CHAPTER 12

Radar Exciters

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Chapter Outline

12.1 INTRODUCTION

The earliest radar systems had the ability to detect a target and determine the amplitude of the received signal and the distance to the target by measuring the time delay between the transmit pulse and the received signal. Although the intrapulse phase of such a signal was usually quite consistent, the starting phase of the transmitted signal was random, so there was no significance to the phase of the received signal. As a consequence, there was also no fixed or predictable relationship of the phase of the target echoes from one pulse to the next. Because of this lack of pulse-to-pulse coherence in the signal, these systems were termed *noncoherent* radars.[1] Most (though not all) modern radar systems are *coherent*; that is, they detect the phase of the received signal, relative to a well-controlled reference, as well as the time delay and the amplitude. The received signal is treated as a vector having an amplitude and a direction (phase angle). Whether there is a need for coherence depends on the specific application for the radar system; however, any system that needs to cancel clutter, to measure the Doppler characteristics of targets, or to image a target needs to be coherent.[2]

[1] Sometimes the term *incoherent* is used, which has the connotation of a coherent system somehow losing its coherence. The more accurate term is *noncoherent*.

[2] Some techniques used with noncoherent systems, such as coherent-on-receive processing or clutter-referenced moving target indication (MTI) to detect moving targets, provide a crude measurement of the phase of the received signal, but modern radars seldom implement these techniques, opting instead for full coherence.

To implement the ability to detect the phase of the received signal, the transmitted signal has to have a known phase. By far the most common technique is to develop a transmit signal from a set of very stable continuously operating oscillators. These oscillators are then used as the phase reference for the received signals. Such a system transmits a signal that is "in phase with" (having a fixed phase difference with) a reference oscillator or combination of oscillators. There is then significance to the phase of the received signal in the sense that any change in the received signal phase relative to the reference oscillator phase can be attributed to the target characteristics, principally its range. The radar subsystem that develops the transmit signal and supplies coherent local oscillator (LO) signals to the receiver is (usually) called the *exciter*. Often, it is physically combined with the receiver subsystem, in which case the subsystem is called the receiver/exciter (REX). However, in this chapter the exciter is treated as a separate subsystem, independent of the physical architecture of the system.

In addition to supplying the transmit and LO signals, the exciter usually supplies the timing and control signals for the radar. These include, but are not limited to, transmit timing, receiver protection timing, analog-to-digital converter (ADC) sample timing, pulse repetition frequency (PRF), and pulse repetition interval (PRI) timing.

12.2 | EXCITER-RELATED RADAR SYSTEM PERFORMANCE ISSUES

Coherence is required in three major categories of applications: clutter reduction, Doppler processing, and imaging. The reduction of strong, close-in clutter signals is the most demanding of these applications in terms of the system instabilities that can be tolerated. The major design issues involve phase noise, timing jitter, and spurious signal (spur) generation.

Ideal oscillators operate at the specific exact frequency desired from the oscillator. In practice, though, the average frequency over time might be close to the specified frequency, from one instant to the next the frequency will drift due to thermal, mechanical, and aging effects. Oscillator exactness in terms of center frequency is usually expressed in parts per million (ppm). One Hz drift for a 1 MHz oscillator would be specified as a drift of 1 ppm. A 100 MHz oscillator with 6 ppm drift would vary 600 Hz.

Due to the thermal effect on the dimensions of the frequency-determining elements in an oscillator, the frequency of an oscillator will drift with changes in the ambient temperature. The thermal coefficient of expansion for the critical components is somewhat predictable; therefore, the frequency's sensitivity to temperature is likewise predictable. Though design details can minimize the thermal effects, it is difficult to eliminate the effects entirely. Therefore, most oscillators will specify the maximum thermal drift in frequency. Usually this is specified in terms of ppm per degree (Celsius) of temperature change (ppm/degree).

Slow frequency drift in an oscillator does not adversely affect radar performance. One effect would be to produce an incorrect Doppler frequency measurement. The amount of the error, expressed in ppm, would be the same as the oscillator frequency error in ppm. For instance, with a 6 ppm oscillator error, a 100 mph target would look like a 100.0006 mph target. Such small errors in Doppler measurements are usually not considered a problem.

Another effect is to spread the spectrum of the received signal if the frequency were to change during a coherent processing interval (CPI). In the context of this discussion, a CPI is the time it takes to transmit and receive the sequence of pulses used in a pulse-Doppler process. It is usually on the order of several milliseconds. Again, the scale of the error is nearly insignificant for most radar applications. For example, with the same 6 ppm frequency error, a Doppler bin with an ideal bandwidth of 1,000 Hz would appear 1000.006 Hz wide.

12.2.1 Phase Noise Issues

Although slow frequency drift does not have an adverse effect on coherent radar performance, fast frequency modulation does. The objective of a coherent radar system is to measure the phase of the received signal. Analysis of the signal phase leads to the ability to separate targets from clutter in the frequency (Doppler) domain and to image targets by providing high range and cross-range resolution. In all these cases, the integrity with which the phase can be measured depends on the stability of the oscillators that generate the transmit signal and provide local oscillator references for the downconversion process.

The purpose of the exciter is to generate an ensemble of coherent signals used to produce the transmitted signal (not including the final amplifier) and to provide the coherent reference signals for the coherent receiver (downconverter.) These are all phase-stable signals such that they form an accurate reference for determining the phase of the received signal. The received phase is determined by comparing the phase of the received signal with that of the reference signal from which the transmitted signal was generated, as depicted in Figure 12-1. The dark sine wave segments in the figure represent the times for which the oscillator signal is applied to the transmit amplifier and radiated out of the antenna system. The dashed sine wave signal represents the local oscillator signals, which are continuously available for use as the phase reference for the receiver circuits. The segment labeled “Received Signal” is a delayed version of the transmitted signal, resulting from the reflected wave from the target at some range R, leading to the delay time $2R/c$. This received signal phase is compared with the phase of the reference oscillator to determine the target-induced phase shift. If the target changes range from one pulse to

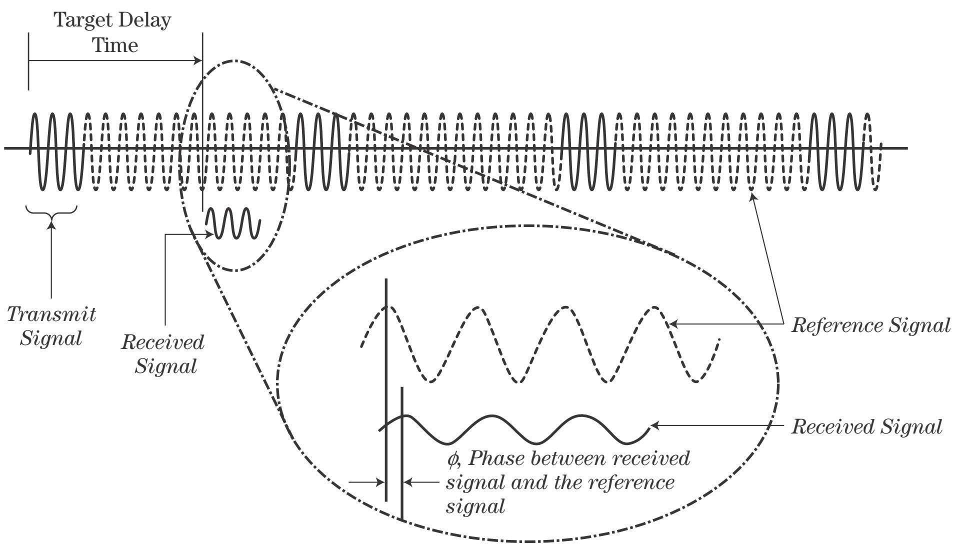

FIGURE 12-1 ■ Measurement of the received signal phase.

the next, this phase will change at a rate proportional to the radial component of target velocity, identifying the target as a moving target.

