radar receiver, it is also shown in Figure 6.1 and will be discussed in this chapter. The purpose of Figure 6.1 is to illustrate the functions typical of a modern radar receiver and exciter.

Virtually all radar receivers operate on the superheterodyne principle shown in Figure 6.1. Through this architecture, the receiver filters the signal to separate desired target signals from unwanted interference. After modest RF amplification, the signal is shifted to an intermediate frequency (IF) by mixing with a local-oscillator (LO) frequency. More than one conversion stage may be necessary to reach the final IF without encountering serious image- or spurious-frequency problems in the mixing process. The superheterodyne receiver varies the LO frequency to follow any desired tuning variation of the transmitter without disturbing the filtering at IF. This simplifies the filtering operation as the signals occupy a wider percentage

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bandwidth at the IF frequency. These advantages have proven to be so significant that competitive forms of receivers have virtually disappeared.

In conventional antenna systems, the receiver input signal is derived from the duplexer, which permits a single antenna to be shared between transmitter and receiver. In active array systems, the receiver input is derived from the receive beam-forming network. Active array antennas include low-noise amplifiers prior to forming the receive beams; although these are generally considered to be antenna rather than receiver components, they will be discussed in this chapter.

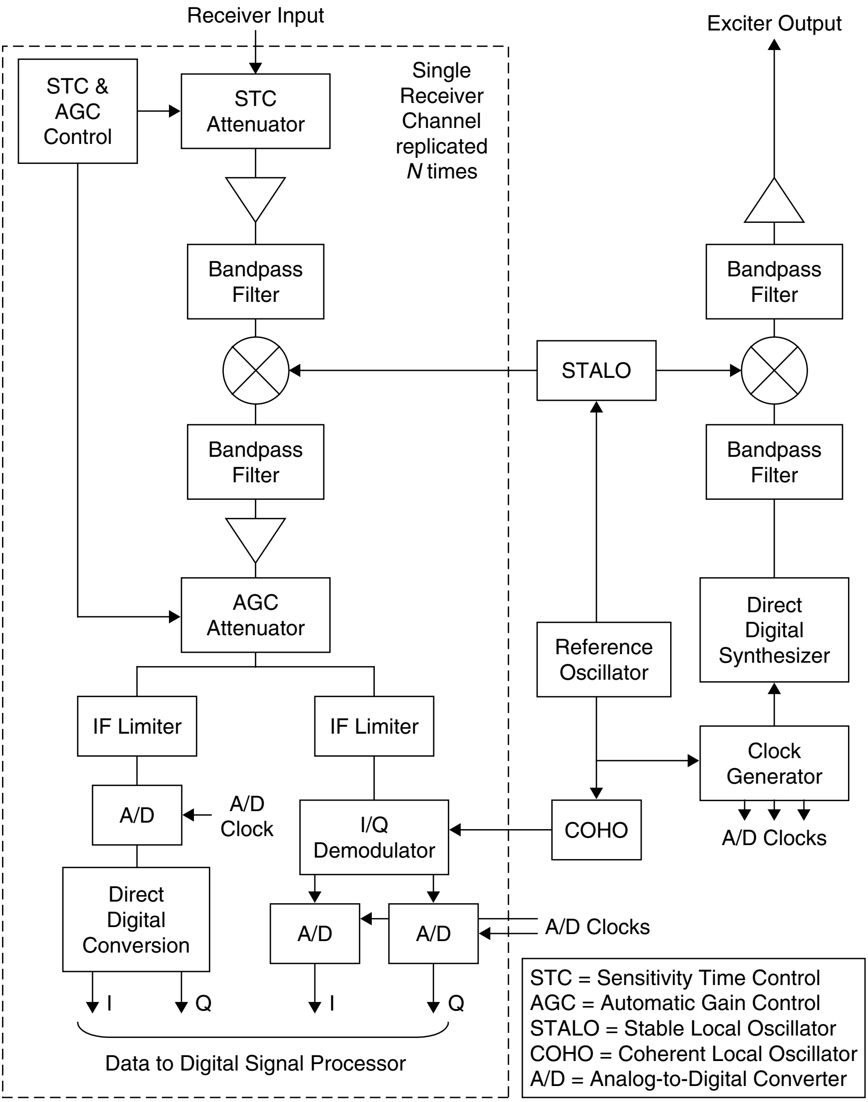

FIGURE 6.1 General configuration of a radar receiver

The block diagram shown in Figure 6.1 includes sensitivity time control (STC) attenuation at the RF input. Alternatively adjustable RF attenuation may be used. Either form provides increased dynamic range above that provided by the analog-to-digital (A/D) converters. RF attenuation is described in more detail in Section 6.6. The STC attenuator is followed by an RF amplifier, often referred to as a low-noise amplifier (LNA). This amplifier provides sufficient gain with a low noise figure to minimize the subsequent degradation of the overall radar noise figure by subsequent components. If sufficient gain is provided in the antenna prior to the receiver, it may be possible to eliminate this gain stage. The RF filter provides rejection of out-of-band interference, including rejection at the RF image frequency. After downconversion to IF, a bandpass filter provides rejection of unwanted signals and sets the receiver analog-processing bandwidth. Additional gain is provided at IF to overcome losses and raise the signal level required for subsequent processing and to set the correct signal level into the A/D converters. An IF limiter provides graceful limiting of large signals that would otherwise overload the A/D converters.

The two dominant methods of digitization, IF sampling and analog I/Q demodulation with baseband A/D conversion, are included for illustration in Figure 6.1, though in general, receivers will not include both techniques. Prior to the availability of affordable digital signal processing, a number of functions, such as monopulse comparison, currently performed in the digital domain, were performed using analog processing within the receiver. Readers interested in the details of these analog processing techniques will find details in the first and second editions of this handbook.[1,2]

All but the simplest of radars require more than one receiver channel. Figure 6.1 shows a single receiver channel that may be replicated any number of times depending on the radar system requirements. Monopulse radars typically include three receiver channels, sum, delta azimuth, and delta elevation channels, used to provide improved angle accuracy. Additionally, many military radar systems include a sidelobe blanker or several sidelobe canceler channels to combat jamming. Since the advent of digital beamforming radar systems, the number of receiver channels required has increased dramatically, with some systems now requiring hundreds of receiver channels. In these multichannel receiver systems, close matching and tracking of gain and phase is required. Receiver channel tracking and equalization are discussed in Section 6.11.

The stable local oscillator (STALO) block provides the local oscillator frequencies for downconversion in the receiver and upconversion in the exciter. For true coherent operation, the STALO is locked to a low frequency reference, shown by the reference oscillator in Figure 6.1 that is used as the basis for all clocks and oscillators such as the coherent local oscillator (COHO) within the receiver and exciter. The clock generator provides clocks to the A/D converters and the direct digital synthesizer and provides the basis for the signals that define the radar transmit and receive intervals.

The direct digital synthesizer in Figure 6.1 is used to generate the transmit waveforms at an IF frequency prior to upconversion to the RF output frequency. Filtering in the exciter is required to reject aliased signals from the direct digital synthesizer and unwanted mixer products. RF gain is typically required to provide a sufficient drive level to the transmitter or phased array antenna.

Almost all modern radar systems use digital signal processing to perform a variety of functions, including pulse compression and the discrimination of desired targets from interference on the basis of velocity or the change in phase from one pulse to the next. Previously, pulse compression was performed using analog processing with dispersive delay lines, typically surface acoustic-wave (SAW) devices. Analog pulse compression has largely been replaced by pulse compression using digital signal processing.

In the case of very wideband waveforms, analog stretch processing (see Section 6.3) may be used to reduce the signal bandwidth before subsequent digital signal processing.

The receiver discussed herein focuses on those functions that provide analog processing and digitization of the individual pulse signals with the minimum of distortion, enabling subsequent digital signal processing to maximize the performance of the radar. The digital signal processing function is not normally considered to be part of the receiver.

6.2 NOISE AND DYNAMIC-RANGE CONSIDERATIONS

Receivers generate internal noise that masks weak signals being received from the radar transmissions. This noise contribution, which can be expressed as either a noise temperature or a noise figure, is one of the fundamental limitations on the radar range.

The noise temperature or noise figure 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 other performance parameters.

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

- Dynamic range and susceptibility to overload
- Instantaneous bandwidth and tuning range
- Phase and amplitude stability

A direct compromise must be made between the noise figure 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:

	Example 1	Example 2	Example 3
Ratio of front-end noise to mixer noise	6 dB	10.0 dB	13.3 dB
Sacrifice in mixer dynamic range	7 dB	10.4 dB	13.5 dB
Degradation of system noise temperature due to mixer noise	1 dB	0.4 dB	0.2 dB

The same considerations apply to the setting of the noise level at the input to the A/D converters. Traditionally, the noise contribution of the A/D converter was considered by the system engineers as a separate contribution to the overall radar system noise, distinct from receiver noise, and was accounted for at the system level. Today, it has become common to include the A/D converter noise as part of the overall receiver noise. Consequently, it is important to understand whether or not the contribution of the A/D converter is included in the specification for the noise figure of a receiver.

In active array antennas, and many conventional antennas, low-noise amplifiers (LNAs) establish the system noise floor prior to the receiver input. The noise from the antenna is usually set well above the receiver noise floor such that the receiver has only a small impact on overall system noise. Again, the trade-off must be performed between system dynamic range and noise figure.

Definitions. Dynamic Range represents the range of signal strength over which the receiver will perform as expected. It requires the specification of a minimum level, typically the noise floor, the maximum level that can be handled with some allowable deviation from the ideal response, and the type of signal to be handled. These parameters are defined through a variety of characteristics as described below.

Modern radars systems increasingly rely solely on linear receiver channels followed by digital signal processing, providing both increased flexibility and near ideal signal-detection characteristics. Previously, a variety of limiting or logarithmic receiver approaches were used to perform various signal-processing functions. These receivers must define an allowable error in their outputs relative to their ideal nonlinear response.

Receivers that include some form of gain control must distinguish between instantaneous dynamic range and the total dynamic range that is achieved as a result of programmed gain variation.

Receiver Input Noise Level. Because many radar systems include low-noise amplifiers prior to the input of the receiver, it is important to understand and specify the noise level at the receiver input. This noise level is set by the antenna noise temperature and its total effective noise gain or loss. The noise level can be specified either as an rms power in a specified bandwidth or as a noise power spectral density.

System Noise. The system noise level is the combined antenna and receiver noise. Typically, the receiver input noise will exceed that of the noise due to the receiver itself, so that the receiver has only a small impact on the system noise temperature or noise figure. Thus, when defining dynamic-range parameters, such as signal-to-noise ratio, it is important to specify whether the noise level being referenced is the receiver noise or total system noise.

Minimum Signal of Interest. Minimum signal definitions such as minimum-detectable-signal or minimum-discernable-signal have been used in the past; however, these definitions have become less common due to the extensive use of digital signal-processing techniques. Digital signal processing of the receiver output allows the detection of signals well below the receiver noise floor and the minimum detectable level depends on the nature of the processing performed.

Signal-to-Noise Ratio (SNR). SNR is the ratio of the signal level to that of the noise. SNR is typically expressed in decibels (dB). The maximum receiver SNR is set by the noise contribution and maximum signal capability of every component in the chain; however, since the limiting technology is often the Analog-to-Digital (A/D) converter, the preceding components and gain structure are often chosen such that the maximum SNR is driven by the performance of the A/D converter. More details of the relationship between A/D converter and receiver SNR are included in Sections 6.10 and 6.11.

Spurious Free Dynamic Range (SFDR). SFDR is the ratio of the maximum signal level to that of largest spurious signal created within the receiver. SFDR is typically expressed in decibels (dB). This parameter is determined by a variety of factors including the mixer intermodulation spurious (described in more detail in Section 6.4), the spurious content of the receiver local oscillators, the performance of the A/D converter, and the many sneak paths that may result in unwanted signals coupling onto the receiver signal path.

Intermodulation Distortion (IMD). Intermodulation distortion is a nonlinear process that results in generation of frequencies that are linear combinations of the fundamental frequencies of the input signals. Second and third order intermodulation are the most commonly specified, and the performance of the receiver is usually specified in terms of two-tone second and third order input intercept points. The intercept point is the extrapolated level at which the power in the intermodulation product equals that of the two fundamental signals.

For input signals at frequencies f_1 and f_2, second order intermodulation distortion produces signals at frequencies: $0, f_1 - f_2, f_1 + f_2, 2f_1$ and $2f_2$. Third order intermodulation distortion produces signals at frequencies: $2f_1 - f_2$, $2f_2 - f_1$, $2f_1 + f_2$, $f_1 + 2f_2$, $3f_1$ and $3f_2$. For narrow band signals, only the third order products $2f_1 - f_2$ and $2f_2 - f_1$ fall in band, and consequently, third order distortion is typically the primary concern. The power levels of these third order intermodulation products are given by

$$P_{2f1-f2}(dBm) = 2P_{f1}(dBm) + P_{f2}(dBm) - 2P_{\mathrm{IP}}(dBm) \tag{6.1}$$

$$P_{2f2-f1}(dBm) = P_{f1}(dBm) + 2P_{f2}(dBm) - 2P_{\mathrm{IP}}(dBm) \tag{6.2}$$

where P_{f1}(dBm) = power of input signal at frequency f_1 in dBm
P_{f2}(dBm) = power of input signal at frequency f_2 in dBm
P_{IP}(dBm) = third order intercept point in dBm

Intermodulation can result in a variety of undesirable effects such as

- Intermodulation of clutter returns causing broadening of clutter doppler width, resulting in the masking of targets
- Unwanted in-band signals due to out-of-band interfering signals, resulting in false targets
- Intermodulation products from in-band signals that cannot be readily cancelled through linear cancellation techniques, resulting in susceptibility to jammers

Intermodulation distortion occurs throughout the receiver chain. Consequently, the receiver will have a significantly different input intercept point, depending on the signal frequency relative to the radio frequency (RF), IF, and video filter bandwidths. It is, therefore, important to distinguish between the requirements for in-band and out-of-band intermodulation distortion as different signals have different effects on the receiver.

Cross-Modulation Distortion. Cross-modulation occurs as a result of third order intermodulation, whereby the amplitude modulation (AM) of one signal, typically an unwanted interference signal in the operating RF band but usually outside the tuned signal bandwidth, is transferred onto the desired signal.

The resultant percent AM modulation, $\%d$, on the desired signal is given by[3]

$$\%d = \%u \frac{4P_U}{P_{IP} + 2P_U} \tag{6.3}$$

where $\%u$ = percent AM modulation of the unwanted signal
P_U = power of unwanted signal
P_{IP} = third order intercept point

Cross modulation can result in the modulation of clutter and target returns due to large amplitude modulated out-of-band interferences resulting in poor clutter cancellation and poor range sidelobe performance.

1 dB Compression Point. The input 1 dB compression point of a receiver is a measure of the maximum linear signal capability and is defined as the input power level at which the receiver gain is 1 dB less than the small signal linear gain. Receiver gain compression can result from compression in amplifiers, mixers, and other components throughout the receiver chain. Typically, the receiver is designed to provide controlled gain compression through a limiting stage at the final IF as described in Section 6.8.

Analog-to-Digital Converter Full Scale. The A/D converter full scale level determines the maximum level that can be digitized. Receivers typically provide controlled limiting (Section 6.8) to prevent the signal level from exceeding the full scale level of the A/D converter. Practical considerations mean that the hard limit level is typically set 1 dB below full scale to prevent overload as a result of component tolerance variations.

Types of Signals. Various types of signals are of interest in determining dynamic-range requirements: distributed targets, point targets, wideband noise jamming, and narrow band interference. 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 for distributed clutter; the time-bandwidth product of the coded pulse indicates the added dynamic range that the decoder will extract from the point targets. Conversely, if the radar incorporates an excessively wide-bandwidth RF amplifier, its dynamic range may be severely restricted due to wideband noise interference.

When low-noise amplifiers (LNAs) are included in the antenna, prior to forming the receive beams, the antenna 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 returns 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. In modern radar architectures that employ digital beamforming, nonlinearity at any stage of the receiver channel will create similar problems.

System calibration techniques and adaptive beamforming techniques can compensate for linear gain and phase deviations; however, as for the case of the LNA nonlinearities described above, compensation for nonlinear characteristics is either impractical or impossible when the cause of the nonlinear distortion is outside the digitized bandwidth.