Classically, the received signal is synchronously detected, producing the two rectangular coordinate system components of the received signal: the in-phase (I) and the quadrature (Q) components. The expressions for the I and Q signals as functions of range to the target, R, and wavelength, λ, are

$$I = A\cos\left(\frac{4\pi R}{\lambda}\right) \tag{12.1}$$

$$Q = A\sin\left(\frac{4\pi R}{\lambda}\right) \tag{12.2}$$

The integrity with which the I and Q signals can be measured depends on the phase stability of the transmit signal reference between the times when the transmit signal is generated and the received signal is detected. If there is any unintentional phase modulation, $\delta\phi$, of the reference signal, then the measurement will be flawed by that amount. In this case, the expressions for I and Q become

$$I = A\cos\left(\frac{4\pi R}{\lambda} + \delta\phi\right) \tag{12.3}$$

$$Q = A\sin\left(\frac{4\pi R}{\lambda} + \delta\phi\right) \tag{12.4}$$

Though the goal is to have a stable reference signal ($\delta\phi = 0$), it is never perfectly phase stable. At a minimum thermal noise will cause a phase modulation, termed phase noise.

Phase noise is not "white" in its spectral characteristics. That is, it does not have a uniform power spectral density. Rather, the phase noise power spectrum is highest at frequencies close to the carrier frequency of the oscillator and reduces with increased frequency offset. The general characteristics of the phase noise (spectrum) are depicted in Figure 12-2, which shows a typical, hypothetical oscillator's single-sideband phase noise power spectrum, $S(f_m)$, as a function of the modulation frequency, f_m, expressed in dB relative to the carrier power (dBc) in a 1 Hz bandwidth. The vertical axis is labeled in units

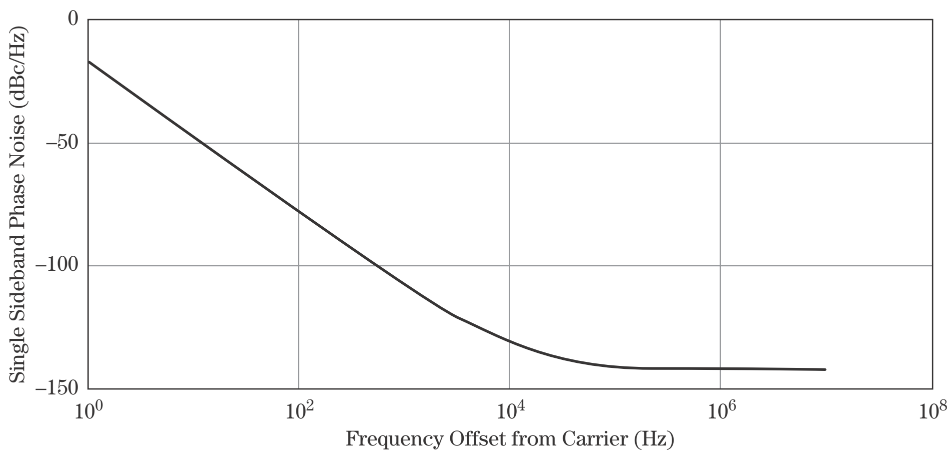

FIGURE 12-2 ■ Typical phase noise spectral characteristics.

of dBc/Hz. The figure depicts the offset frequencies above the carrier, The same curve exists at offset frequencies below the carrier.

One of the most important consequences of phase noise is that in a clutter cancellation system, such as a moving target indication or pulse-Doppler processor, some of the clutter signal energy will be spread throughout the passband of the processor, limiting the target's detectability.

The typical oscillator phase noise spectral characteristic of Figure 12-2 does not represent any particular oscillator but does demonstrate the general shape expected of many stable oscillator phase noise curves. Analysis of a typical oscillator circuit [1,2] reveals that there will be a low-frequency region in which the phase noise power falls off as $1/f^3$ and a region farther from the carrier that falls off as $1/f^2$ before the phase noise reaches a noise floor. A simple mathematical model that can be used to approximate the power spectral density of such an oscillator is

$$S(f) = 10\log 10\left\{\left[10^9\left(\frac{S_{1k}}{f^3} + \frac{2.5S_{50k}}{f^2}\right)\right] + S_{min}\right\} \tag{12.5}$$

where

S(f) is the single-sideband power spectral density of the phase noise in dBc/Hz.

f is the offset frequency.

S_{1k} is the phase noise at 1 kHz offset frequency (in power relative to the carrier power/Hz).

S_{50k} is the phase noise at 50 kHz offset frequency in (in power relative to the carrier power/Hz).

S_{min} is the minimum phase noise at high offset frequencies in (in power relative to the carrier power/Hz).

Figure 12-2 is a plot of (12.5), for the following conditions:

$S_{1k} = 1 \times 10^{-11}$ per Hz (equivalent to –110 dBc/Hz)

$S_{50k} = 1 \times 10^{-13}$ per Hz (equivalent to –130 dBc/Hz)

$S_{min} = 1 \times 10^{-14}$ per Hz (equivalent to –140 dBc/Hz)

Notice that this model would produce an infinite value at 0 Hz, so use of the model is subject to some maximum single-sideband phase noise power, which usually occurs in the vicinity of 5 to 10 Hz offset frequency. Close-in phase noise can be characterized by the spectral "line width." The 3 dB line width is typically on the order of a few Hz, depending on the oscillator frequency. Therefore, plotting or analyzing the effects of phase noise this close to the carrier is usually not considered. Any target with a Doppler frequency offset this low will not be detectable in a clutter environment.

12.2.1.1 Spectral Folding Effects

Two additional calculations are required for a complete analysis of the phase noise requirements. First, the phase noise spectrum extends from a receiver bandwidth below the carrier frequency to a receiver bandwidth above the carrier frequency. Since the radar samples this signal at the much lower radar PRF, there is often significant spectral folding

(aliasing). All spectral components must be included in calculating the total phase noise at any given Doppler frequency. The folded components must be added to the phase noise at a Doppler frequency of interest, $S(f_d)$. In general, an estimate of the sum of all folded spectral components, $S(f)_{total}$, can be found by assuming that the phase noise is at the noise floor, $S(f)_{min}$, for all spectral components above the PRF. The number of spectral folds between the single-sideband receiver bandwidth BW and the PRF is $n = BW/PRF$, so the sum of all these components is simply the phase noise in the first PRF interval plus n times the noise floor, as given in equation (12.6):

$$S(f)_{total} = S(f_d) + S(PRF - f_d) + 2\left(\frac{BW}{PRF}\right) S(f)_{min} \tag{12.6}$$

This must be evaluated at every Doppler frequency of interest.

Figure 12-3a is a plot of equation (12.5) for the same parameters as used previously. This plot is different from that of Figure 12-2 in that the horizontal axes (frequency) for Figures 12-3a and 12-3b are linear, whereas the horizontal axis for Figure 12-2 is logarithmic. Also, Figures 12-3a and 12-3b are plotted for offset frequencies from near 0 to only 10 kHz.

Figure 12-3b is the same plot with the phase noise components below the carrier folded into the spectral region above the carrier. This folding is due to the sampling effects of a pulsed radar.

The total phase noise is the sum of the curves in Figures 12-3a and 12-3b, plus the frequency components from the PRF out to the receiver bandwidth. This is found by exercising equation (12.6). The result is plotted in Figure 12-3c. This shows the total phase noise spectrum including the components below and above the carrier, as well as all folded components from near zero to the receiver bandwidth.

In developing these plots, it was assumed that the phase noise spectrum above the PRF is flat, so the effect of folding all components from the PRF to the receiver bandwidth is simply the number of PRF intervals times the noise floor. In most cases, however, the phase noise spectrum is not "flat" beyond the first PRF line, so the actual calculated values must be used. In this case, the total phase noise at each Doppler frequency is the sum of the noise from all PRF intervals from $-BW$ to $+BW$. Equation (12.7) shows how this summation is performed:

$$S(f)_{total} = S(f_d) + \sum_{i=1}^{n} S(f_d + i \cdot PRF) + \sum_{i=1}^{n} S(f_d - i \cdot PRF) \tag{12.7}$$

where f_d is the offset frequency of interest.

12.2.1.2 Self-Coherence Effect

The second factor affecting phase noise provides some help in meeting detection requirements. If the same local oscillator is used in the receiver downconversion process as is used to develop the transmit signal, there is some "self-coherence" in the system, reducing the clutter residue. Specifically, the noise in the received signal is partially correlated with the noise in the local oscillator, reducing the phase noise out of the receiver downconverter. The effective signal-to-clutter ratio (SCR) improvement depends on the time delay, t_c, to

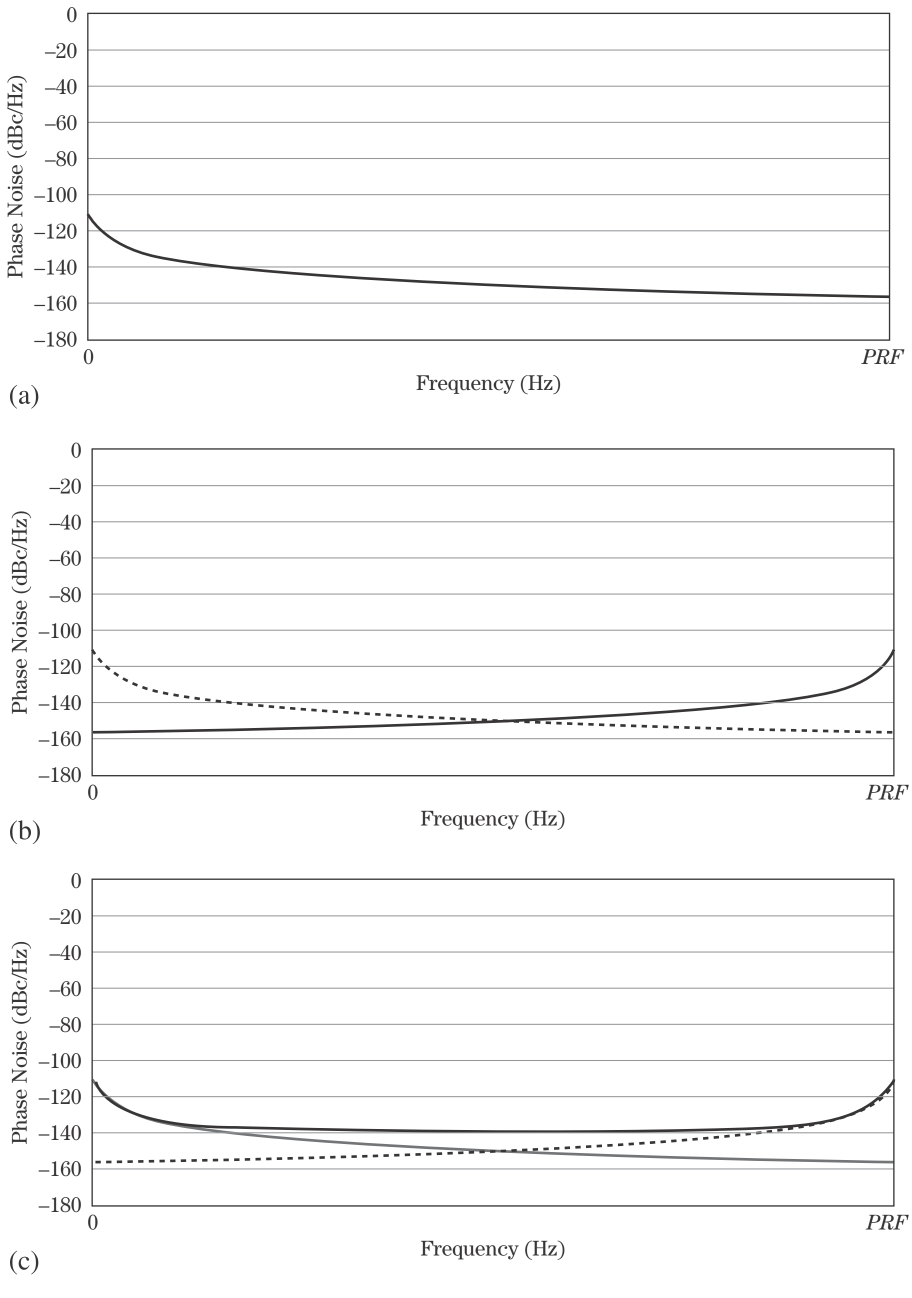

FIGURE 12-3 (a) Phase noise plot. (b) Components of phase noise below the carrier frequency. (c) The sum of all folded components. Parameters are the same as Figure 12-2.

the interfering clutter cell and the Doppler frequency, f_d, of interest. Raven [3] develops the following expression for the improvement, y:[3]

$$y\,(t_c,\ f_d) = 4\sin^2(\pi t_c f_d) \tag{12.8}$$

[3]Note that an equivalent expression found in the literature can be derived from equation (12.8) using the double angle formula, resulting in $y(t_c,\ f_d) = 2[1 - \cos(2\pi t_c f_d)]$.

For small Doppler offset frequencies and close-in clutter the factor is less than unity, reducing the effect of phase noise in conditions for which it is needed the most. This equation predicts an improvement of about 7.6 dB at a Doppler frequency of 10 kHz and 1 km range. The factor must be applied for each range ambiguity and Doppler ambiguity before using the summation given in equation (12.7).

The previously described self-coherence factor applies only in cases for which the same oscillator is used in the upconversion and downconversion processes. If a different oscillator is used in the receiver, then its phase noise adds to the stable local oscillator (STALO) phase noise, with no self-coherence factor applied. Since the effect of self-coherence is significant—and often is the difference between a system meeting detection requirements and not—this is a very important design consideration.