Evaluation. A thorough evaluation of all elements of the receiver is necessary to prevent unanticipated degradation of noise figure or dynamic range. Inadequate dynamic range makes the radar receiver vulnerable to interference, which can cause saturation or overload, masking or hiding the desired signals. A tabular format for such a computation (a typical example of which is shown in Table 6.1) 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.

TABLE 6.1 Noise and Dynamic-Range Characteristics

	Units	Input	STC Attenuator	Amplifier	Bandpass Filter	Mixer	Bandpass Filter	Amplifier	AGC Attenuator	Limiter	A/D Converter
Component Noise Figure	dB		3.0	5.0	0.5	6.5	5.0	4.0	6.0	14.0	
Component Gain	dB		−3.0	12.0	−5.0	−6.5	−0.2	20.0	−6.0	0.0	
Component Output 3rd Order Intercept	dBm		43.0	32.0	50.0	20.0	50.0	38.0	40.0	30.0	
Component Output 1dB Compression Point	dBm		30.0	18.0	40.0	10.0	40.0	23.0	30.0	−1.0	
Cumulative Gain	dB		−3.0	9.0	8.5	2.0	0.0	20.0	14.0	14.0	
Cumulative Noise Figure	dB		3.00	8.00	8.01	8.33	9.13	9.86	9.88	10.29	
Cumulative Output 3rd Order Intercept	dBm		43.0	32.0	31.4	18.8	16.8	34.3	28.1	25.9	
Cumulative Output 1dB Compression Point	dBm		30.0	18.0	17.5	7.4	5.4	21.0	14.9	−1.1	
Receiver Noise Level	dBm/Hz		−174.0	−157.0	−157.5	−163.7	−164.9	−144.1	−150.1	−149.7	
System Noise Level	dBm/Hz	−149.0	−152.0	−139.0	−140.4	−146.9	−148.9	−128.9	−134.9	−134.9	
Bandwidth	MHz		1000	1000	100	100	10	10	10	10	
A/D SNR in Nyquist BW	dB										70.0
A/D Converter Sample Rate	MHz										100.0
A/D Full Scale Level	dBm	−14.0	−17.0	−5.0	−5.5	−12.0	−14.0	6.0	0.0	0.0	0.0
A/D Noise Level	dBm/Hz										−147.0
System Noise Relative to A/D Noise	dB										12.1
Maximum Point Clutter or Target Level	dBm	−20.0	−23.0	−11.0	−11.5	−18.0	−20.0	0.0	−6.0	−6.0	

6.3 BANDWIDTH CONSIDERATIONS

Definitions. The instantaneous bandwidth of a component is the frequency band over which the component can simultaneously process two or more signals to within a specified accuracy. When the term *instantaneous bandwidth* is used as a radar receiver parameter, it refers to the resulting bandwidth set by the combination of RF, IF, video, and digital filtering that occurs within the receiver.

When the radar receiver employs stretch processing (defined later in this section), the RF processing bandwidth is significantly larger than the IF bandwidth. Consequently, the term *instantaneous bandwidth* can be confusing. Confusion can be avoided by using the terms *RF waveform bandwidth, LO linear FM (chirp) bandwidth*, and *IF processing bandwidth*. The relationship between RF, LO, and IF bandwidths used in stretch processing is explained in more detail later.

The tuning range is the frequency band over which the component may operate without degrading the specified performance. Tuning is typically accomplished by adjusting the local oscillator frequency and adjusting the RF filtering characteristics. The frequency range over which the radar operates is often referred to as the *operating bandwidth*.

Important Characteristics. The environment in which a radar must operate includes many sources of electromagnetic radiation, which can mask the relatively weak returns 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 Section 6.7. However, one cannot ignore the RF portion of the receiver merely by making it have wide bandwidth. Section 6.2 discussed 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 (e.g., other radars, TV stations, or microwave communication links) that, 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. The nonlinearities and imbalance of mixers is described in more detail in Section 6.4.

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 flexibility to escape interference, but if the interference is intentional, as in the case of jamming, a change in RF frequency on a pulse-to-pulse basis may be required using switchable or electronically tuned filters. If the RF filtering is located prior to RF amplification, the filter insertion loss will have a dB for dB impact on the receiver noise figure, another sacrifice in noise temperature to achieve more vital objectives. Yttrium iron garnet (YIG) filters and pin diode switched filters have been used to provide the necessary frequency agility.

Stretch Processing. Stretch processing is a technique frequently used to process wide bandwidth linear FM waveforms. The advantage of this technique is that it allows the effective IF signal bandwidth to be substantially reduced, allowing digitization and subsequent digital signal processing, at more readily achievable sample rates. By applying a suitably matched chirp waveform to the receiver first LO, coincident with the expected time of arrival of the radar return, the resultant IF waveform has a significantly reduced bandwidth for targets over a limited range-window of interest. Provided that the limited-range window can be tolerated, a substantially reduced processing bandwidth allows more economical A/D conversion and subsequent digital signal processing. It also allows a greater dynamic range to be achieved with lower-rate A/D converters than would be achievable if digitization of the entire RF signal bandwidth were performed.

If the LO chirp rate is set equal to the received signal chirp rate of a point target, the resultant output is a constant frequency tone at the output of the stretch processor receiver, with frequency $\Delta t B/T$, where Δt is the difference in time between the received signal and the LO chirp signal, and B/T is the waveform chirp slope (chirp bandwidth/pulse width). Target doppler is maintained through the stretch processing, producing an output frequency offset equal to the doppler frequency, though the wide percentage bandwidth often used means that the doppler frequency can change significantly over the duration of the pulse.

Ignoring the effect of target doppler, the required RF signal bandwidth is equal to the transmitted waveform bandwidth. Given the RF signal bandwidth B_R, the received pulse width T_R, and the range interval ΔT, the required LO reference waveform duration is given by

$$T_L = T_R + \Delta T \tag{6.4}$$

the LO reference chirp waveform bandwidth is given by

$$B_L = \frac{T_R + \Delta T}{T_R} B_R \tag{6.5}$$

and the IF processing bandwidth is given by

$$B_I = \frac{\Delta T}{T_R} B_R \tag{6.6}$$

6.4 RECEIVER FRONT END

Configuration. The radar *front end* consists of a low-noise amplifier (LNA) and bandpass filter followed by a downconverter. The radar frequency is downconverted to an IF, 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. Occasionally, receivers will include filtering before the LNA in order to limit the effects of intermodulation distortion that can occur in the LNA. Even when filtering is included before the LNA, a second filter is often still required between the LNA and the mixer in order to reject the amplifier noise at the image frequency. Without this filter, the noise contribution of a broadband LNA would be doubled.

The receiver front end may also include a limiter, used to protect the receiver circuitry from damage due to high power that may occur either from leakage during transmit mode or as a result of interference from another system such as a radar at close range. Front-end limiters are discussed in more detail in Section 6.8.

The radar or receiver front end often includes some form of gain or attenuation control as shown in Figure 6.1. Gain control is described in more detail in Section 6.6.

Effect of Characteristics on Performance. Noncoherent pulse radar performance is affected by front-end characteristics in three ways. Noise introduced by the front end increases the radar noise temperature, degrading sensitivity, and limits the maximum range at which targets are detectable. Front-end saturation on strong signals may limit the minimum range of the system or its ability to handle strong interference. Finally, the front-end spurious performance affects the susceptibility to off-frequency interference.

Coherent radar performance is even more affected by spurious mixer characteristics. Range and velocity accuracy is degraded in pulse doppler radars; stationary target cancellation is impaired in MTI (moving-target indication) radars; 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. Spurious doppler spectra levels are dictated by requirements to suppress clutter interference through doppler filtering.

Harmonics are generated by any component that becomes nonlinear when subjected to the power level created by the transmitter and that 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 devices and can generate harmonics.

Spurious doppler spectra are created by any process that does not reoccur identically 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 Response of Mixers. The ideal mixer acts as a multiplier, producing an output proportional to the product of the two input signals. The input RF signal at frequency f_R is frequency shifted or modulated by the LO signal at frequency f_L. Balanced mixers are used to minimize conversion loss and unwanted spurious responses. In active mixers, modulation is performed using transistors, and in passive mixers, the modulation is performed using Schottky-barrier diodes or other solid-state devices (e.g., MESFET) where increased dynamic range is required.

The resulting output signal frequencies ($f_L + f_R$ and $f_L - f_R$) are the sum and difference of the two input frequencies. In practice, all mixers produce unwanted intermodulation spurious responses[4] with frequencies $\pm nf_L \pm mf_R$ (where m and n are integers), and the degree to which these spurious products impact the radar performance depends upon the type of mixer and the overall radar performance requirements. Analysis of mixer spurious levels is nontrivial, and the receiver designer typically requires tabulated data generated through mixer characterization measurements to predict mixer spurious levels.

Advances in mixer technology have resulted in a wide variety of commercially available devices employing balanced, double balanced, and double-double balanced topologies covering a wide range of RF, LO, and IF frequencies and a range of performance characteristics.

Mixer Spurious-Effects Chart. A graphical display of mixer spurious components up to the sixth order is shown in Figure 6.2. This chart allows identification of those combinations of input frequencies and bandwidths that are free of strong low-order spurious components. Such charts are most useful in determining optimum IF and LO frequencies during the initial design phase. Once the frequency plan has been determined, computer analysis of spurious responses is typically used to ensure spurious free performance over the entire range of LO frequencies and RF and IF bandwidths.

The heavy line in Figure 6.2 represents the desired signal and shows the variation of normalized output frequency $(H - L)/H$ with normalized input frequency L/H. All other lines on the chart represent the unwanted spurious signals. To simplify use of the chart, the higher input frequency is designated by H and the lower input frequency by L.

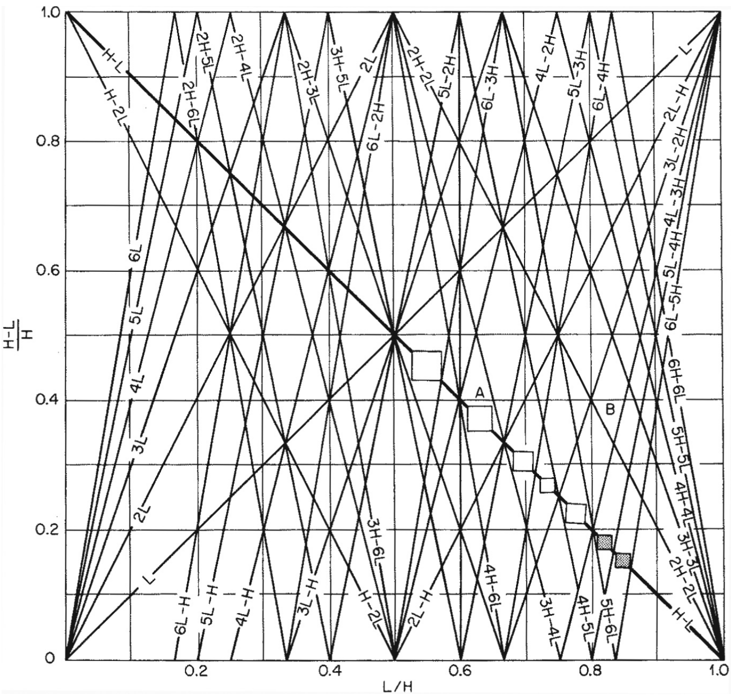

FIGURE 6.2 Downconverter spurious-effects chart: H = high input frequency; L = low input frequency

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 cannot be corrected by IF filtering. The $4H - 6L$ and $3H - 4L$ spurious frequencies, like all spurious IF frequencies, arise from cubic or higher-order intermodulation.

The available spurious-free bandwidth in any of the designated regions is roughly 10% 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 high-order terms in the power-series model and are consequently low enough in amplitude that they can often be ignored. For this reason, a low IF generally provides better suppression of spurious responses.

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. If sufficient filtering cannot be applied prior to the mixing process, spurious products that fall within the operating band will no longer be filterable, which will seriously degrade system performance.

Spurious responses not predicted by the chart occur when two or more RF input signals produce other frequencies by intermodulation that lie within the RF passband.

Image-Reject Mixer. A conventional mixer has two input responses 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 Figure 6.3. The RF hybrid produces a 90° phase differential between the LO inputs to the two 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. In practice, image reject mixers often do not provide sufficient rejection of the image response alone without filtering. In this case, they can be used in conjunction with an image rejection filter, reducing the magnitude of rejection required by the filter.

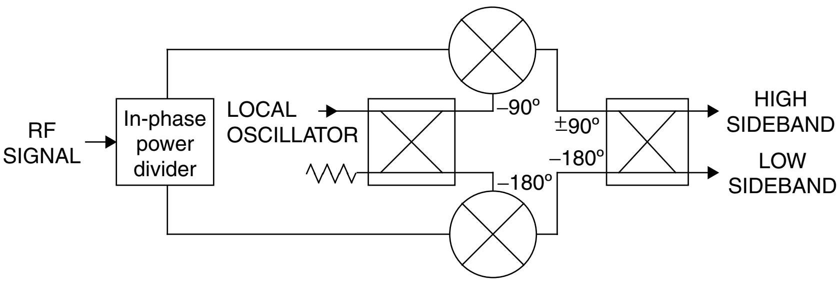

FIGURE 6.3 Image reject mixer

Characteristics of Amplifiers and Mixers. Noise figure, amplifier gain, mixer conversion loss, 1 dB compression point, and third order intercept point are the most common performance parameters specified for amplifiers and mixers. Occasionally, a second order intercept point specification is also required for very wide bandwidth signals. It should be noted that for amplifiers, compression point and third order intercept are usually specified at their output whereas for mixers these parameters are usually specified at their input.

Additional specifications for mixers include LO drive power, port-to-port isolation, and single tone intermodulation levels. The LO drive power specification defines how much LO power is required by the mixer to meet its specified performance levels. Typically, the higher the LO power, the higher the 1 dB compression point and third order intercept point. Radar receivers often require high LO drive level mixers in order to meet the challenging dynamic-range requirements. The port-to-port isolation is used to determine the power level coupled directly between the mixer ports without frequency translation. The single tone intermodulation levels specify the levels of the $\pm nf_L \pm mf_R$ spurious signals, as discussed previously.

6.5 LOCAL OSCILLATORS

Functions of the Local Oscillator. The superheterodyne receiver utilizes one or more local oscillators and mixers to convert the signal 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. These LOs are generally also used in the exciter to upconvert modulated waveforms to RF for output to the transmitter.

In many early radars, the only function of the local oscillators was conversion of the input signal frequency to the correct intermediate frequency. Many modern radar systems, however, coherently process a series of returns from a target. The local oscillators act essentially as a timing standard by which the signal 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.

STALO Instability. The first local oscillator, generally referred to as a *stable local oscillator (STALO)*, typically has the greatest effect on receiver-exciter stability; however, when evaluating the overall performance, other contributions should not be neglected. Advances in state-of-the-art STALO oscillator performance and the stringent clutter cancellation requirements of modern radars means that the phase noise of all oscillators and timing jitter of A/D converter and D/A converter clocks and T/R strobes may be significant.

The short-term stability requirements of the STALO are generally characterized by device noise relative to carrier (dBc), specified in terms of a phase noise spectrum and measured in the frequency domain. Long-term stability is typically characterized by aging and environmental effects, specified in terms of frequency drift and measured using an Allan Variance[5] technique. Requirements are typically specified in terms of an absolute frequency tolerance or a maximum frequency deviation over some time interval.

It should be noted that measurements of phase noise are typically performed by measurement of double-sideband noise, the sum of the power in both the upper and lower sidebands, but more typically reported and specified as single sideband (SSB) values. Double-sideband noise can be translated to a single-sideband value by subtracting 3 dB. Unequal sideband power can only result from additive signals or noise or correlated amplitude and phase noise components.

Amplitude modulation (AM) of the STALO is typically not a significant factor as it is usually at a lower level than the phase noise (at small offset frequencies from carrier) and can be further reduced through limiting. Modern mixers typically provide a significant reduction in the effect of STALO amplitude modulation as their conversion gain is relatively insensitive to LO power variation when operated at their specified drive level.

For systems requiring high sensitivity, AM noise can become disruptive if unintentional conversion of AM to PM noise occurs in the receiver chain. This process can occur via suboptimum component bias techniques where high amplitude signals or noise create a phase shift resulting in another phase noise contribution to the receiver chain.

Vibration Sensitivity. In addition to the phase noise generated by the STALO in a benign environment, sources of unwanted phase modulation include the effects of power supply ripple and spurious signals as well as mechanical or acoustic vibration from fans, motors, and other sources. The effects of vibration can be severe, especially in airborne environments where high vibration levels are present. The vibration sensitivity of an oscillator is specified by the factional frequency vibration sensitivity, commonly know as the *g-sensitivity*. Typically, a single constant value is specified. In practice, the sensitivity varies significantly with vibration frequency and is different for each axis. Equation 6.7 can be used to determine the effect on oscillator phase noise due to random vibration in each axis.[6]

$$L(f_v) = 20\log_{10}\left[\frac{\Gamma_i f_0 \sqrt{\gamma_i(f_v)}}{f_v}\right] \quad \text{dBc SSB in a 1 Hz bandwidth} \qquad (6.7)$$

where f_v = vibration frequency (Hz)
f_0 = oscillator frequency (Hz)
Γ_i = oscillator fractional frequency vibration sensitivity (g^{-1}) in axis i
$\gamma_i(f_v)$ = vibration power spectral density (g^2/Hz) in axis i at the vibration frequency f_v

The composite STALO vibration sensitivity (Γ) is defined by the root sum square of the sensitivity in each of the three prime axes, as shown in Eq. 6.8

$$|\Gamma| = \sqrt{\Gamma_x^2 + \Gamma_y^2 + \Gamma_z^2} \qquad (6.8)$$

Range Dependence. Most modern radars use the STALO in both the receiver for downconversion and the exciter for upconversion. This double use of the STALO 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 return from a specified range.

This range-dependent filter characteristic is given by

$$|F_R(f_m)|^2 = 4\sin^2(2\pi f_m R/c) = 4\sin^2(\pi f_m T) \tag{6.9}$$

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

A short time delay can tolerate much higher disturbances at low modulation frequencies, as illustrated by the two cases in Figure 6.4. Consequently, the effects of STALO stability need to be computed for several time delays or ranges to ensure sufficient stability exists for the intended application.

Close to carrier phase modulation is typically dominated by that of the oscillators due to the inherent feedback process within the oscillator circuitry. Noise contributors within the oscillator loop that exhibit a $1/f$ characteristic (10 dB/decade) noise slope, are enhanced by 20 dB via the feedback mechanism with a resulting net $1/f^3$ characteristic (30 dB/decade) noise signature close to carrier, within the oscillator loop bandwidth. Outside this loop bandwidth, the oscillator noise signature resumes a $1/f$ slope until reaching a flat thermal noise floor. At larger frequency offsets, significant noise contributions can result from other components such as amplifiers in the STALO signal path. Depending on the location of these amplifiers, they may either create phase modulation that is common to both the receiver and exciter (correlated noise) or add phase noise to only the receiver or exciter (uncorrelated noise). Uncorrelated or uncommon noise is not subject to the range dependent factor described above so it must be accounted for separately. Other significant contributors of uncommon noise are the noise on the exciter waveform before upconversion, along with amplifiers in the receiver and exciter signal paths.

The undesired SSB phase noise after downconversion by the STALO is the sum of the uncommon phase noise and the common phase noise reduced by the range factor.

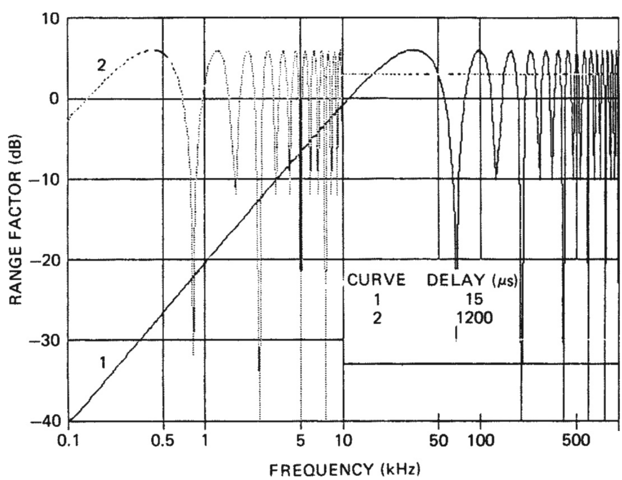

FIGURE 6.4 Effect of range delay on clutter cancellation

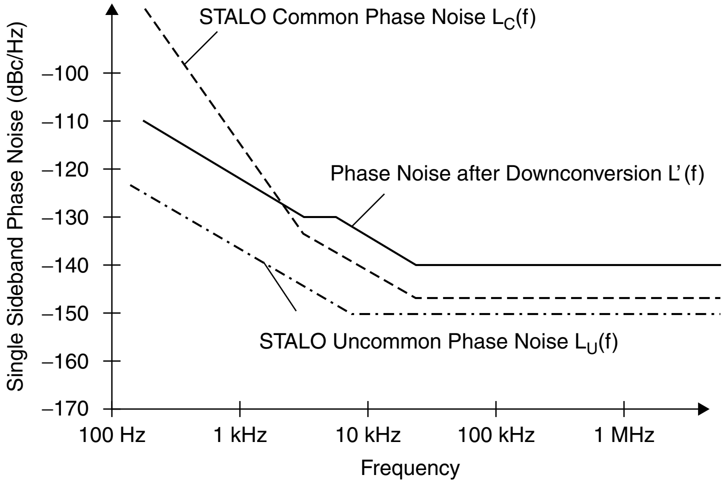

FIGURE 6.5 Phase noise components

Figure 6.5 illustrates typical common and uncommon phase noise components and the resulting mixer output phase noise as calculated using

$$L'(f) = L_C(f)\,|\,F_R(f)\,|^2 + L_U(f) \tag{6.10}$$

where

$L_C(f)$ = STALO SSB phase noise spectrum common to the receiver and exciter
$L_U(f)$ = total receiver-exciter uncorrelated STALO SSB phase noise
$F_R(f)$ = range dependence factor

Residue Power and MTI Improvement Factor. Subsequent stages of the receiver and signal processor have responses that are functions of the doppler modulation frequency, so the output spectrum can be obtained by combining the responses of these filters with the spectrum present at the mixer input. In MTI systems, it is common to describe the ability to suppress clutter in terms of an MTI improvement factor. The MTI improvement factor I is defined as the signal-to-clutter ratio at the output of the clutter filter divided by the signal-to-clutter ratio at the input of the clutter filter, averaged uniformly over all target radial velocities of interest. The MTI improvement factor limitation due to the STALO may be expressed as the ratio of the STALO power to the total integrated power of the return modulation spectrum it creates at the output of the MTI filters. Figure 6.6 illustrates the effect of the overall filtering, consisting of MTI filtering and receiver filtering on the residue power spectrum.

The integrated residue power due to the STAMO phase noise is given by

$$P_{\text{residue}} = \int_{-\infty}^{\infty} |\,H(f)\,|^2\, L'(f)\,df \tag{6.11}$$

where

$H(f)$ = combined response of receiver and doppler filters, normalized to 0 dB noise gain
$L'(f)$ = phase noise after downconversion as defined in Eq. 6.10

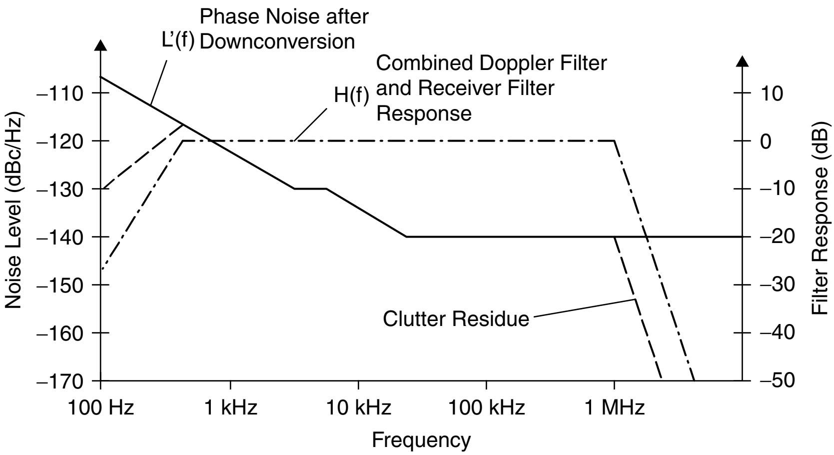

FIGURE 6.6 Clutter residue due to LO phase noise

and the limit on the MTI improvement factor due to the STALO phase noise is given by

$$I = -10\log_{10} P_{\text{residue}} \tag{6.12}$$

If the radar utilizes more than one doppler filter, the effect of STALO instability should be calculated for each individually.

Pulse Doppler Processing. In pulse doppler systems, a series of pulses are transmitted at a fixed pulse repetition frequency (PRF), and doppler processing is performed within the digital signal processor, using samples separated at the PRF rate. The resulting sampling of the receiver output at the PRF produces aliasing of the phase noise spectrum periodically at the PRF interval, as shown in Figure 6.7, where each curve represents the phase noise at the output of the receiver, including the effects of receiver filtering and offset by a multiple of the PRF frequency. The combined phase noise due to each aliased component is calculated using Eq. 6.13 with the result illustrated in Figure 6.8. This sampled phase noise spectrum provides a method for comparing different LO phase noise profiles and their relative impact on the overall performance of the system.

$$\hat{L}(f) = \sum_{k=-\infty}^{\infty} \left[L'(f + kf_{\text{PRF}}) \,|\, H(f + kf_{\text{PRF}}) \,|^2 \right] \tag{6.13}$$

Sinusoidal Modulations. Radar performance is affected by both random and sinusoidal modulations. Sinusoidal modulations can have a significant impact on radar performance, though the degree to which they cause degradation often depends on their relationship to the radar PRF and their magnitude relative to the random modulations. Examples of such undesired sinusoidal modulations are in-band, unfilterable mixer products, or leakage due to insufficient isolation between signal sources within a receiver or exciter. In addition to external sources of interference, the radar designer

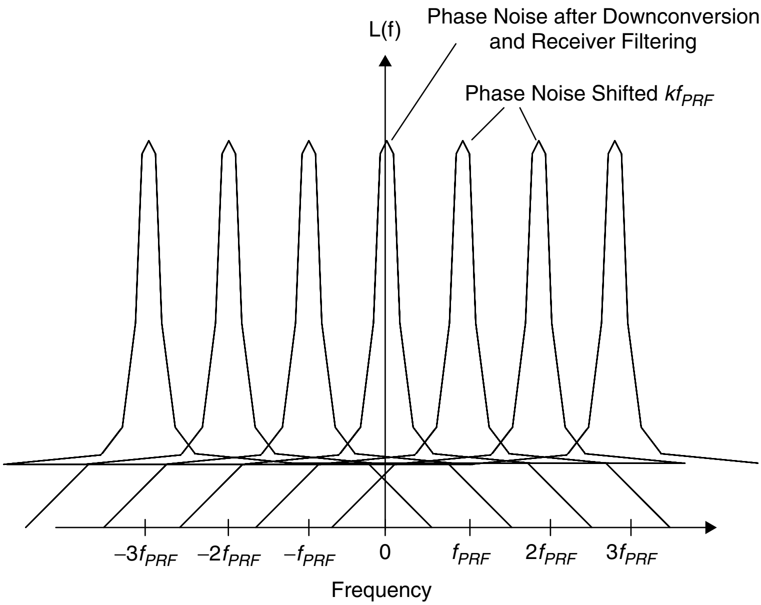

FIGURE 6.7 Phase noise aliasing in a pulse doppler system

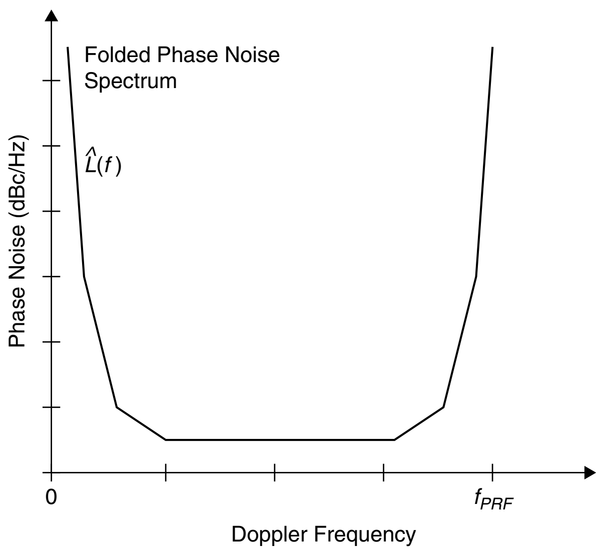

FIGURE 6.8 Sampled phase noise spectrum due to phase noise aliasing

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 is then different for each return, and the ability to reject clutter is degraded.

A truly coherent radar generates all frequencies, including its inter-pulse periods, from a single frequency reference. This fully coherent architecture insures that both the desired frequencies and all the internally generated spurious signals are coherent, eliminating the degradation of clutter rejection.

Many radar systems are pseudo-coherent. The same oscillators are used in both transmit and receive but not necessarily coherent with each other. The result is that the phase of the target remains constant, but the phase of many of the spurious signals varies from pulse to pulse. In this type of configuration, signal isolation and frequency architecture is critical to minimize the occurrence of spurious signals that could erroneously be interpreted as false targets.

COHO and Timing Instability. The majority of this discussion has focused on the STALO as the major contributor to receiver stability. Other contributors such as the second LO, the coherent oscillator (COHO) (if used), A/D and D/A converter clocks can all become significant. A/D and D/A converter clock jitter becomes increasingly significant as sample rates and IF frequencies are increased. The effects of A/D and D/A converter clock phase noise and jitter is described in Sections 6.10 and 6.13. The jitter on timing strobes used to perform transmit/receive (T/R) switching is typically less stringent than that of A/D clocks, as it does not have a direct impact on the signal phase. However, if components such as transmit/receive switches or power amplifiers have a transient phase response of significant duration, time jitter on the switching time can be translated into a phase modulation of the transmitter or receiver signal.

Total Radar Instability. The primary sources of radar instability are usually the receiver-exciter common phase noise, receiver and exciter uncommon phase noise, and the transmitter phase noise. If the spectra of these components are available, either through measurements or through predictions based on similar devices, the convolution of receiver-exciter common phase noise, modified by the range-dependent effect with the other components, provides an estimate of the spectrum of returns from stable clutter, which is then modified by the receiver filters and integrated to obtain the residue power caused by these contributors. 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 and to allow the allocation of stability requirements to critical components or subsystems within the radar.

Measurement of total radar instability can be conducted with the radar antenna search-lighting a stable point clutter reflector that produces a signal return close to (but below) the dynamic-range limit of the receiver. 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 return 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 return, 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.

In addition to the amplitude and phase noise of the receiver-exciter and the transmitter, mechanically scanning antennas produce a modulation that is predominantly AM. The combined effect is the sum of the residue powers produced by each component individually.

Low Noise Frequency Sources. Many radar systems operate over a range of RF frequencies, requiring a number of LO frequencies that are typically generated using frequency synthesis. Frequency synthesis is the process of creating one or more frequencies from a single reference frequency using frequency multiplication, division, addition, and subtraction to synthesize the required frequencies. The fundamental building block of any frequency synthesis approach is the oscillator. Crystal oscillators have historically been the most common source technology. VHF crystal oscillators employing doubly-rotated (SC, IT, etc.) crystal resonators are able to support higher power levels than single axis crystals. This enables them to achieve lower phase noise and improved vibration immunity due to properties unique to the particular axis of rotation. Frequency multiplication of these VHF sources is often used to generate the radar RF frequencies required; however, this multiplication process results in increase in phase noise performance by $20 \log_{10}(M)$ dB where M is the multiplication factor. A variety of other source technologies, such as Surface Acoustic Wave (SAW) oscillators, have been exploited to achieve improved phase noise performance. SAW oscillators enable lower far-from-carrier phase noise, largely due to their higher frequency operation and the resulting lower frequency multiplication factor required to generate the equivalent radar RF output frequencies.