12.2.2 Effect of Phase Noise on Clutter Reduction—MTI Processing

The simplest clutter reduction process is the MTI system, which is implemented as a time-domain canceller that is functionally equivalent to a periodic notch filter having notches at 0 Hz, at the PRF, and at multiples of the PRF (see Chapter 17). Many legacy hardware (analog or digital) systems use a single-delay filter having a frequency response, $H_1(f)$, of

$$H_1(f) = 2\sin\left(\frac{\pi f_d}{PRF}\right) \tag{12.9}$$

or a double-delay filter, having a frequency response of

$$H_2(f) = 4\sin^2\left(\frac{\pi f_d}{PRF}\right) \tag{12.10}$$

These two filter characteristics are depicted in Figure 12-4. For a stationary radar, the clutter will be centered at DC (0 Hz), with some spectral spread due to wind-induced motion of the individual scattering centers. The spectral width of the clutter is typically on the order of 100 Hz or less, as discussed in Chapter 5. The clutter residue, C_0, which is the amount of clutter power at the output of the MTI filter, is found by integrating the product of the frequency response of the filter and the spectrum of the clutter over the PRF interval

$$C_0 = \int_{-PRF/2}^{PRF/2} |H(f)C(f)|^2 \tag{12.11}$$

where $C(f)$ is the power spectrum of the clutter. Were it not for hardware instabilities, the clutter spectrum might be modeled by the exponential function

$$C(f) = k_c e^{-af^2} \tag{12.12}$$

However, the phase noise characterized by equation (12.5) spreads the clutter spectrum out to a bandwidth of 1 MHz or more, depending on the receiver instantaneous bandwidth. The clutter residue is then significantly higher than that due to intrinsic spectral spread of the clutter alone. The transmitted spectrum for a pulsed radar system is an ensemble of spectral lines spaced by the radar PRF and having a $\sin(x)/x$ envelope as was shown in Chapter 8. Given these sampled data system characteristics of a pulsed radar, both upper

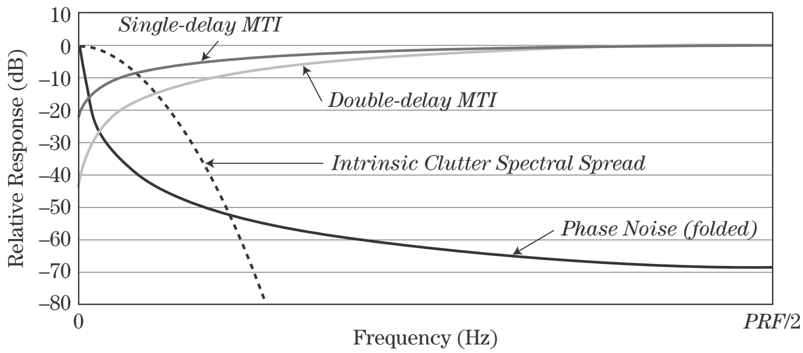

FIGURE 12-4 ■ MTI filter response, showing clutter spectrum and phase noise spectrum.

and lower phase noise sidebands for each of the PRF lines—above as well as below the carrier—fold into the PRF region.

Figure 12-4 depicts the frequency responses of a single- and double-delay MTI filter for a 1 kHz PRF MTI system, along with the natural clutter spectrum and double-sideband phase noise, demonstrating the effect of the phase noise in increasing the amount of clutter power in the MTI filter passband. Notice that the logarithmic plot distorts the MTI filter response shapes from the sine or sine-squared shape of equations (12.9) or (12.10).

12.2.3 Effect of Phase Noise on Pulse-Doppler Processing

Probably the most demanding application regarding phase noise specifications is found in a radar system designed to detect a (small) target in the presence of clutter. The first step in the performance analysis is to determine the signal-to-noise ratio (SNR) for the target signal, as described in Chapter 2. Assuming this is sufficient for detection, given the improvements obtained by whatever signal processing is employed, the next step is to determine the SCR using the radar range equation. The SCR is simply the ratio of the power received from a target to the power received from a clutter cell. The target radar cross section (RCS), σ_t, depends on the size, shape, and materials of the target. Target reflectivity characteristics were presented in Chapter 6. Assuming the interference is due to surface clutter, the clutter RCS, σ_c, depends on the area of the surface clutter being illuminated, A_c, and the average reflectivity per unit area of the clutter, σ^0. Chapter 5 provides a thorough description of the characteristics of various types of clutter.

For a pulse-Doppler system, the waveform is usually range-ambiguous; that is, the distance to the clutter cell may be significantly different from the distance to a target appearing in the same range bin due to foldover of the target, clutter, or both. This can be challenging to the radar in two ways. Not only is clutter RCS larger than the target RCS, but also, for a medium PRF radar, close-in high RCS clutter may appear in the same range bin as a faraway low RCS target. In this case, the target's echo power relative to that of the competing clutter will be decreased by ratio of their ranges raised to the fourth power. In addition, the clutter contributions from several range ambiguity intervals must be summed to determine the true effective ratio between the clutter and target signal. Figure 12-5 depicts a target at some long range, R_t, and several range-ambiguous clutter cells that compete with the target signal. The ranges to the clutter cells are designated Rc_1 through Rc_5. The areas of the clutter cells are determined by the range resolution, which

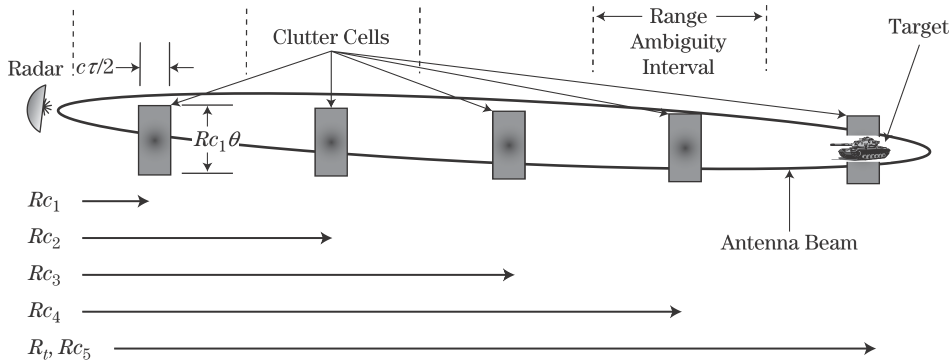

FIGURE 12-5 Far-away target signal competes with close-in clutter signal in a range-ambiguous system.

is proportional to the pulse length for a simple unmodulated pulse, and the azimuthal beamwidth and range to the cell,

$$A_c = \frac{c\tau}{2} R_c \theta_3 \tag{12.13}$$

where θ_3 is the 3 dB azimuth beamwidth of the antenna.

The target signal-to-clutter ratio due to clutter at range Rc_1 is found from

$$SCR = \frac{\sigma_t}{A_c \sigma^0} \left(\frac{R_{c1}}{R_t}\right)^4 \left(\frac{G_t}{G_{c1}}\right)^2 \tag{12.14}$$

where

SCR is the target-to-clutter ratio for a clutter cell at range R_{c1} and a target at range R_t.

σ_t is the target radar cross section.

σ^0 is the average clutter reflectivity per unit area.

A_c is the area of the clutter cell.

G_t is the gain of the antenna in the direction of the target. Assuming that the main beam is centered or nearly centered on the target, this is the peak antenna gain.

G_{c1} is the antenna gain in the direction of the first clutter cell. If the clutter cell is in a direction that is, for example, below the target, it may not be in the main beam of the antenna or may be somewhat down the slope of the main beam. In this case, the antenna gain in the direction of the clutter is less than the antenna gain in the direction of the target. To include the effects of multiple clutter cells "folded" into the same range bin, as depicted in Figure 12-5, the summation of several range-ambiguous clutter cells must be considered.

The net SCR resulting from these effects is

$$SCR = \frac{G_t^2 \sigma_t}{\sigma^0 R_t^4} \sum_{i=1}^{n} \frac{R_{ci}^4}{A_{ci} G_{ci}^2} \tag{12.15}$$

where n is the number of range ambiguities between the nearest clutter cell and the target. Notice that for a radar whose clutter cell area, A_c, is proportional to the range to the clutter cell, R_c, one factor of R_c cancels, leaving R_{ci}^3 in the numerator.

As an example, typical values for these parameters might be as follows:

$\sigma_t = 1\ \text{m}^2$

$\sigma^0 = .01$

$\tau = 1\ \mu\ \text{sec}$

$R_c = 5$ km

$R_t = 40$ km

$\theta_3 = 50$ mrad

The resulting SCR is –62 dB. Clearly, pulse-Doppler processing (discussed in Chapter 17) is required to detect the target in the presence of the clutter energy.

Ideally, the remaining (residual) clutter power is zero in the Doppler bin of interest after the pulse-Doppler fast Fourier transform (FFT) processing is performed. Phase noise, however, spreads the spectrum of the clutter signal into essentially all Doppler bins. As already described, the amount of residual clutter in a given Doppler bin depends on the intrinsic clutter power and spectrum, the power spectral density of the phase noise, and the bandwidth of the Doppler bin. The allowable residual clutter power in a Doppler bin depends on the required probabilities of detection and false alarm—P_D and P_{FA}, respectively—and the probability density function (PDF) of the clutter.

For conventional FFT processing, the Doppler filter bandwidth is proportional to the reciprocal of the dwell time, T_d. If no weighting function were used, the 3 dB bandwidth of the filter would be $0.89/T_d$; however, since a weighting function is usually required to reduce Doppler sidelobes, an aggressive weighting function (e.g., Blackman[4]) leading to a bandwidth of about $1.6/T_d$ is more common. Thus, a typical Doppler bin bandwidth will be

$$B_d = 1.6/T_d \tag{12.16}$$

For a 2 ms dwell time, the resulting Doppler filter bandwidth is 800 Hz (29 dBHz).

Suppose that a P_D of 90% and P_{FA} of 10^{-4} are considered to provide reliable detection. If the interference is noise only and the target exhibits Swerling 1 or 2 fluctuations, the required signal-to-interference ratio (SIR) is about 19 dB. If the interference is clutter having a probability density function PDF with longer "tails," then the SIR needs to be significantly higher (e.g., 28 dB, depending on the specific clutter PDF) for equivalent performance.

The required power spectral density (PSD) of the phase noise, $S(f)_{reqd}$, at any offset frequency of interest is found from

$$S(f)_{reqd} = SCR\,(\text{dB}) - B_d\,(\text{dBHz}) - SIR_{reqd}\,(\text{dB}) \tag{12.17}$$

[4] *Blackman* is the name of one of a number of commonly used signal processing weighting functions. These functions have various degrees of sidelobe suppression and main lobe spreading. The Blackman function provides a high degree of sidelobe suppression, with the attendant large amount of main lobe spreading. Detailed descriptions of many of these functions can be found in digital signal processing texts.

For the previous example,

$$SCR\ (\mathrm{dB}) = -62\ \mathrm{dB}$$

$$B_d\ (\mathrm{dBHz}) = 29\ \mathrm{dB}$$

$$SIR_{reqd} = 28\ \mathrm{dB}$$

which results in a phase noise requirement of –119 dBc/Hz at any Doppler frequency of interest. This numerical example assumed that only one clutter cell, at 5 km, coincided with the target signal at 40 km. If the effect of range folding and spectral folding had been included in the analysis, the required phase noise would have been even lower. Typical coherent radar systems with modest target detection requirements will require that the total phase noise power density from all contributors be no higher than perhaps –110 dBc/Hz at any Doppler frequency of interest. Detection of a stealth target in the presence of range-folded clutter will require a more aggressive system, which may necessitate that the phase noise be no more than, say, –140 dBc/Hz at any Doppler frequency of interest.

The previous example was given for a particular target range, clutter range, and target RCS. The result scales directly with target RCS. For example, for a 10 dB smaller target, the phase noise has to be 10 dB lower. The result also scales with space loss, R^4. If the target range is doubled, the phase noise must be 12 dB lower.

The analysis of target detectability can be reduced to a step-by-step procedure, or algorithm. If the target RCS and range, clutter characteristics and range, and phase noise power spectrum are known, then the following procedure can be used to determine target detectability while accounting for the effects of phase noise:

- Determine the target range.
- Determine clutter RCS from the clutter cell area and average reflectivity.
- Fold in clutter from multiple range ambiguities, adjusted for R^4, to obtain total clutter power in the range bin of interest.
- Determine the space loss, R^4, differences between clutter and target echo power.
- Develop a model for the system phase noise, or obtain a point value from the vendor data sheet.
- Determine the Doppler filter bandwidth for the operating parameters of interest.
- Fold in phase noise from multiple Doppler ambiguities, including negative frequencies, to obtain the total phase noise power in the Doppler bin of interest.
- Multiply the phase noise PSD and the filter bandwidth to determine clutter reduction at a given Doppler frequency.
- Adjust clutter reduction by the self-coherence term (range-dependent) of equation (12.8).
- Determine the resulting signal-to-interference ratio.
- Determine the minimum SIR required to obtain the desired P_D and P_{FA} for the expected interference PDF.
- Determine the minimum detectable target RCS from the required SIR.

12.2.4 Effect of Phase Noise on Imaging

The unavoidable phase noise of the exciter oscillators manifests itself as a spectral spreading of the transmit signal and of the local oscillators used in the receiver downconversion

process. This spectral spreading has several effects, including spectral spreading of the desired target signal and its spectral (Doppler) characteristics, spectral spreading of the interfering clutter signal, and increased sidelobe levels in imaging processes, such as used with a stepped frequency system or synthetic aperture system. If the goal of a target imaging system is a recognizable set of scattering centers such that a pattern recognition system can be used to identify or classify the target, then the image sidelobe structure will alter that image. Most target imaging systems use only the top 40 dB or so of image dynamic range, so the sidelobes in the image need be suppressed only about 40 dB. While this is not as demanding as the clutter cancellation applications, phase noise still needs to be considered.

12.3 EXCITER DESIGN CONSIDERATIONS

12.3.1 Transmit Signal

Modern coherent radar systems require that the phase of the receiver LO represents the phase of the transmitted signal, delayed by the round-trip propagation time for the radiofrequency (RF) signal. In the master oscillator-power amplifier (MOPA) approach, which is the most common architecture for coherent radars, the transmitted signal is developed by modulating the signals developed in a master oscillator and amplifying the modulated signal with a power amplifier. The master oscillator usually comprises a set of very stable oscillators, each of which is often phase-locked to a common stable crystal oscillator. This network of oscillators and the associated timing and control circuits are called the exciter. Figure 12-6 shows a top-level block diagram of a coherent radar using a MOPA technique. The exciter is the part indicated by the dashed box.

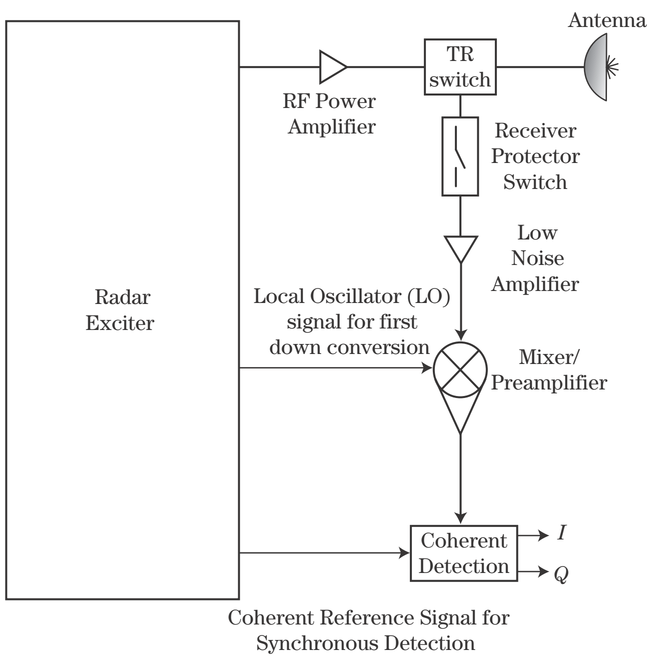

FIGURE 12-6 ■ General diagram of exciter as part of the radar system.