Very accurate frequency timing is often required in radars where coordination or hand-off from one radar to another, or communication to a missile in flight, is required. This is typically the case where a search radar acquires a target and queues a precision tracking radar. Accurate timing for these applications may be achieved by phase locking the low phase noise radar oscillators to a low frequency reference generated from either a rubidium oscillator or a GPS receiver. In this configuration, the long-term stability of the reference oscillator is superior to that of the radar oscillator, and the short-term stability of the radar oscillator is superior to that of the reference oscillator. The phase lock loop (PLL) architecture is established to exploit the strengths of both technologies by selecting a PLL bandwidth at the offset frequency where the source stabilities cross over. For typical radar and reference oscillator technologies, this usually occurs in the 100 Hz to 1 kHz offset region.

Frequency Synthesis Techniques. The most common techniques are direct synthesis, direct digital synthesis, and frequency multiplication. Direct synthesis is the process of generating frequencies through the multiplication and mixing of a number of signals at different frequencies to produce the required output frequency. Frequency multiplication and direct digital synthesis are described in Section 6.13. Conventional phase locked loop synthesizers are occasionally used, but their frequency switching times and phase settling responses are generally inadequate to meet the stringent radar receiver-exciter requirements. Phase locked loops are more likely used to lock fixed high-frequency oscillators to stable low-frequency references to ensure coherence of all oscillators within the receiver-exciter and obtain an optimum balance of long- and short-term stability.

Coherence After Frequency Switching. Long range radars often transmit a series of pulses before receiving returns from the first in the sequence. Pulses may be transmitted at a number of different operating frequencies requiring switching of the LO frequency between pulses. If target returns are processed coherently, the phase of the LO signal must be controlled such that each time it switches to a particular frequency, the phase of the LO is the same phase that it would have been had no frequency switching occurred. This requirement drives the architecture used to generate LO frequencies. Generating all the frequencies from a single reference frequency does not guarantee phase coherence when frequency switching occurs. Three sources of phase ambiguity are common: frequency dividers, direct digital synthesizers, and voltage controlled oscillators (VCO). Frequency dividers produce an output signal that can have any one of N phases, where N is the divide ratio; switching dividers can result in phase ambiguity of $2\pi/N$. If frequency dividers are used in the frequency synthesis process, they must be operated constantly without switching the input frequency or divide ratio to avoid this phase ambiguity. Direct digital synthesizers (DDSs) can be used either to generate LO frequencies directly or to generate modulated waveforms prior to upconversion. When pulse-to-pulse phase coherence is required, the starting phase is reset to zero at the start of each pulse. If all the LO frequencies used are multiples of the pulse repletion frequency, the resulting phase will be the same for each pulse. VCOs can be used to create a tunable LO but are usually phase locked to another stable source for improved stability. The tuning voltage design and filter capacitor technology used to achieve phase lock must be carefully designed to ensure rapid voltage and stored charge transitions. Otherwise, the VCO may properly acquire and achieve phase lock, but the residual voltage decay from the transition will manifest itself in an insidious phase ambiguity called *post-tuning drift*.

Stretch Processing. In stretch processing, the LO signal frequency is modulated with a chirp waveform similar to that of the received signal to reduce the bandwidth of the IF signal as described in Section 6.3. The wideband chirp waveform is typically generated by passing a narrower bandwidth linear frequency modulation (LFM) waveform through a frequency multiplier that increases both the operating frequency and bandwidth of the chirp waveform. Frequency multipliers multiply the phase distortion of the input signal and often have significant phase distortion themselves. Distortion of the LO chirp signal phase can have a significant effect on the compressed pulse performance, either distorting the compressed pulse shape or degrading sidelobe performance (Section 6.13). Phase errors can be measured using a test target injected into the receiver and measuring the phase ripple at the receiver output. By performing this measurement with targets injected at different simulated ranges, the errors associated with the receiver LO and test signal can be separated. Correction of receiver LO phase distortion can be readily corrected when using a direct digital synthesizer as described in Section 6.13.

6.6 GAIN CONTROL

Sensitivity Time Control (STC). The search radar detects returns of widely differing amplitudes, often so great that the dynamic range of a fixed-gain receiver will be exceeded. Differences in return strength are caused by differences in radar cross sections, in meteorological conditions, and in range. The effect of range on radar return strength overshadows the other causes and can be mitigated 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 return strength is independent of range.

Time sidelobes of compressed pulses in radars that transmit coded waveforms can be degraded by STC. Gradual changes can usually be tolerated, but at close range, the rate of change of attenuation can be very large. Most modern radars that include STC use digital STC control, which can lead to large step sizes at close range unless high digitization rates are used. The phase stability of the STC attenuator is also an important consideration as excessive phase variation as a function of attenuation can have a dramatic impact on range sidelobes.

Clutter Map Automatic Gain Control. In some radars, mountain or urban clutter can create returns that would exceed the dynamic range of the receiver. The spatial area occupied by such clutter is typically a very small fraction of the radar coverage, so clutter map AGC has been used as an alternative to boosting the STC curve. This technique uses a digital map to record the mean amplitude of the clutter in each map cell over many scans and adds receiver attenuation where necessary to keep the clutter returns below the saturation level of the receiver.

Programmable Gain Control. Reduced gain may be desirable in a variety of situations such as high clutter or high interference environments or in short range modes. Fixed attenuation is often preferable to STC or clutter map control. High PRF pulse doppler radars, for example, cannot tolerate STC due to the range ambiguity of targets. Additional attenuation may be programmed either manually via operator control or automatically to increase the receiver's large signal handling capability or to reduce its sensitivity.

Gain Normalization. Receiver gain can vary due to component tolerances, frequency response, variation with temperature, and aging. Accurate receiver gain control is required for a variety of reasons that include target radar cross-section measurement, monopulse angle accuracy, maximizing the receiver dynamic range, and noise level control. Digital gain control permits the calibration of receiver gain by injecting test signals during radar dead time or during some scheduled calibration interval. Calibration coefficients can be stored as a function of commanded attenuation, operating frequency, and temperature as needed. Measurements over time can also be used to assess component aging and potentially predict receiver failure prior to degradation beyond acceptable limits. Accurate gain control is essential for receiver channels used to perform monopulse angle measurements, where amplitudes received in two or more beams simultaneously are compared to accurately determine the target's position in azimuth or elevation. Receiver dynamic range is maximized with accurate gain control as too little gain can result in noise figure degradation and too much gain results in large signals exceeding the A/D converter full-scale or creating unwanted gain compression, intermodulation, or cross modulation distortion.

Automatic Noise-Level Control. Another widely employed use for AGC is to maintain a desired level of receiver noise at the A/D converter. As will be described in Section 6.10, too little noise relative to the quantization increment of the A/D converter causes a loss in sensitivity. Samples of noise are taken at long range, often beyond the instrumented range of the radar or during some scheduled period. If the radar has

RF STC prior to any amplification, it can be set to full attenuation to minimize 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. The noise level calibration algorithm must be designed to tolerate external interference and returns from rainstorms or mountains at extreme range.

Another concern with amplification prior to STC is that the noise level at the output of the STC attenuator varies with range. At close range, the noise level into the A/D converter may fall below the quantization interval. Also, a constant noise level as a function of range at the receiver output is desirable in order to maintain a constant false alarm rate. Noise injection after the STC attenuator is used to overcome this problem. A noise source and attenuator are often employed at IF to inject additional noise to compensate for the reduced noise after the STC attenuator. Digital control of the noise injection is synchronized with the STC attenuation to provide an effective constant noise level at the A/D converter input.

Gain Control Components. Most modern radars perform gain control digitally. Digital control permits calibration of each attenuation value to determine the difference between the actual attenuation and that commanded by injecting test signals during dead time.

In the past, gain controlled amplifiers were used extensively to control and adjust receiver gain. Recently, this approach has largely been replaced using digital switched or analog (voltage or current) controlled attenuators distributed throughout the receiver chain. Variable attenuators have a number of advantages over variable gain amplifiers; they typically provide broader bandwidths, greater gain control accuracy, greater phase stability, improved dynamic range, and faster switching speed.

The choice between voltage controlled and switched attenuation depends on tradeoffs between performance of a variety of parameters. Switched attenuators generally provide maximum attenuation accuracy, faster switching speed, improved amplitude and phase stability, greater bandwidth, higher dynamic range, and higher power handling capability. Voltage or current controlled attenuators, controlled via a D/A converter, typically provide improved resolution and lower insertion loss.

Gain control attenuators are often incorporated within the receiver at both RF and IF. RF attenuation is used to provide increased dynamic range in the presence of large target returns. By placing the attenuation as close to the front end as possible, large signals can be handled by minimizing gain compression, intermodulation, or cross-modulation distortion in the majority of receiver components. The disadvantage of using front-end attenuation is that it will typically have a larger impact on receiver noise figure than attenuation placed later in the receiver. This is not usually an issue when the intent of adding attenuation is to desensitize the receiver as is the case for STC. Back-end or IF attenuation is often used to adjust the gain of the receiver to compensate for receiver gain variations due to component variations where receiver noise figure degradation cannot be tolerated.

6.7 FILTERING

Filtering of the Entire Radar System. Filtering provides the principal means by which the radar discriminates between target returns and interference of many types. The filtering is performed by a variety of filters throughout the receiver and in the subsequent digital signal processing. Most radars transmit multiple pulses at a target before the antenna beam is moved to a different direction, and the multiple returns are combined in some fashion. The returns may be combined using coherent integration or various doppler processing techniques (including MTI) to separate desired targets from clutter. From the radar system standpoint, these are all filtering functions, and in modern radar systems, these functions are performed using digital signal processing on the receiver output I and Q data. These functions are discussed in other chapters of this handbook. The purpose of the filtering within the receiver is to reject out-of-band interference and digitize the received signal with the minimum of error so that optimum filtering can be performed using digital signal processing.

Matched Filtering. Although matched filtering is typically now performed within the digital signal-processing function, the concept is explained here for completeness. The overall filter response of the system is chosen to maximize the radar performance. If the signal spectrum $X(\omega)$, in the presence of white noise with power spectral density $N_0/2$, is processed with a filter with frequency response $H(\omega)$, the resulting signal-to-noise ratio (SNR) at time T is given by[7]

$$\chi = \frac{\left| \frac{1}{2\pi} \int_{-\infty}^{\infty} X(\omega) H(\omega) e^{j\omega T} \, d\omega \right|^2}{\frac{N_0}{4\pi} \int_{-\infty}^{\infty} |H(\omega)|^2 \, d\omega} \tag{6.14}$$

The ideal filter response from the standpoint of maximizing SNR is the matched filter that maximizes the SNR at time T_M when

$$H_M(\omega) = X^*(\omega) e^{-j\omega T_M} \tag{6.15}$$

Deviations from the ideal matched filter response $H_M(\omega)$ produce a reduction in SNR termed *mismatch loss*. This loss can occur for a number of reasons such as target doppler or because a filter response is chosen that is different from the matched filter response in order to minimize another parameter such as range sidelobes.

Receiver filtering is often modified for different waveforms used. When radar systems use waveforms of widely varying bandwidths, different I/Q data rates may be used to minimize the digital signal-processing throughput requirements. With different data rates comes the need to adjust the receiver filtering in order to avoid aliasing signals beyond the Nyquist rate. Although these radars adjust their filtering to the waveform bandwidth, they do not typically implement the matched filtering within the receiver. This function is usually implemented in digital signal processing.

Receiver Filtering. Filtering is required at various points throughout the receiver chain including RF, IF, baseband if used, digital filtering prior to decimation (reduction of the sample rate), and as an integral part of I/Q generation.

Section 6.4 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 signal 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 primary function of RF filtering is the rejection of the image response due to the first downconversion. Image rejection filtering can be alleviated using an image reject mixer; however, the maximum rejection achievable by image reject mixers is typically inadequate without the use of additional rejection through filtering. This image-suppression problem is the reason why some receivers do not translate from the received signal 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 (Figure 6.2) shows that there are certain choices of frequency ratio that provide spurious-free frequency bands, approximately 10% 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. RF filtering is also important as it reduces out-of-band interference before it can cause intermodulation or cross-modulation distortion within the receiver.

If the receiver operating bandwidth is a large percentage of the RF frequency, some form of switched or tunable RF filtering may be required so that the image response is rejected as it moves through the operating bandwidth. The choice between using switched or tunable filtering depends on the switching speed, linearity, and stability requirements of the receiver. Switched filters provide the fastest response time, with excellent linearity and stability but can be bulky and suffer from the additional loss of the switch components.

An alternate approach that is sometimes used with large operating bandwidths is to first upconvert the input RF signal to an IF frequency higher than the RF operating band. This process virtually eliminates the image response problem, allowing the use of a single RF filter spanning the entire operating bandwidth. Narrow bandwidth filtering can be used on the high IF as defined by the signal bandwidth before downconversion to a lower IF for digitization or baseband conversion.

IF filtering is the primary filtering used to define the receiver bandwidth prior to A/D conversion in receivers using either IF sampling or baseband conversion. In IF sampling receivers, the IF filter acts as the anti-aliasing filter and limits the bandwidth of signals entering the A/D converter. In receivers using baseband conversion, the IF filter sets the receiver bandwidth. Subsequent video filtering should be of greater bandwidth to prevent the introduction of I/Q imbalance due to filter differences between I and Q channels.

In IF sampling receivers, digital filtering is usually the primary means of setting the final receiver bandwidth and provides anti-alias rejection required to prevent aliasing in the decimation of the I/Q data rate. Digital filtering can be precisely controlled, tailored to almost any desired passband and stop band rejection requirements. The digital filters used are typically linear phase FIR filters, but they can also be tailored to compensate for variations in the passband phase and amplitude responses of RF and IF analog filters.

Filter Characteristics. Filter responses are characterized fully by either their frequency response $H(\omega)$ or their impulse response $h(t)$; however, they are usually specified by a variety of parameters as described below. Digital filters may be specified using the same measures, or because they can be specified exactly, they are frequently specified by their transfer function $H(z)$ or impulse response $h(n)$.

Key passband characteristics are: insertion loss, bandwidth, passband amplitude and phase ripple, and group delay. Bandwidths are frequently specified in terms of a 3 dB bandwidth; however, if a low passband variation is required, the specified bandwidth may be, for example, specified as a 0.5 dB or 0.1 dB bandwidth. Passband amplitude variation relative to the insertion loss is a key parameter that has potential impact on range sidelobes and channel-to-channel tracking. Phase ripple, if specified, is relative to a best-fit linear phase and has similar effects as amplitude ripple. Group delay, the rate of change of phase vs. frequency, is ideally constant for linear phase filters. The absolute value of group delay does not impact the range sidelobe performance; however, the relative group delay between channels must be tightly controlled or compensated in monopulse, sidelobe canceler, and digital beamforming systems.

Although stopband rejection is clearly a key parameter, filters with fast roll-off may not provide the required phase and impulse response characteristics. Figure 6.9 shows the magnitude response of six different fifth order low-pass filters with equal 3 dB bandwidth.[8] The Chebshev filters (0.1 and 0.01 dB ripple) have flat passband response and improved stopband rejection relative to the remaining filters; however, as shown in Figure 6.10 and Figure 6.11, they have inferior phase (group delay) and impulse response characteristics.

Digital filters can be either Finite Impulse Response (FIR) or Infinite Impulse Response (IIR). FIR filters are typically preferred as their finite response is desirable along with their linear phase characteristic. Phase linearity is achieved with the symmetric impulse response condition[9] defined by Eq. 6.16 or the anti-symmetric impulse response conditions defined by Eq. 6.17:

$$h(n) = h(M-1-n) \qquad n = 0, 1, \ldots, M-1 \tag{6.16}$$

where M is the length of the FIR filter impulse response

$$h(n) = -h(M-1-n) \qquad n = 0, 1, \ldots, M-1 \tag{6.17}$$

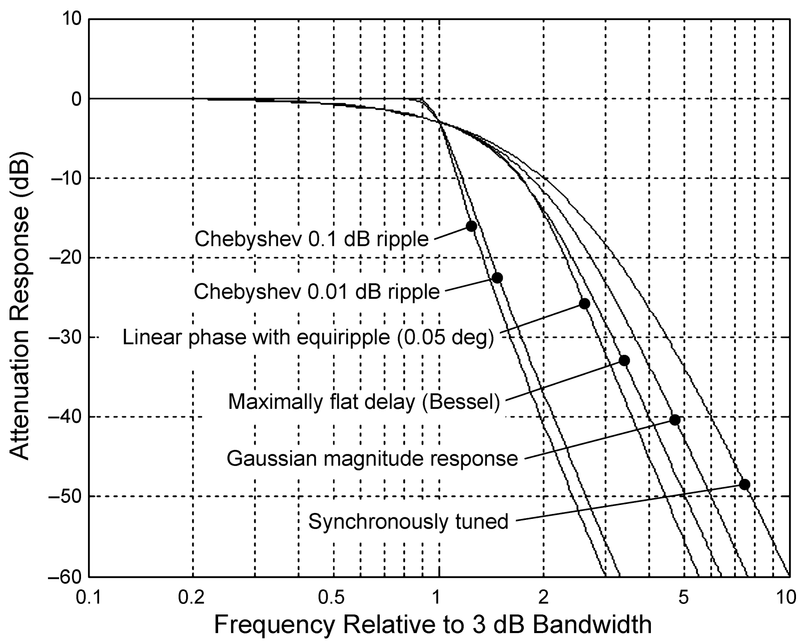

FIGURE 6.9 Magnitude response of lowpass filters

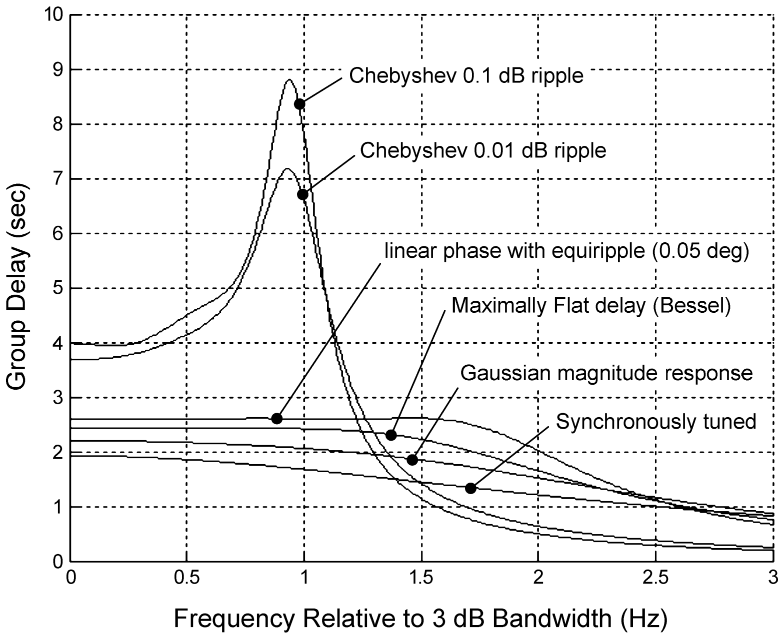

FIGURE 6.10 Group delay response of lowpass filters

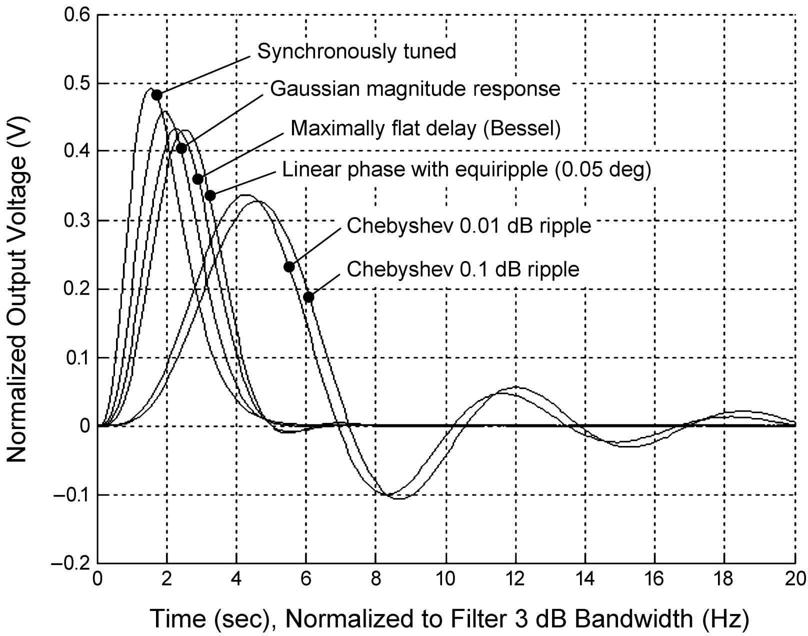

FIGURE 6.11 Normalized impulse response of lowpass filters

Range Sidelobes. Errors in filter responses can produce degradation in pulse compression range sidelobes. The effect of a filter response on range or time sidelobes can be seen by taking the filter impulse response $h(t)$ and adding to this a delayed impulse response $20\log_{10}(\alpha)$ dB below the main response to produce the modified response $h'(t)$, which is given by

$$h'(t) = h(t) + \alpha h(t-T_0) \tag{6.18}$$

Using the property of time shifting of the Fourier transform, the resultant frequency response is given by

$$H'(\omega) = H(\omega) + \alpha e^{-j\omega T_0} H(\omega) \tag{6.19}$$

Thus, for small values of α, the resulting magnitude and phase response is that of the original filter modified by a sinusoidal phase and amplitude modulation as given here:

$$|H'(\omega)| = |H(\omega)|(1 + \alpha\cos(\omega T_0)) \tag{6.20}$$

$$\angle H'(\omega) = \angle H(\omega) - \alpha\sin(\omega T_0) \tag{6.21}$$

Therefore, if there are n ripples across the filter bandwidth B, the range sidelobe occurs at time T_0 given by

$$T_0 = n/B \tag{6.22}$$

Assuming a compressed pulse width of 1/B, values of $n < 1$ will put the range sidelobe within the main lobe of the target return, resulting in a distortion of the mainlobe response.

Channel Matching Requirements. Radar receivers with more than one receiver channel typically require some degree of phase and amplitude matching or tracking between channels. In order to operate effectively, sidelobe canceler channels must track very closely. Constant offsets in gain or phase do not degrade sidelobe canceler performance, but small variations in phase and amplitude across the bandwidth cause significant degradation. For example, achieving a cancellation ratio of 40 dB requires a gain tracking of less than 0.1 dB across the receiver bandwidth. Filters are the main source of amplitude and phase ripple across the signal bandwidth as other components such as amplifiers and mixers are typically relatively broadband. The degree of tracking required for sidelobe canceler operation was previously achieved by providing matched sets of filters with tightly tracking amplitude and phase responses. Modern digital signal processing allows the correction of these channel-to-channel variations using FIR equalization (Section 6.11) or correction in the frequency domain in the digital signal processor, allowing the use of less tightly controlled filters.

6.8 LIMITERS

Applications. Limiters are used to protect the receiver from damage and to control saturation that may occur within the receiver. When received signals saturate some stage of the radar receiver that is not expressly designed to cope with such a situation, the distortions can result in severely degraded radar performance, and the distortion 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. Limiting prior to the A/D converter also prevents the distortion that occurs when signals exceed full-scale. Although A/D converters can often handle modest overload with fast recovery, the distortion that occurs degrades signal processing such as digital pulse compression and clutter rejection. With IF limiting, these harmonics are filtered out using bandpass filtering after limiting prior to A/D conversion, minimizing the degradation due to limiting.

All radar systems contain some form of Transmit/Receive (T/R) device to protect the receive electronics from the high-power transmit signal. In many systems, an RF front-end limiter is also required in order to prevent the receiver from being damaged by high input power levels from the antenna that may occur as a result of leakage from the T/R device during transmit mode or from interference due to jammers or other radar systems. These limiters are typically designed to limit well above the maximum signals to be processed by the receiver.

In the past, limiters were used to perform a variety of analog signal-processing functions. Hard limiters with as much as 80 dB of limiting range were used with some designed to limit on-receiver noise. Applications that utilize hard limiting, including phase-detectors and phase-monopulse receivers, are described in Section 3.10 of the second edition[1] of this handbook. Modern radar systems are mostly designed to maximize the linear operating region, with limiters used only to handle excessively large signals that inevitably exist under worst case conditions.

Characteristics. The ideal limiter is perfectly linear up to the power level at which limiting begins followed by a transition region beyond which the output power remains constant. In addition, the insertion phase is constant for all input power levels, and recovery from limiting is instantaneous. The output waveform from a bandpass limiter is sinusoidal, whereas the output waveform from a broadband limiter approaches a square wave. Deviations from the ideal characteristics can degrade radar performance in a variety of ways.

Linearity Below Limiting. One major drawback of adding a limiter stage to a receiver channel is that it is inherently nonlinear. Since any practical limiter has a gradual transition into limiting, the limiter is often the largest contributor to receiver channel nonlinearity in the linear operating region and can cause significant intermodulation distortion of in-band signals. For this reason, the primary limiting stage is usually located at the final IF stage where maximum filtering of out-of-band interference has been achieved. The lower operating frequency also allows implementation of a limiter that more closely matches the ideal characteristics.

Limiting Amplitude Uniformity. No single-stage limiter will exhibit a constant output over a wide range of input signal amplitudes. One cause is apparent if one considers the effect of a single-stage limiter having a perfectly symmetrical clipping at voltages $\pm E$. For a sinusoidal input, the output signal at the threshold of limiting is

$$v_0 = E\sin(\omega t) \tag{6.23}$$

and when the limiter is fully saturated and the output waveform is rectangular, it is given by the Fourier series:

$$v_o' = \frac{4E}{\pi} \sum_{n=1,3,5,\ldots}^{\infty} \frac{1}{n} \sin(n\omega t) \tag{6.24}$$

which is an increase of 20 log $(4/\pi) = 2.1$ dB in the power of the fundamental.

In practice, the amplitude performance is also degraded by capacitive coupling between input and output of each limiting stage, charge storage in transistors and diodes, and *RC* time constants that permit changes in bias with signal level. For these reasons, two or more limiter stages may be cascaded when good amplitude uniformity is required over a wide dynamic range.

Phase Uniformity. The change of insertion phase of the limiter with amplitude is less of a concern for modern radar systems that operate primarily in the linear operating region. However, maintaining constant insertion phase during limiting preserves the phase of target returns in the presence of limiting clutter or interference. The change of insertion phase with signal amplitude is generally directly proportional to the frequency at which it is operated.

Recovery Time. The recovery time of a limiter is a measure of how quickly the limiter returns to linear operation after the limiting signal is removed. Fast recovery is particularly important when the radar is exposed to impulsive interference.

6.9 I/Q DEMODULATORS

Applications. The I/Q demodulator, also referred to as a quadrature channel receiver, quadrature detector, synchronous detector, or coherent detector, performs frequency conversion of signals at the IF frequency to a complex representation, $I + jQ$ centered at zero frequency. The baseband in-phase (I) and quadrature-phase (Q) signals are digitized using a pair of A/D converters providing a representation of the IF signal, including phase and amplitude without loss of information. The resulting digital data can then be processed using a wide variety of digital signal-processing algorithms, depending on the type of radar and mode of operation. Processing such as pulse compression, doppler processing, and monopulse comparison, all require amplitude and phase information. The predominance of digital signal processing in modern radar systems has led to almost universal need for Nyquist rate sampled data. In many modern radar systems, digital I and Q data is now generated using IF sampling followed by digital signal processing used to perform the baseband conversion as described in Sections 6.10 and 6.11. I/Q demodulators are still used, though their use is increasingly limited to wider bandwidth systems where A/D converters are not yet available with the required combination of bandwidth and dynamic range to perform IF sampling.

Implementation. Figure 6.12 shows the basic block diagram of a I/Q demodulator. The IF signal described by Eq. 6.25 is split and fed to a pair of mixers or analog multipliers. The mixer LO ports are fed with a pair of signals in quadrature, generated from the reference frequency signal, or coherent oscillator (COHO), and represented

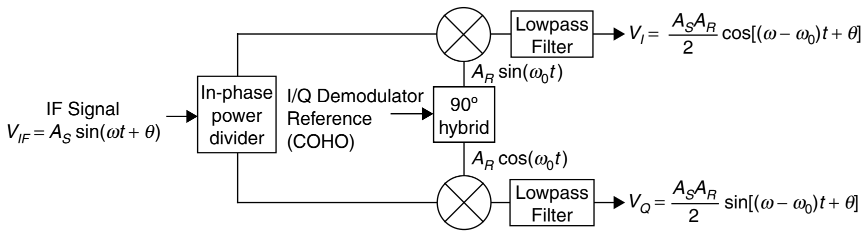

FIGURE 6.12 I/Q demodulator

in complex form in Eq. 6.26. Ignoring any mixer insertion loss or loss associated with the IF split, the complex representation of the mixer output is given by Eq. 6.27. Ideal low-pass filtering rejects the second (sum frequency) term of Eq. 6.27, producing the I/Q demodulator output as represented by Eq. 6.28.

$$V_{\text{IF}} = A_S \sin(\omega t + \theta) = \frac{A_S}{2j}\left(e^{j(\omega t+\theta)} - e^{-j(\omega t+\theta)}\right) \tag{6.25}$$

$$V_{\text{COHO}} = A_R\left[\sin(\omega_0 t) + j\cos(\omega_0 t)\right] = jA_R e^{-j\omega_0 t} \tag{6.26}$$

$$V_{\text{IF}} V_{\text{COHO}} = \frac{A_S}{2}\left(e^{j(\omega t+\theta)} - e^{-j(\omega t+\theta)}\right)A_R e^{-j\omega_0 t} = \frac{A_S A_R}{2} e^{j[(\omega-\omega_0)t+\theta]} - \frac{A_S A_R}{2} e^{-j[(\omega+\omega_0)t+\theta]} \tag{6.27}$$

$$V_I + jV_Q = \frac{A_S A_R}{2}\cos[(\omega-\omega_0)t+\theta] + j\frac{A_S A_R}{2}\sin[(\omega-\omega_0)t+\theta] = \frac{A_S A_R}{2} e^{j[(\omega-\omega_0)t+\theta]} \tag{6.28}$$

In implementing an I/Q demodulator, it is important to provide well-balanced *I* and *Q* channels in order to maximize image rejection, as explained below. The mixers must have DC coupled IF output ports and be presented with a good match at both the wanted low frequency output and the unwanted sum frequency. A match at the sum frequency can be provided using a diplexer filter. Video filtering is required to reject the sum frequency mixer outputs and also provides rejection of wideband noise from the video amplifiers, which would otherwise alias to baseband through the A/D converter sampling process, producing an unwanted degradation of receiver noise figure. Video amplification is often required to increase the signal level to the full-scale signal level of the A/D converter and also allows for impedance matching of the mixer and A/D converter.

The convention for the *I* and *Q* relationship is that the *I* signal phase leads the *Q* signal phase for radar signals with positive doppler (approaching targets). Frequency conversions within the receiver using LO frequencies greater than the RF frequency will cause a doppler frequency inversion, so each conversion must be considered in order to achieve the correct sense of *I* and *Q* at the receiver output. Fortunately, an incorrect *I* and *Q* relationship can easily be fixed either in the receiver or the signal processor, by switching the *I* and *Q* digital data or by changing the sign of either *I* or *Q*.

Gain or Phase Imbalance. If the gains of the *I* and *Q* channels are not exactly equal or if their COHO phase references are not exactly 90 degrees apart, an input signal at frequency ω will create an output at both the desired frequency $\omega - \omega_0$ and at the image frequency $-(\omega - \omega_0)$. The image signals generated by gain and phase imbalance are given by Eq. 6.29 and Eq. 6.30. For small errors, if the ratio of voltage gains is $(1 \pm \Delta)$ or if the phase references differ by $(\pi/2 \pm \Delta)$ radians, the ratio of the spurious image at $-\omega_d$ to the desired output of ω_d is $\Delta/2$ in voltage, $\Delta^2/4$ in power, or 20 (log Δ) −6 in decibels.

$$V_I + jV_Q = E\cos(\omega_d t) + j(1+\Delta)E\sin(\omega_d t) = \left(1+\frac{\Delta}{2}\right)Ee^{j\omega_d t} - \frac{\Delta}{2}Ee^{-j\omega_d t} \tag{6.29}$$

$$V_I + jV_Q = E\cos(\omega_d t) + jE\sin(\omega_d t + \Delta) = \cos\left(\frac{\Delta}{2}\right)Ee^{j\left(\omega_d t + \frac{\Delta}{2}\right)} - \sin\left(\frac{\Delta}{2}\right)Ee^{-j\left(\omega_d t + \frac{\Delta+\pi}{2}\right)} \tag{6.30}$$

Historically, *I* and *Q* phase and gain corrections have been performed using adjustments in the analog signal paths, as shown in Figure 6.13. Gain errors may be corrected by a change in gain in the IF or video stages of either or both *I* and *Q* channels. Video gain control must be implemented with care as it can exaggerate the nonlinearity of those stages. These corrections can now be implemented more precisely in the digital domain.