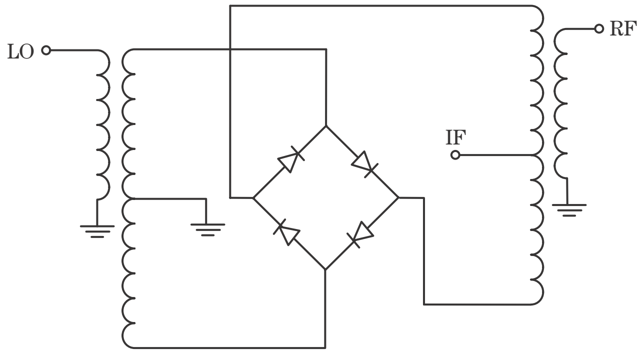

FIGURE 12-7 ■ Diagram of a double-balanced mixer.

Typically, the transmit signal is developed by mixing two or more oscillator signals to produce the sum of the frequencies of these two (or more) oscillators. Though the RF transmit signal could be generated using a single oscillator, the received signal must be downconverted to a lower frequency than the RF for high-fidelity processing. For example, an X-band radar having a carrier frequency of 10 GHz would have a received signal frequency that would be difficult to filter with a narrow bandwidth filter. This is because filter technology is such that it is difficult to produce filters with less than, for example, 1% bandwidth. This means an X-band filter would have a bandwidth of 100 MHz or more. A matched filter for a 1 microsecond pulse, for example, would be a bandwidth of about 1 MHz, a narrower bandwidth than available at X-band. Also, current coherent radar system designs include A/D conversion for digital signal processing. Current analog-to-digital technology does not provide for high-fidelity conversion of a 10 GHz signal. Therefore, the received frequency must be converted down to an intermediate frequency (IF) that is easier to process. Typical IFs for current systems are in the range from 100 MHz to 300 MHz. To downconvert the received RF to an IF, a local oscillator is required in the receiver. This oscillator must be coherent with the transmitted signal. The typical technique for producing an IF in the received signal is to upconvert an IF signal in the transmitter (exciter) and to use the same oscillators in the receive process.

Figure 12-7 is the schematic diagram of a frequency mixer, commonly called simply a double-balanced mixer. A mixer is a three-port device that produces the sum and difference of the input frequencies. The RF and LO ports normally operate at a high frequency (HF) relative to the IF port. For example, a C-band mixer might have a signal capability of from 2 to 4 GHz at the RF and LO ports and an IF port operational band from, say, 100 MHz to 500 MHz. As discussed in Chapter 11, in a receiver a mixer is usually used to downconvert the RF input to an IF output. In this case, the RF might be 3.2 GHz and the LO might be 2.9 GHz, resulting in an IF of 300 MHz. In a radar exciter, the intent is to produce the desired RF frequency from the IF and LO inputs. Usually, this is the sum of the LO frequency and the IF (LO + IF); however, in some cases the difference frequency (LO – IF) is desired. For example, it would be typical to have an LO frequency of 9.0 GHz and an IF input frequency of 300 MHz, resulting in an RF of 9.3 GHz.

Figure 12-8 depicts the use of a mixer to produce a signal at a frequency determined by the frequencies of two oscillators. One signal is input to the local oscillator, or “L” port of the mixer. This signal is normally at a frequency on the order of the desired output frequency, or at least in the same band. For an X-band radar, this might be on the order of 9 GHz. The other input is a signal into the intermediate frequency, or “I” port of the mixer. This signal in the exciter is usually the same frequency as the final IF of the radar receiver.

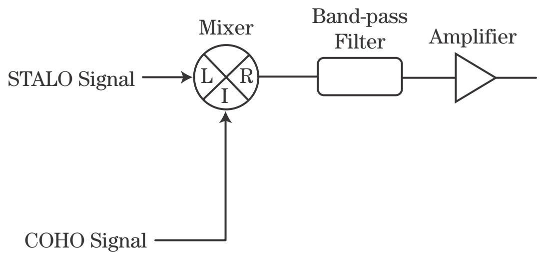

FIGURE 12-8 ■ Method for generating a signal whose frequency is the sum of two other signals' frequencies.

Though potential IFs vary widely, one might choose 300 MHz. The radiofrequency (RF), or "R" port of the mixer, will produce a signal whose frequency is the *sum* of the frequencies of the two inputs. For example, if the L port has a signal at 9.0 GHz and the I port has a signal at 300 MHz, then the output of the mixer at R will be at 9.3 GHz. For the typical coherent radar exciter architecture, the local oscillator is usually called a STALO, and the IF oscillator is usually called the coherent oscillator (COHO). Though these terms are most common, some major radar contractors may have other (proprietary) names for these functions.

The mixer also produces other unwanted mixing products, such as the two individual input frequencies (9.0 GHz and 300 MHz in this example) and the difference frequency (8.7 GHz). Additional unwanted mixing products are generated at harmonics (multiples) of the input signals and sums of these harmonic components and differences between these harmonic components. Figure 12-9 depicts the frequencies that would be generated in such a case, showing only the first and second harmonic terms. Table 12-1 lists the frequencies that are generated when mixing a 9 GHz STALO with a 300 MHz COHO. These frequencies are depicted graphically in Figure 12-9. A band-pass filter, chosen to pass only the desired sum frequency (9.3 GHz), is placed after the mixer to ensure that only that frequency is processed by the transmitter amplifier.

Figure 12-10 depicts the frequencies that are produced by the mixer, up to the second harmonic signals. Note that the filter must pass the desired $(f_2 + f_1)$ frequency and reject all others, some of which may be close to the desired frequency, particularly if higher-order harmonics are considered.

The bandwidth of the band-pass filter must be sufficient to include the bandwidth of the transmitted signal. The RF and IF selected for the process must be carefully considered to avoid the possibility of an unwanted signal (mixing product) existing in the passband of the filter. For a fixed-frequency radar, this is usually not a significant problem because the passband filter can have a relatively narrow bandwidth (though a bandwidth less than about 1% of the center frequency is difficult to realize).

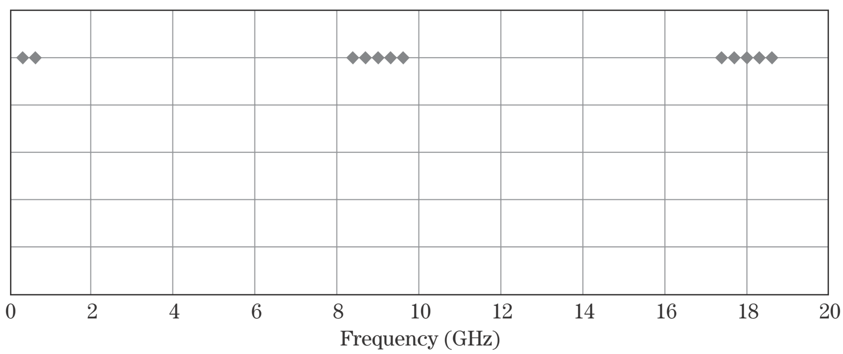

FIGURE 12-9 ■ Mixer output frequencies given two inputs, at 9.0 and 0.3 GHz.