A measurement of the signal spectrum at the center of the IF bandwidth indicates the degree of gain and phase imbalance compensation. However, as the following discussion will explain, the suppression of image energy across the IF bandwidth may be substantially less than indicated by this measurement at IF center.

Time Delay and Frequency Response Imbalance. If the responses of the *I* and *Q* channels are not identical across the entire signal bandwidth, unwanted image responses will occur that are frequency dependent. Optimum bandpass filtering should

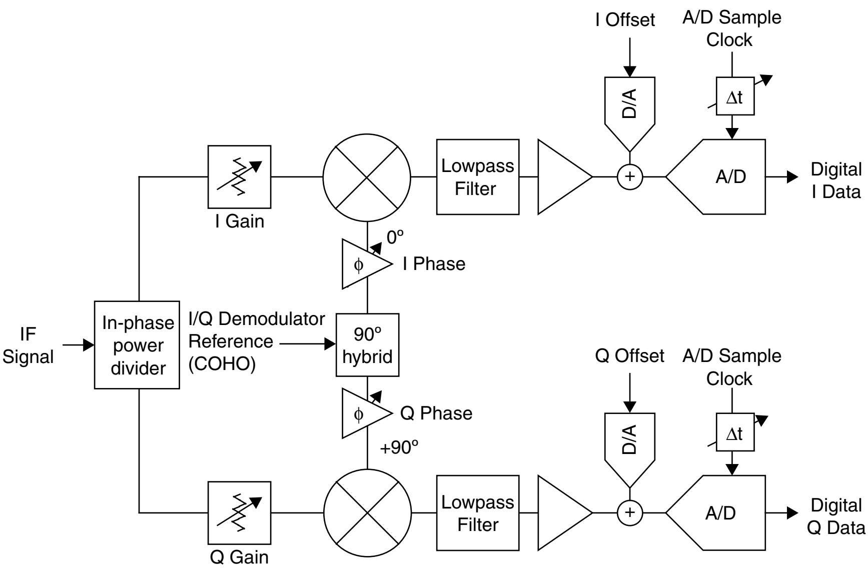

FIGURE 6.13 I/Q demodulator with gain, phase, DC offset, and time-delay adjustments

be at IF, where it affects I and Q channels identically, not at baseband. Video filter bandwidth should be more than half the IF bandwidth and controlled by precision components in order to minimize the creation of image signals. Substituting $\Delta(\omega)$ for Δ in Eq. 6.29 and Eq. 6.30 gives the image components for frequency dependent gain and phase errors. Similarly, substituting $\omega\Delta T$ for Δ in Eq. 6.30 gives the image component due to time-delay imbalance in the I and Q paths. Small time-delay imbalances can be corrected by adding time delay to the A/D sample clock, as shown in Figure 6.13. Large time-delay corrections should be avoided as they can cause problems aligning the I and Q digital data. When adding time delay to the sample clock, care must be taken to avoid adding jitter, which could degrade the A/D converter SNR performance. Time-delay correction can also be implemented effectively in the digital domain, and if frequency dependent phase and amplitude imbalance correction is required, this is most easily and effectively performed in the digital domain using FIR filtering of the I and Q data or by performing corrections in the frequency domain data as part of the radar signal processing.

Nonlinearity in *I* and *Q* Channels. Component tolerances often lead to somewhat different nonlinearities in I and Q, which can generate the variety of spurious doppler components.

The ideal input signal is

$$V = Ae^{j\omega_d t} = I + jQ \tag{6.31}$$

Each video channel response can be expressed as a power series. For simplicity, only symmetrical distortion will be considered. The A/D output, including a residual gain imbalance of Δ, is

$$V'_{\mathrm{IQ}} = V'_I + jV'_Q \tag{6.32}$$

$$V'_I = V_I - aV_I^3 - cV_I^5 \tag{6.33}$$

$$V'_Q = (1+\Delta)V_Q - bV_Q^3 - dV_Q^5 \tag{6.34}$$

Substitution of Eqs. 6.33 and 6.34 into Eq. 6.31 yields the amplitudes of the spectral components listed in Table 6.2. Note that if the nonlinearities in I and Q were identical ($a = b$; $c = d$), spurious components at -5ω and $+3\omega$ would not be present and the image ($-\omega$) would be proportional to input signal amplitude. Spurious at zero doppler is not due to dc offset; it is the result of even-order nonlinearities that were omitted from the above equations. The negative third harmonic is the dominant component produced by nonlinearity.

TABLE 6.2 Spurious Signal Components Generated by *I*/*Q* Nonlinearity

Signal Frequency	Amplitude of Spectral Component
-5ω	$A^5(c-d)/32$
-3ω	$A^3(a+b)/8 + 5A^5(c+d)/32$
$-\omega$	$A(\Delta/2) + 3A^2(a-b)/8 + 5A^5(c-d)/16$
(Input) ω	$A(1+\Delta/2) - 3A^2(a+b)/8 - 5A^5(c+d)/16$
$+3\omega$	$A^3(a-b)/8 + 5A^5(c-d)/32$
$+5\omega$	$A^5(c+d)/32$

DC Offset. Small signals and receiver noise can be distorted by an offset in the mean value of the A/D converter output unless the doppler filter suppresses this component.

False-alarm control in receivers without doppler filters is sometimes degraded by errors of a small fraction of the least significant bit (LSB), so correction is preferably applied at the analog input to the A/D. DC offsets can be measured using digital processing of the A/D converter outputs and a correction applied using D/A converters, as shown in Figure 6.13. DC offset correction can also be performed effectively in the digital domain, provided that the DC offset at the input of the A/D converter is not so large that it results in a significant loss of available dynamic range.

Many of the I/Q demodulator errors described above are either reduced dramatically or eliminated using IF sampling. This, along with the reduction of hardware required, are the reasons that IF sampling (described in Sections 6.10 and 6.11) is becoming the dominant approach.

6.10 ANALOG-TO-DIGITAL CONVERTERS

The high-speed A/D converter is a key component in receivers of modern radar systems. The extensive use of digital signal processing of radar data has resulted in a demand for converters with both state-of-the-art sampling rates and dynamic range.

Analog to digital converters transform continuous time analog signals into discrete time digital signals. The process includes both sampling in the time domain, converting from continuous time to discrete time signals and quantization, converting from continuous analog voltages to discrete fixed-length digital words. Both the sampling and quantization process produce errors that must be minimized in order to limit the radar performance degradation. In addition, a variety of other errors such as additive noise, sampling jitter, and deviation from the ideal quantization, result in non-ideal A/D conversion.

Applications. The conventional approach of using a pair of converters to digitize the *I* and *Q* outputs of an I/Q demodulator is, in many cases, being replaced by digital receiver architectures where a single A/D converter is followed by digital signal processing to generate *I* and *Q* data. Digital receiver techniques are described in Section 6.11.

Although the dividing line is arbitrary and advancing with the state-of-the-art, radar receivers are often classified as either wideband or high dynamic range. Different radar functions put a greater emphasis on one or the other of these parameters. For example, imaging radars put a premium on wide bandwidth, whereas pulse doppler radars require high dynamic range. Because radars are often required to operate in a variety of modes with differing bandwidth and dynamic range requirements, it is not uncommon to use different types of A/D converter, sampling at different rates for these different modes.

Data Formats. The most frequently used digital formats for A/D converters are 2's complement and offset binary.[10]

The 2's complement is the most popular method of digital representation of signed integers and is calculated by complementing every bit of a given number and adding one.

The most significant bit is referred to as the sign bit. If the sign bit is 0, the value is positive; if it is 1, the value is negative. The representation of voltage in 2's complement form is given by

$$E = k(-b_N 2^{N-1} + b_{N-1}2^{N-2} + b_{N-2}2^{N-3} + \cdots + b_1 2^0) \tag{3.36}$$

where E = analog voltage
N = number of binary digits
b_i = state of ith binary digit
k = quantization voltage

Offset binary is an alternate coding scheme in which the most negative value is represented by all zeros and the most positive value is represented by all ones. Zero is represented by a most significant bit (MSB) of one followed by all zeros. The representation of voltage in offset binary is given by

$$E = k[(b_{N-1})2^{N-1} + b_{N-1}2^{N-2} + b_{N-2}2^{N-3} + \cdots + b_1 2^0] \tag{3.35}$$

The Gray code[10] is also used in certain high-speed A/D converters in order to reduce the impact of digital output transitions on the performance of the A/D converter. The Gray code allows all adjacent transitions to be accomplished by the change of a single digit only.

Delta-Sigma Converters. Delta-sigma converters differ from conventional Nyquist rate converters by combining oversampling with noise-shaping techniques to achieve improved SNR in the bandwidth of interest. Noise shaping may be either lowpass or bandpass depending on the application. Delta-sigma architectures provide potential improvements in spurious-free dynamic range (SFDR) and SNR over conventional Nyquist converters where tight tolerances are required to achieve very low spurious performance. Digital filtering and decimation is required to produce data rates that can be handled by conventional processors. This function is either performed as an integral part of the A/D converter function or can be integrated into the digital downconversion function used to generate digital I and Q data, as described in Section 6.11.

Performance Characteristics. The primary performance characteristics of A/D converters are the sample rate or usable bandwidth and resolution, the range over which the signals can be accurately digitized. The resolution is limited by both noise and distortion and can be described by a variety of parameters.

Sample Rate. Sampling of band-limited signals is performed without aliasing distortion, provided that the sample rate (f_s) is greater than twice the signal bandwidth and provided the signal bandwidth does not straddle the Nyquist frequency ($f_s/2$) or any integer multiple ($Nf_s/2$).

In conventional baseband approaches, sampling is usually performed at the minimum rate to meet the Nyquist criteria. Since the baseband I and Q signals have bandwidths ($B/2$) equal to half the IF signal bandwidth, a sample rate just greater than the IF bandwidth is required (see Figure 6.14).

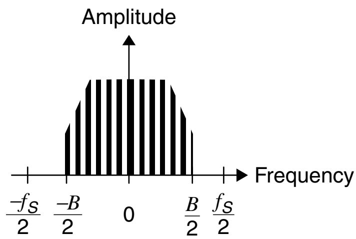

FIGURE 6.14 Baseband sampling

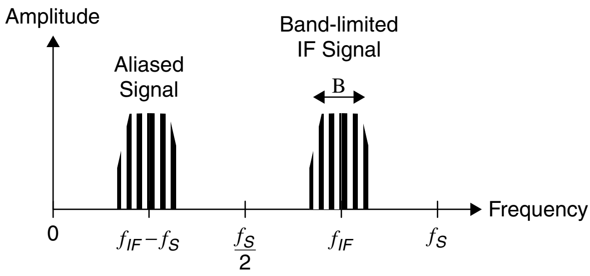

FIGURE 6.15 IF sampling in second Nyquist region

For IF sampling, a frequency at least twice the IF bandwidth is required; however, oversampling is typically employed to ease alias rejection filtering and to reduce the effect of A/D converter quantization noise. IF sampling is often performed with the signal located in the second Nyquist region, as shown in Figure 6.15 or in higher Nyquist regions.

Stated Resolution. The stated resolution of an A/D converter is the number of output data bits per sample. The full-scale voltage range of a Nyquist rate converter is given by $V_{FS} = 2^N Q$, where N is the stated resolution and Q is the least significant bit (LSB) size.

Signal-to-Noise-Ratio (SNR). SNR is the ratio of rms signal amplitude to rms A/D converter noise power. For an ideal A/D converter, the only error is due to quantization. Provided that the input signal is sufficiently large relative to the quantization size and uncorrelated to the sampling signal, the quantization error is essentially random and is assumed to be white. The rms quantization noise is $Q/\sqrt{12}$, and signal-to-quantization-noise ratio (SQNR) of an ideal A/D converter is given by

$$SQNR(dB) = 6.02N + 1.76 \tag{6.37}$$

Practical A/D converters have additional sampling errors other than quantization, including thermal noise and aperture jitter. Provided that these additional errors can be characterized as white, they can be combined with the quantization noise with a resulting SNR less than the theoretical SNR of the ideal converter. Because various A/D converter error mechanisms are dependent on input signal level and frequency, it is important to characterize devices over the full range of input conditions to be expected. The available signal-to-noise ratio of state-of-the-art high-speed A/D converters has been shown[11] to fall off by one-bit (6 dB) for every doubling of the sample rate. Over-sampling of the signal followed by filtering and decimation provides an improvement of one half-bit (3 dB) in the achievable signal-to-noise-ratio for each doubling of the sample rate. Thus, for high dynamic-range applications, the best performance is achieved using a state-of-the-art A/D converter that has a maximum sample rate just sufficient for the application.

Spurious Free Dynamic Range (SFDR). SFDR is the ratio of the single-tone signal amplitude to the largest spurious signal amplitude and is usually stated in dB. Similar to SNR, the spurious performance of an A/D converter is dependent on the input signal frequency and amplitude. The frequency of spurious signals is also dependent on the input signal frequency with the highest values typically due to low order harmonics or their aliases. When using IF sampling with a significant over-sampling ratio ($f_s \gg B/2$), the worst spurious signals may be avoided by choosing the sample frequency relative to signal frequency such that the unwanted spurious signals fall outside the signal bandwidth of interest. If the worst case spurious can be avoided, the specified SFDR is less important than the levels of the specific spurious components that fall within the bandwidth of interest. Again, it is important to characterize devices over the range of expected operating conditions.

The impact of A/D converter spurious signals on radar performance depends on the type of waveforms being processed and the digital signal processing being performed. In applications using chirp waveforms with large time-bandwidth products, spurious signals are less critical as they are effectively rejected in the pulse compression process because their coding does not match that of the wanted signal. In pulse doppler applications, spurious signals are of much greater concern because they can create components with doppler at a variety of frequencies that may not be rejected by the clutter filtering.

Signal-to-Noise-and-Distortion Ratio (SINAD). SINAD is the rms signal amplitude to the rms value of the A/D converter noise plus distortion. The noise plus distortion includes all spectral components, excluding DC and the fundamental up to the Nyquist frequency. SINAD is a useful figure of merit for A/D converters, but in digital receiver applications, where the worst spurious components may fall outside of the bandwidth of interest, it is not necessarily a key discriminator between competing converters for a specific application.

Effective Number of Bits (ENOB). The term *effective number of bits* is often used to state the true performance of an A/D converter and has been stated in the literature[11] in terms of SINAD and SNR, as given below. Consequently, it is important to differentiate between definitions when using this term.

$$N_{\mathrm{eff}} = [SINAD(dB) - 1.76]/6.02 \tag{6.38}$$

$$N_{\mathrm{eff}} = [SNR(dB) - 1.76]/6.02 \tag{6.39}$$

Two Tone Intermodulation Distortion (IMD). Two tone intermodulation distortion is also important in receiver applications. Testing is performed with two sinusoidal input signals of unequal frequency and levels set such that the sum of the two inputs does not exceed the A/D converter full-scale level. Similar to IMD for amplifiers, the most significant distortion is usually second order or third order IMD products. However, due to the complex nature of the distortion mechanism in A/D converters, the amplitude of IMD products is not easily characterized and predicted by the measurement of an input intercept point.

Input Noise Level and Dynamic Range. Accurate setting of the A/D converter input noise level relative to the A/D converter noise is critical to achieving the optimum trade-off between dynamic range and system noise floor. Too high a level of noise into the A/D converter will degrade the available dynamic range; too low a level will degrade the overall system noise floor. Sufficient total noise should be applied to the A/D converter input to randomize or "whiten" the quantization noise.