TABLE 12-1 ■ List of Mixer Output Frequencies

Input Frequencies	Output Frequencies
f_1	9.0 GHz
f_2	300 MHz
$f_1 + f_2$	9.3 GHz (Desired, the closest undesired frequency is 300 MHz away.)
$f_1 - f_2$	8.7 GHz
$2f_1$	18.0 GHz
$2f_2$	600 MHz
$f_1 - 2f_2$	8.4 GHz
$f_1 + 2f_2$	9.6 GHz
$2f_1 + 2f_2$	18.6 GHz
$2f_1 + f_2$	18.3 GHz
$2f_1 - f_2$	17.7 GHz
$2f_1 - 2f_2$	17.4 GHz

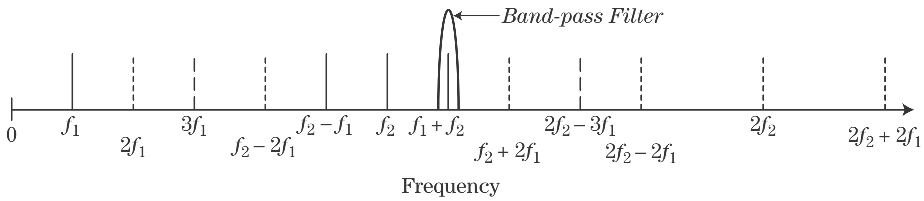

FIGURE 12-10 ■ Frequencies out of the mixer, and filter characteristics.

The amplitude of the signal is reduced as it is processed by the mixer and filter, so an amplifier is placed after the filter to recover the signal loss. The conversion loss of most mixers is on the order of 6 to 7 dB, and the filter will attenuate the signal by an amount on the order of 1 dB, so only modest gain (typically 10 dB) is required of this amplifier.

For the exciter, the desired output is usually the sum of the two input frequencies.[5] Consequently, a single-sideband upconverter is sometimes used in place of the mixer. This device is an assembly of two mixers—signal in-phase and quadrature splitters—and combiners that produces only the desired sum frequency (upper sideband [USB] or lower sideband [LSB]) and greatly suppresses the undesired sideband and carrier frequencies. The undesired signals are not fully suppressed, but the filtering process is made easier with the partial suppression. Figure 12-11 depicts a circuit diagram of a single-sideband frequency converter, which has both the USB and LSB outputs available. The user would terminate the unused output port using a matched load. The desired output would have the undesired frequency suppressed by about 20 dB, relative to a typical mixer shown in Figure 12-7.

For a pulsed radar, the transmit signal exists for only a short period of time, consistent with the pulse length. Except during the transmit pulse, it is desirable not to have any signal out of the mixer. To achieve this, two switches controlled by the timing and control circuits

[5]Some radars, such as those operating in the HF band (3–30 MHz) may employ a high-side local oscillator with the desired output frequency being the *difference* between the two input frequencies.

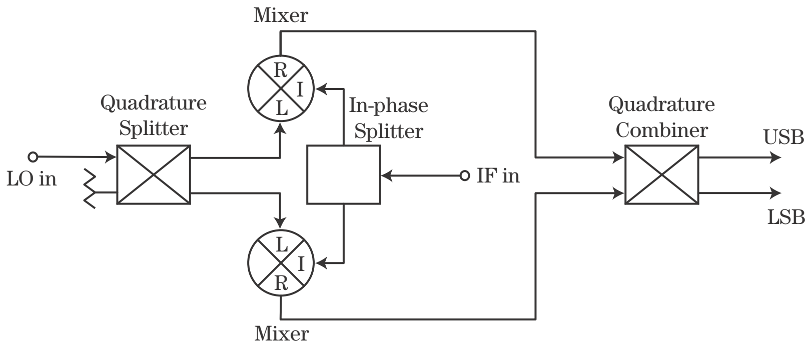

FIGURE 12-11 ■ Diagram of a single-sideband frequency converter.

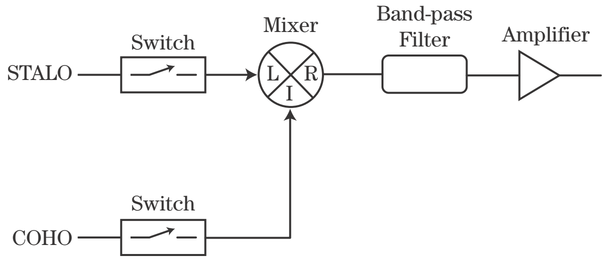

FIGURE 12-12 ■ Method for pulsing the transmit frequency using switches.

are used to pulse the signals into the mixer. Certainly, only one of the two signals needs to be pulsed; however, better isolation is achieved through the mixer if both signals are gated "off" during the receive time. Figure 12-12 depicts the use of the two switches with the mixer circuit. Usually the transmit amplifier is also pulsed on only during the transmit time. Gridded traveling-wave tube (TWT) amplifiers, such as described in Chapter 10, are a common type of amplifier for this application.

Figure 12-13 depicts the diagram of a more complete exciter circuit, which produces a transmit signal to be applied to the transmit amplifier, and the LO and COHO signals for the receiver downconversion and coherent detection process. In typical designs, the STALO is at a frequency somewhat close to the transmit RF, and the COHO is at a significantly lower frequency. For example, for an X-band radar, the STALO might be at 9 GHz and the COHO at 300 MHz. The combination would produce a transmit frequency of 9.3 GHz. Both of the stable oscillators (STALO and COHO) are phase-locked to an extremely stable low frequency crystal oscillator. The most stable of such oscillators is a stress-compensated-cut (SC-cut) quartz crystal oscillator.[6] These are available at frequencies up to about 100 MHz. It is desirable to use as high a frequency as available for this reference oscillator, so that the final output frequency is no larger a multiple of the reference oscillator than necessary.

The STALO and COHO are each followed by a signal splitter, ahead of the switch, so that continuous wave (CW) versions of these signals are available for the receiver LO and COHO requirements. If the desired transmit signal bandwidth B is wide due to the use of intrapulse coding such as a phase code or linear frequency modulation (LFM), then

[6]There are many other techniques for stabilizing the frequency or phase of an oscillator. The use of a quartz crystal is common for oscillators in the HF, VHF, and UHF ranges.

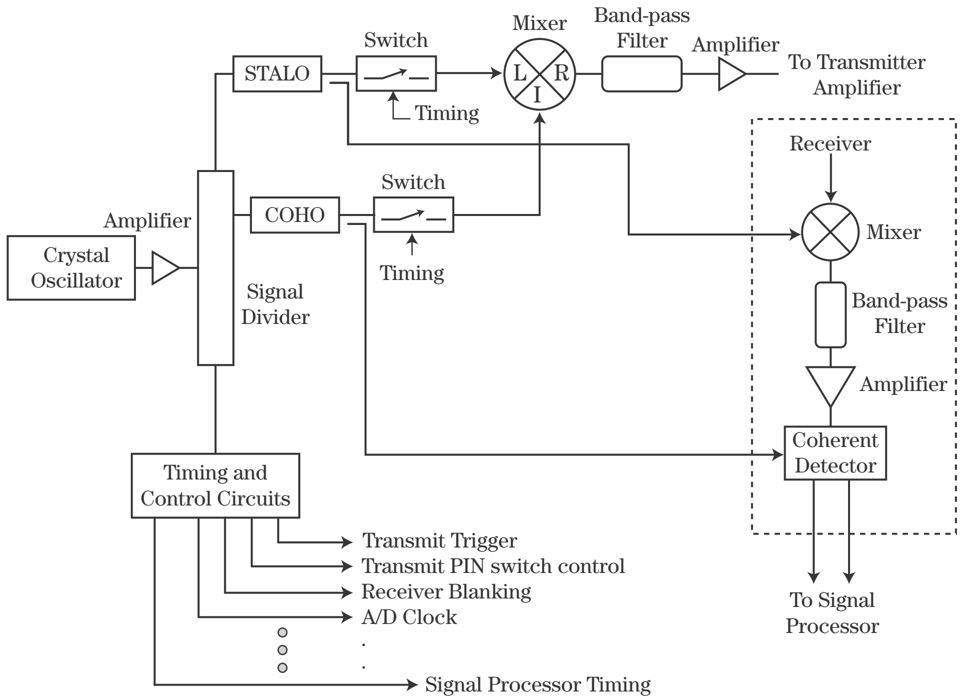

FIGURE 12-13 ■ Diagram of internal components of a simple exciter.