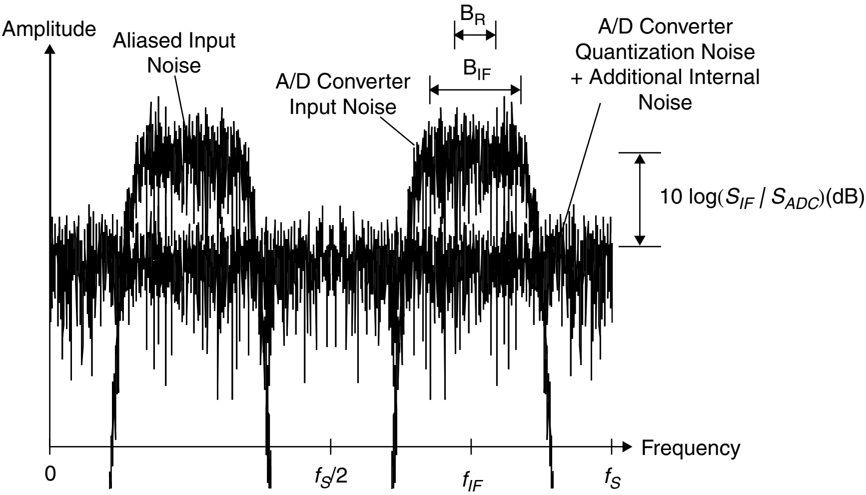

FIGURE 6.16 IF sampling noise spectrums

This can be achieved with rms input noise (σ) equal to the LSB step size (Q). In addition, the input noise power spectral density should be sufficient to minimize the impact on system noise due to the A/D converter noise. The impact on overall noise due to quantization noise is given by[7]

$$\frac{\tilde{\sigma}^2}{\sigma^2} = 1 + \frac{Q^2}{12\sigma^2} \qquad \sigma \geq Q \tag{6.40}$$

Typical operating points are in the range of $\sigma/Q = 2$ to $\sigma/Q = 1$, with corresponding noise power degradation due to quantization of 0.09 dB and 0.35 dB, respectively.

In practice, the SNR of high-speed converters is often such that the noise of the A/D converter is significantly greater than the theoretical quantization noise. In addition, the A/D converter input signal noise bandwidth may be significantly less than the Nyquist bandwidth. This is a significant factor in IF sampling applications where the IF noise bandwidth is often less than ¼ of the Nyquist bandwidth. In this case, the total input and A/D converter noise must be sufficient to whiten the quantization noise, and the power spectral density of the input noise should be sufficiently greater than that of the A/D converter, as illustrated in Figure 6.16. In some cases, out-of-band noise may be added to whiten the A/D converter quantization noise and spurious signals. The out-of-band noise is then rejected through subsequent digital signal processing.

The resulting SNR of the system after digital filtering with receiver bandwidth B_R and sample rate f_s is given by

$$SNR_{\mathrm{SYS}}(dB) = SNR_{\mathrm{ADC}}(dB) + 10\log_{10}\left(\frac{f_S}{2B_R}\right) - 10\log_{10}\left(1 + S_{\mathrm{IF}}/S_{\mathrm{ADC}}\right) \tag{6.41}$$

where $S_{\mathrm{IF}}/S_{\mathrm{ADC}}$ is the ratio of noise power spectral density of the A/D converter input signal to the power spectral density of the A/D converter. The degradation of overall sensitivity due to the A/D converter noise is given by

$$L(dB) = 10\log_{10}(1 + S_{\mathrm{ADC}}/S_{\mathrm{IF}}) \tag{6.42}$$

A/D Converter Sample Clock Stability. The stability of the sample clock is critical to achieving the full capability of an A/D converter. Sample-to-sample variation in the sampling interval, called *aperture uncertainty* or *aperture jitter*, produces a sampling error, proportional to the rate of change of input voltage. For a sinusoidal input signal, the SNR due to aperture uncertainty alone is given by[12]

$$SNR(dB) = -20\log_{10}(2\pi f \sigma_j) \tag{6.43}$$

where f = input signal frequency
σ_j = rms aperture jitter

Similarly, close-to-carrier noise sidebands present on the sample clock signal are transferred to sidebands on the sampled input signal, reduced by $20\log_{10}(f/f_S)$ dB. For example, in an IF sampling application with the input signal ¾ of the sample frequency, the close-to-carrier phase noise of the sample clock will be transferred to the output of the A/D converter output data signal, reduced by 2.5 dB.

6.11 DIGITAL RECEIVERS

The availability of high-speed analog-to-digital converters capable of direct sampling of radar receiver IF signals has resulted in the almost universal adoption of digital receiver architectures over conventional analog I/Q demodulation. In a digital receiver, a single A/D converter is used to digitize the received signal, and digital signal processing is used to perform the downconversion to I and Q baseband signals. Continuing advances in sampling speeds are leading to sampling at increasing frequencies, sometimes eliminating the need for a second downconversion, with the possibility approaching of sampling directly at the radar RF frequency. The benefits of IF sampling over conventional analog I/Q demodulation are

- Virtual elimination of I and Q imbalance
- Virtual elimination of DC offset errors
- Reduced channel-to-channel variation
- Improved linearity
- Flexibility of bandwidth and sample rate
- Tight filter tolerance, phase linearity, and improved anti-alias filtering
- Reduced component cost, size, weight, and power dissipation

The use of a high IF frequency is desirable as it eases the downconversion and filtering process; however, the use of higher frequencies places greater demands on the performance of the A/D converter. Direct RF sampling is considered the ultimate goal of digital receivers, with all the tuning and filtering performed through digital signal processing. The advantage being the almost complete elimination of analog hardware. However, not only does the A/D converter have to sample the RF directly, but unless it is preceded by tunable RF preselector filters, the A/D converter input must have the dynamic range to handle all of the signals present in the radar band simultaneously. Generally, the interference power entering the A/D converter is proportional to the bandwidth of components in front of the

A/D converter. The required A/D converter SNR to avoid saturation on the interfering signals is given by

$$SNR_{\text{ADC}}(dB) = 10\log_{10}\left(\frac{P_I C^2}{N_{\text{ADC}}}\right) \tag{6.44}$$

where

P_I = interference power at A/D converter input
C = crest factor
N_{ADC} = A/D converter noise

The crest factor is the peak level that can be handled within the full-scale range of the A/D converter relative to the rms interference level. It is set to achieve a sufficiently high probability that full-scale will not be exceeded. For example, with gaussian noise, a crest factor of 4 sets the peak level at the 4σ level (12 dB above the rms level) with a probability of 0.999937 that the full-scale is not exceeded on each A/D converter sample.

Setting the system noise level power spectral density into the A/D converter $R(dB)$ above the A/D converter noise gives

$$R(dB) = 10\log_{10}\left(\frac{f_s N_{\text{SYS}}}{2B_{\text{IF}} N_{\text{ADC}}}\right) \tag{6.45}$$

where

N_{SYS} = system noise at A/D converter input in bandwidth B_{IF}

Combining Eq. 6.43 and 6.44 gives the required SNR as

$$SNR_{\text{ADC}}(dB) = 10\log_{10}\left(\frac{2P_I C^2 B_{\text{IF}}}{f_S N_{\text{SYS}}}\right) + R(dB) \tag{6.46}$$

The generation of baseband I and Q signals from the IF sampled A/D converter data is performed using digital signal processing and can be implemented through a variety of approaches.[7] Two approaches are described next.

Digital Downconversion. The digital downconversion approach is shown in Figure 6.17. The signal is sampled by the A/D converter, frequency shifted to baseband, low-pass filtered, and decimated to produce I/Q digital data. The signal spectrum at each stage of the process is shown in Figure 6.18. In continuous-time (Fig. 6.18*a*), frequency is in hertz and is represented by F. In discrete-time (Fig. 6.18*b–e*), frequency is in radians per sample and is represented by ω. The spectrum of the analog input signal $x(t)$ is shown in Figure 6.18*a*, with the signal spectrum centered at F_0 hertz. The signal is sampled by the A/D converter at frequency F_s, producing the time sequence $\tilde{x}(n)$ and frequency spectrum $\tilde{X}(\omega)$ centered at frequency ω_0 with the image centered at $-\omega_0$. The A/D converter output signal is then frequency shifted by complex multiplication with the reference signal $e^{-j\omega_0 n}$, corresponding to a reference signal rotating at ω_0 radians per sample, centering the signal spectrum $\bar{X}(\omega)$ about zero. The unwanted image is re-centered at $-2\omega_0$ if $\omega_0 > \pi/2$ or $-2\omega_0 + 2\pi$ if $\omega_0 \le \pi/2$. The unwanted image is then rejected using the FIR filter with impulse response $h(n)$ producing output $\hat{x}(n)$ with spectrum $\hat{X}(\omega)$. Finally, the sample rate is reduced by

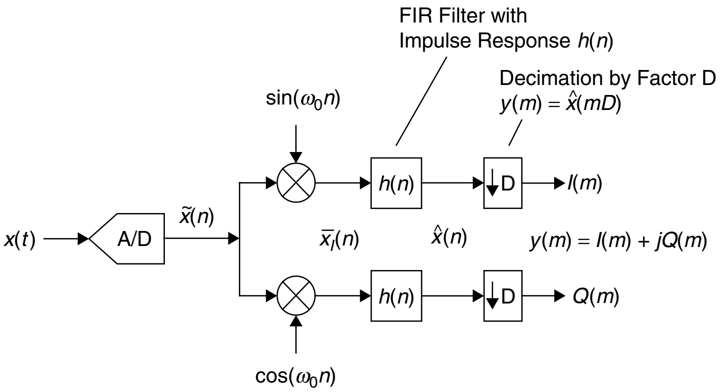

FIGURE 6.17 Digital downconversion architecture

selecting every Dth sample. Provided the filter response $H(\omega)$ has sufficient rejection for frequencies $|\omega| \geq \pi/D$, there will be negligible aliasing and loss of information in the decimation process.

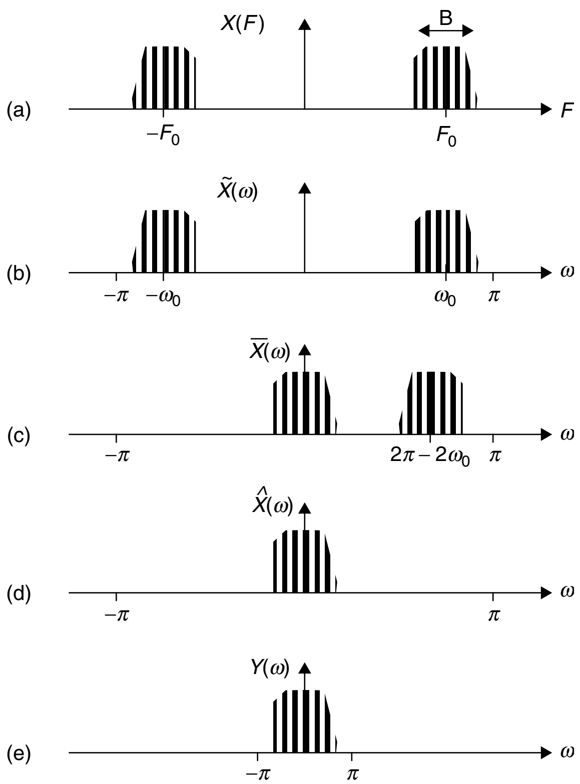

FIGURE 6.18 Digital downconversion spectra

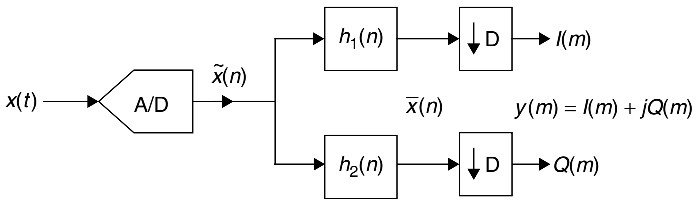

FIGURE 6.19 Hilbert transformer architecture

Hilbert Transformer. An alternative digital receiver architecture is shown in Figure 6.19 with the relevant signal spectra shown in Figure 6.20. The A/D converter output signal $\tilde{x}(n)$ is processed using a Hilbert transformer comprising FIR filters $h_1(n)$ and $h_2(n)$, where the frequency responses are given by

$$|H_1(\omega)| \approx |H_2(\omega)| \approx 1 \qquad |\omega - \omega_0| \le B \tag{6.47}$$

and

$$\frac{H_1(\omega)}{H_2(\omega)} \approx \begin{cases} -j, & |\omega - \omega_0| \le B \\ j, & |\omega + \omega_0| \le B \end{cases} \tag{6.48}$$

The filter outputs form the desired complex valued signal $\bar{x}(n)$ centered at frequency ω_0, while rejecting the image centered at $-\omega_0$. The final stage is to perform a frequency shift and sample rate reduction by decimating the signal by selecting every Dth sample.

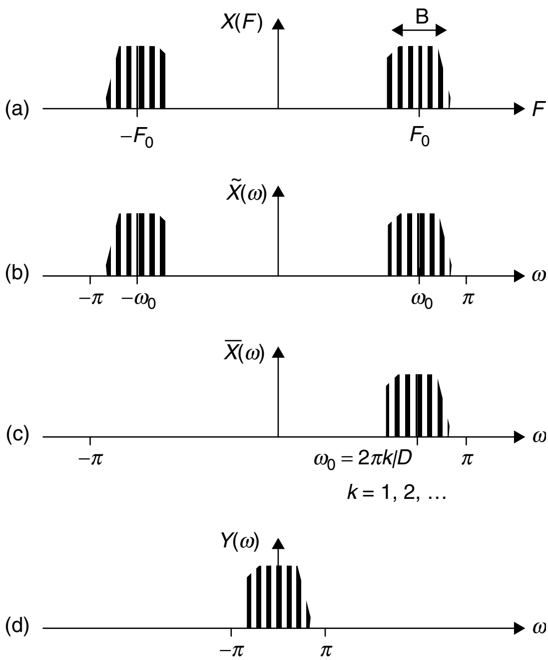

FIGURE 6.20 Spectra of Hilbert transformer receiver

If the spectrum of $\bar{X}(\omega)$ is centered at frequency $\omega_0 = 2\pi k/D$, $k = 1,2,..$, the decimation will center the spectrum $Y(\omega)$ about zero. Provided the filter responses have sufficient rejection for frequencies $|\omega \pm \omega_0| \geq \pi/D$, there will be negligible aliasing and loss of information in the decimation process.

I/Q Errors. Digital I and Q generation does not produce signals without error, as is often stated, but instead allows the generation of these signals with errors that are sufficiently small to be considered negligible. The primary cause of the imbalance is the non-ideal filter responses. An infinite number of taps would be required to set the passband gain to unity and the stopband gain to zero; however, for most applications, sufficient processing resources are available to reduce the errors to insignificant levels. Finite length words for filter coefficients produce non-ideal filter responses. The effect on passband response is typically negligible, but significant distortion of the filter stopband rejection can occur, potentially effecting I/Q balance.

Digital Downconversion Using Multirate Processing and Polyphase filters. There are many variations to these basic approaches, and specific implementations often utilize efficient approaches that minimize the number of calculations required with emphasis on reducing the number of multiplications, as these require significantly more resources than additions. Two techniques used to reduce the FIR filter processing burden are multirate processing and polyphase filtering.[13] The digital downconversion approach is shown in Figure 6.21 using multirate processing. The first FIR filter $h_1(n)$ provides sufficient reduction to prevent aliasing in the first decimation by factor D_1, the second filter $h_2(n)$ provides alias reduction for the second decimation and can also be used to correct passband ripple or droop due to filter $h_1(n)$. For large decimation factors, more than two decimation stages may be used.

A popular filter for the first stage is the Cascaded Integrator Comb[14] (CIC) decimator filter that can be implemented without multipliers. These filters provide rejection in the stopband at frequencies that alias to the passband as a result of decimation. Since they provide relatively large passband droop and slow stopband rejection, they are generally followed by a FIR filter that can both correct for CIC passband droop and

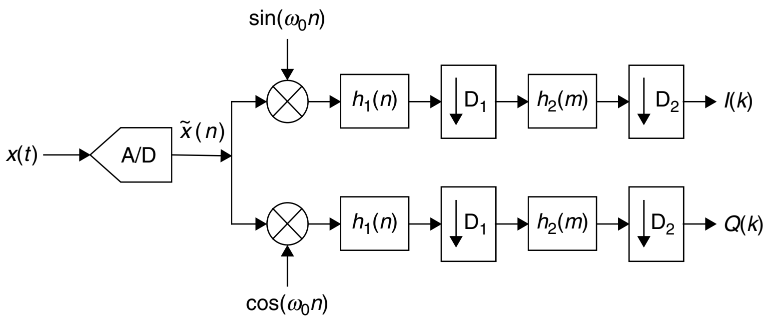

FIGURE 6.21 Digital downconversion architecture

provide the desired stopband rejection response. The kth order CIC filter for decimation factor D has transfer function:

$$H_K(z) = \left[\sum_{m=0}^{D-1} z^{-m}\right]^K = \left[\frac{1-z^{-D}}{1-z^{-1}}\right]^K \tag{6.49}$$

A polyphase filter is a filter bank that splits an input signal into D sub-band filters operating at a sample rate reduced by a factor D, providing a computationally efficient approach to performing the FIR filtering followed by decimation in a digital receiver. Rather than computing all the filter output samples and only using every Dth sample, the polyphase approach calculates only those that are actually used. Figure 6.22 and Eq. 6.50 define how the filter with impulse response $h(n)$, followed with decimation by factor D, is implemented in a polyphase structure. The input signal $x(n)$ is divided into D parallel paths by the "commutator," which outputs samples in turn, rotating in a counterclockwise direction, to each of the FIR filters operating at the reduced sample rate. The outputs of the FIR filters are summed to produce the output signal $y(m)$. This architecture is beneficial as it provides an approach that can be easily parallelized at rate F_X/D.

$$p_k(n) = h(k + nD) \qquad \begin{array}{l} \mathrm{k} = 0, 1, \ldots, D-1 \\ n = 0, 1, \ldots, K-1 \end{array} \tag{6.50}$$

Multi-Channel Receiver Considerations. Modern radar systems rarely contain only one receiver channel. Monopulse processing, for example, requires two or more channels to process sum and delta signals. Additionally, the channels must be coherent, synchronized in time, and well matched in phase and amplitude. Digital beamforming systems require a large number of channels with similar coherence and synchronization requirements and tight phase and amplitude tracking. The coherence requirement dictates the relative phase stability of LO and A/D converter clock signals used for each receive channel. The time synchronization requirement means that A/D converter clock signals for each channel must be aligned in time and decimation must be performed in phase for each channel. Phase and amplitude imbalance between channels is a result of variation in the

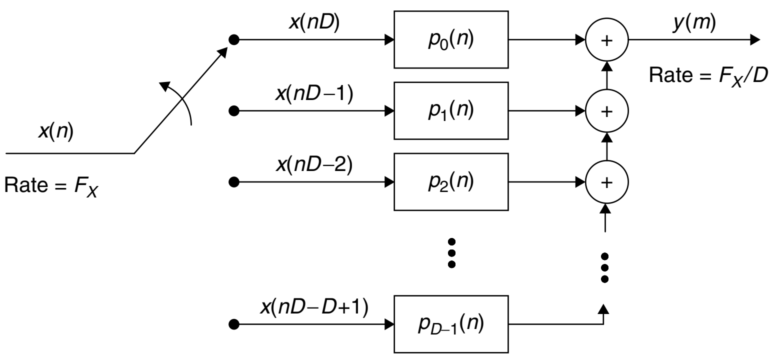

FIGURE 6.22 Decimation using polyphase filters

analog circuitry prior to and within the A/D converter. If the IF filter bandwidth is wide relative to the digital receiver bandwidth, the majority of the error between channels will be a constant gain and phase offset across the receiver bandwidth. A single correction, applied as a complex multiplication of *I*/*Q* data, will correct for gain and phase offsets and is usually adequate to provide the required channel tracking for monopulse applications. When tighter channel tracking is required, such as for sidelobe canceler or digital beamforming applications, FIR filter equalization can be used to correct for frequency dependent variations across the receiver bandwidth. FIR filter equalization can be performed either subsequent to the FIR filtering used to generate *I*/*Q* data or combined with these filters. It should be noted that to correct for frequency and phase variation across the receiver bandwidth requires FIR filters with complex coefficients, applied equally to *I* and *Q* data. Real value coefficients typically used in I/Q generation provide filter responses symmetrical about zero frequency. Correction of IF filter frequency response errors will, in general, require asymmetric frequency correction that can only be provided at baseband using complex coefficients.

The degree to which these multiple receiver channels must track depends on the specific system requirements. Although modern systems typically include some degree of channel equalization function, a reasonable degree of tracking between gain, phase, and timing must be maintained in order to allow the channel equalization to be performed using digital signal processing without consuming excessive processing resources. Also, the relative stability of the radar channels as a function of time and temperature must be such that the corrections can maintain adequate tracking during the time between calibration intervals.

Digital beamforming systems require a large number of receiver channels. In these applications, size, weight, power dissipation, and cost are critical considerations.

6.12 DIPLEX OPERATION

Diplex Benefits. Diplex operation consists of two receivers that simultaneously process returns from transmissions on different frequencies. Transmissions are usually non-overlapping in time to avoid a 6 dB increase in peak power and because most radar transmitters are operated in saturation and simultaneous transmission at multiple frequencies would produce significant transmitted intermodulation distortion.

The sensitivity benefit of diplex operation for detecting Swerling 1 targets is shown in Figure 6.23, increasing with probability of detection (P_D). For example, diplex operation achieves 90% P_D with 2.6 dB less total signal power than simplex. Assumptions made in deriving Figure 6.23 are

1. Returns on the two frequencies are added in voltage or power prior to the detection decision rather than being subjected to individual detection decisions.
2. Separation of the two frequencies is sufficient to make their Swerling 1 fluctuations independent. This depends on the physical length of the target in the range dimension λ_R. The minimum frequency separation is 150 MHz/λ_R (m); 25 MHz will maintain the diplex benefit for aircraft longer than 6 m (20 ft).
3. Equal energy is transmitted in both pulses. A 2:1 imbalance sacrifices only 0.2 dB of the benefit at 90% P_D.

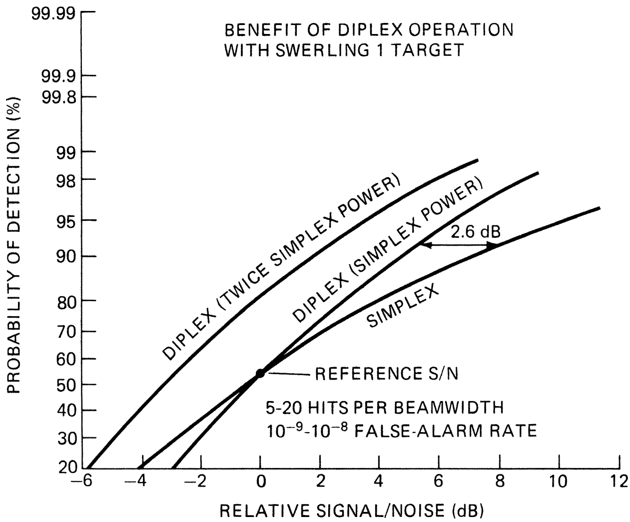

FIGURE 6.23 Diplex operation improves the sensitivity of the receiver

Both linear and asymmetrical nonlinear FM produce a range error as a function of doppler due to range-doppler coupling. These range displacements must match in the two receivers to within a small fraction of the compressed pulse width; otherwise, the sensitivity benefits of diplex operation are not fully achieved and range accuracy may be degraded.

Implementation. Diplex operation can be implemented with a variety of approaches. Complete replication of the receiver channels is typically the most expensive approach and may be required if the frequency separation is very large. A more common approach is separation of the frequencies at the first IF, as this does not require complete duplication of the RF front end or the first LO signal. Separate second local oscillator or I/Q demodulator reference frequencies can be used to process the different frequencies. With the use of high-speed IF sampling, it is also possible to digitize both signals simultaneously using a single A/D converter and perform the frequency separation using digital signal processing. Whichever approach is used, care must be taken to provide adequate dynamic range and linearity to prevent intermodulation distortion from degrading radar performance.

6.13 WAVEFORM GENERATION AND UPCONVERSION

The exciter function of waveform generation and upconversion is often tightly coupled with the receiver function. The requirement for coherence between the receiver and exciter is a major factor for this tight coupling and the use of the same LO frequencies within the receiver and exciter usually results in hardware savings. Similar to the migration to digital receiver architectures, the exciter functionality is increasingly being implemented using digital approaches.

Direct Digital Synthesizer. The Direct Digital Synthesizer[15] (DDS) produces waveforms using digital techniques and provides significant improvements in stability, precision, agility, and versatility over analog techniques. The main limitations are the noise and spurious signals as described below. The general DDS architecture is shown in Figure 6.24. The double accumulator architecture, comprising the frequency and phase accumulators, enables the generation of CW, linear FM (chirp), nonlinear (piece-wise linear) FM, frequency modulated, and phase modulated waveforms. CW waveforms are generated by applying a constant frequency word (digitized frequency representation) input to the phase accumulator, creating a linear phase sequence that is first truncated then input to a cosine (or sine) lookup table that outputs the corresponding sinusoidal signal value to the digital-to-analog (D/A) converter. The frequency resolution is dependent on the number of bits and the clock frequency of the phase accumulator. The output frequency is given by

$$f_{\text{out}} = \frac{M_f f_{\text{clk}}}{2^{N_\phi}} \tag{6.51}$$

where

M_f = frequency word, input to the phase accumulator
f_{clk} = phase accumulator clock frequency
N_ϕ = number of bits of phase accumulator

Linear FM or chirp waveforms are generated by applying a constant chirp slope word (digitized chirp slope representation) to the input of the frequency accumulator, creating a quadratic phase sequence at the output of the phase register. Piecewise-linear or nonlinear FM waveforms can be generated by applying a time-varying slope input to the frequency register. The frequency accumulator may be clocked either at the same rate as the phase accumulator or at a sub-multiple to provide finer chirp slope resolution. If both accumulators are clocked at the same rate, the chirp slope is given by

$$\frac{\Delta f_{\text{out}}}{\Delta t} = \frac{M_S f_{\text{clk}}^2}{2^{N_f}} \tag{6.52}$$

where

M_S = chirp slope word, input to the frequency accumulator
N_f = number of bits of frequency accumulator

Frequency modulated and phase modulated waveforms can be created applying time-varying inputs to the frequency modulation (FM) and phase modulation (PM) ports.

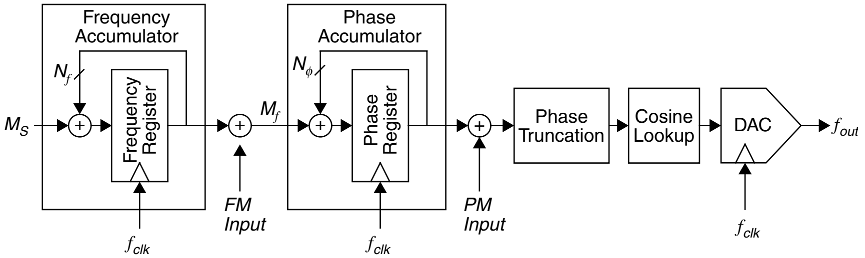

FIGURE 6.24 Direct Digital Synthesizer block diagram

Errors such as phase truncation and D/A converter quantization and nonlinearity produce spurious signals due to their deterministic nature. The spurious signal frequencies generated by a DDS can be readily predicted[16] as they are a function of the digital architecture and programmed frequency. The spurious signal magnitudes are less predictable as the magnitudes of the dominant spurious signals are a function of the D/A converter nonlinearity.

When generating CW waveforms, the D/A converter sequence repeats after 2^K samples where 2^K equals the greatest common divisor of $2^{N\phi}$ and M_f. Thus, spurious signals occur only at frequencies:

$$f_{\text{spur}} = \frac{nf_{\text{clk}}}{2^K}, \quad \text{n} = 0,1,2,... \tag{6.53}$$

In the extreme case where M_f does not contain the factor 2, this creates a spurious frequency spacing of $f_{\text{clk}}/2^{N\phi}$. For example, with a 1 GHz clock and 32-bit frequency accumulator, the spurious frequency spacing can be as close as 0.23 Hz. In most cases, such closely spaced spurious signals cannot be differentiated from noise. Conversely, choosing values of M_f that contain large factors of 2^N creates relatively large spurious spacing. For example, using a 640 MHz clock allows the generation of frequencies at multiples of 10 MHz with all the spurious components occurring at multiples of 10 MHz.

The impact of DDS spurious signals on radar performance depends on the nature of the spurious signals and the type of radar processing involved. Applications using chirp waveforms with large time-bandwidth products are typically less sensitive to DDS spurious signals since the DDS spurious signals chirp at a different rate to that of the wanted signal. The spurious signals are thus rejected during pulse compression. In pulse doppler applications, spurious signals are of much greater concern; however, their effects can be mitigated by ensuring that the DDS generates each waveform from the same initial conditions. Restarting the DDS for every pulse guarantees that the same digital sequence will be input to the D/A converter for each pulse. The result is a DDS output that only contains spectral components at multiples of the PRF.

Techniques have been proposed or incorporated into DDS devices that reduce spurious levels by adding dithering to reduce the effects of limited word lengths. The effect of these techniques and the spurious signals that they are designed to mitigate should be considered carefully as they may be detrimental to radar performance. The use of dithering will randomize the spurious signal, resulting in pulse-to-pulse variations in the digital sequence output to the D/A converter, a result that is undesirable in pulse doppler applications.

Truly random errors are not generated by the digital portion of the DDS. The only nondeterministic errors are a result of the D/A converter performance in the form of internal clock jitter or additive thermal noise and the effect of the phase noise on the input clock signal.

Internal D/A converter clock jitter produces phase modulation of the output signal proportional to the output frequency. Similarly, phase noise present on the clock input signal is transferred to the output signal, reduced by $20\log_{10}(f_{\text{out}}/f_{\text{clk}})$ dB. D/A converter additive thermal noise is independent of output signal frequency and produces both phase and amplitude noise components.

Frequency Multipliers. Frequency multiplication allows signals to be increased in both frequency and bandwidth. Frequency multiplication is frequently used in generating local oscillator CW frequencies where all frequencies are typically based on a

$$A(t)e^{j[\omega t+\phi(t)]} \longrightarrow \boxed{\mathrm{x}M} \longrightarrow \bar{A}(t)e^{jM[\omega t+\phi(t)]}$$

FIGURE 6.25 Frequency multiplier operation

low frequency reference. They also provide the capability for wide-bandwidth chirp waveforms that cannot be generated directly using available DDS devices. Frequency multipliers operate as shown in Figure 6.25, by multiplying the phase of the input signal by the integer multiplication factor M. Since in practice the process typically includes some form of limiting, the output amplitude $\bar{A}(t)$ generally has a lower amplitude variation than the input signal amplitude $A(t)$.

Because the multiplication process multiplies up the variations in the signal phase by factor M, input phase noise and spurious phase modulations are increased by $20 \log_{10}(M)$ dB. Similarly, variations in the phase of the signal as a function of frequency are multiplied up. These variations are produced during signal filtering and may be present on the input signal. For chirp waveforms, this can result in a significant degradation in the range sidelobe performance. Also, practical multipliers may have a significant phase variation as a function of frequency. If the input signal phase distortion is given by

$$\phi(f) = \beta \sin\left(\frac{2\pi n f}{B}\right) \tag{6.54}$$

where

β = peak phase ripple
B = waveform input bandwidth
n = number of cycles of phase ripple

the resulting output distortion produces range sidelobes at times $\pm n/MB$ and magnitude $20 \log_{10}(M\beta/2)$ relative to the main beam of the target return. As an example, generating a chirp waveform that has range sidelobes better than 35 dB using an ×8 multiplier requires that the input signal has less than 0.5 degrees peak-peak phase ripple.

Frequency multipliers can be implemented using a variety of techniques, such as using step recovery diode multipliers or using phase locked loops. Where wide percentage bandwidth and fast settling is required, the most common technique is to cascade a series of frequency doublers or low order multipliers. This type of multiplier can also provide near ideal phase noise performance, but has significant phase modulation as a function of frequency as it contains filters between each stage of multiplication.

Predistortion of the multiplier input waveform is often used in order to produce wideband chirp waveforms with low range sidelobe performance. If the multiplier is characterized by an output phase distortion as a function of input frequency given by $\phi(\omega)$, then a predistortion of the input signal by phase $-\phi(\omega)/M$ will equalize the multiplier response. Predistortion can be performed very precisely by adding the phase modulation via the DDS that is used to generate the chirp waveform.

Waveform Upconversion. Upconversion of exciter waveforms is similar to downconversion within the receiver. Also, similar practical considerations of mixer spurious and image rejection apply. The one significant additional challenge is the rejection of the LO leakage. LO rejection typically imposes tight filter rejection requirements on the RF filters, and for wide-tunable ranges, switched filters are often required.

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