the selection of RF and IF is more critical, as depicted in Figure 12-14. The percentage bandwidth of the band-pass filter must be wide enough to pass the desired signal, with only marginal amplitude reduction over that bandwidth. In this case, for the example shown in Figure 12-14, one of the unintentional mixing products ($f_2 + 2f_1$) is beginning to invade the passband of the filter. As shown in the figure, the condition is probably acceptable as it is shown; however, if the bandwidth were any wider, or if f_1 were any lower, then a different IF would be chosen. In general, for a wideband signal, the use of a very low IF compared with the RF creates a problem. In this case, more than one upconversion stage (and likewise multiple stages of downconversion in the receiver) will alleviate the problem. Figure 12-15 depicts the signals, up through the second harmonics, for a 9.0 GHz and 1.0 GHz pair. Compared with Figure 12-8, the separation between the desired 10.0 GHz signal and the closest neighbor is wider. The instantaneous bandwidth of the filter at 10 GHz and therefore the bandwidth, B, of the signal could be as much as 1 GHz. This frequency plan for the upconversion process would result in a 1.0 GHz IF in the downconversion process. This may be a higher IF than desired, so a second downconversion to a lower IF would be required. To maintain coherence with the transmitter, a similar second upconversion in the transmit process would be required.

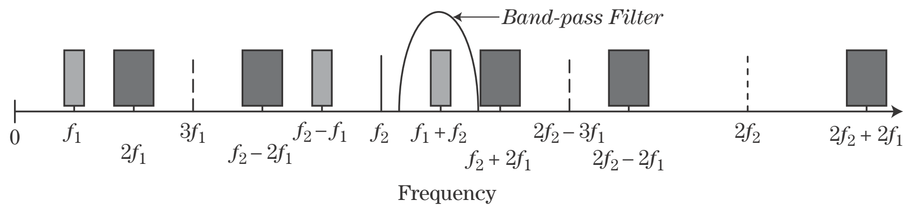

FIGURE 12-14 ■ Frequency plan with a wideband IF signal.

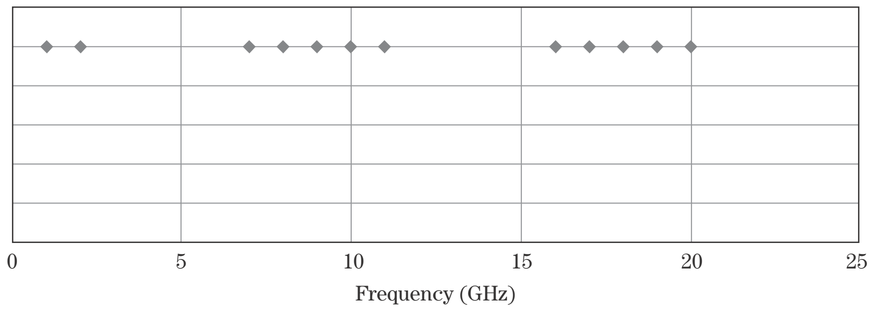

FIGURE 12-15 ■ Frequencies resulting from 9 GHz and 1 GHz pair.

Notice that the bandwidth of the filter selecting the desired signal shown in Figure 12-13 is wider than the equivalent filter shown in Figure 12-10. For a wider bandpass, the so-called skirts of the filter are wider and will pass signals farther from the desired signal than will the narrowband filter. Usually, this condition requires that there be a lower ratio of frequencies into the mixer than for the narrow band system. Whereas the narrowband system may allow a ratio of up to about 50:1 between the RF and IF, the wideband system may allow only a ratio of about 3:1 to 4:1 for each mixing stage. Therefore, two mixing stages might allow for a final ratio of 10:1 to 20:1, and three mixing stages would be required for a ratio on the order of 50:1. Table 12-2 lists the frequencies that are generated when mixing a 9 GHz STALO with a 1 GHz COHO. These frequencies are depicted graphically in Figure 12-15.

Figure 12-16 is a photograph of several typical microwave filters [4]. The devices in Figures 12-16a and 12-16b of the figure are connectorized versions, and Figure 12-16c is a surface-mounted version. Since these filters are passive devices, they typically have only an input and output coaxial connection.

Figure 12-17 shows the band rejection characteristics for a typical tubular filter having a bandwidth of 5% of the center frequency, for example, 500 MHz (3 dB) bandwidth centered at 10 GHz [4]. This filter will pass a signal in the band from 9.750 to 10.250 GHz. If the undesirable mixing products from a mixer ahead of the filter are specified to be at least 40 dB down from the desired signal, then for a five-section filter their frequencies

TABLE 12-2 ■ Output Frequencies

Input Frequencies	Output Frequencies
f_1	9 GHz
f_2	1 GHz
$f_1 + f_2$	10 GHz (Desired, the undesired frequency is 1,000 MHz away.)
$f_1 - f_2$	8 GHz
$2f_1$	18 GHz
$2f_2$	2 GHz
$f_1 - 2f_2$	7 GHz
$f_1 + 2f_2$	11 GHz
$2f_1 + 2f_2$	20 GHz
$2f_1 + f_2$	19 GHz
$2f_1 - f_2$	17 GHz
$2f_1 - 2f_2$	16 GHz

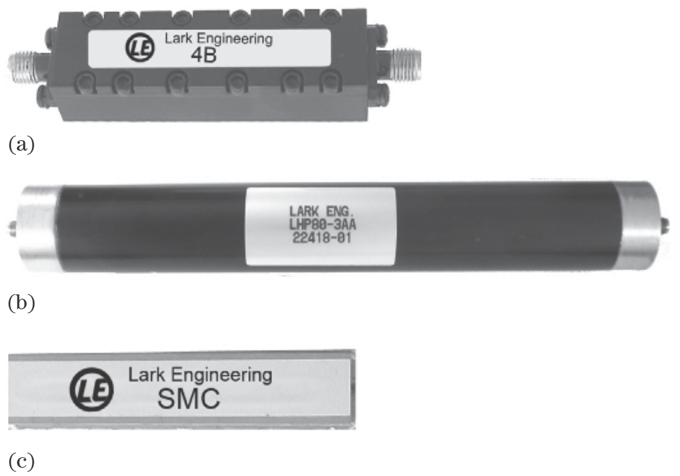

FIGURE 12-16 ■ Typical IF filters. (a) Connectorized. (b) Tubular style. (c) Surface mount style. (Courtesy Lark Engineering [4]. Used with permission.)

must be at least 1.6 bandwidths (800 MHz) above 10.0 GHz, or 1.7 bandwidths (850 MHz) below 10.0 GHz—that is, above 10,800 MHz or below 9,150 MHz. The data points for this calculation are shown as two dark dots on the figure.

There is a trade-off in the selection of filter order between frequency and time-domain characteristics. A higher-order filter, such as a six- or eight-section filter, will allow undesired mixing product signals to be closer to the center frequency (but no less than 1 bandwidth, or 500 MHz away, for the previous example and the filter characteristics shown). However, when a high-order filter (above, e.g., five sections) is used, the time-domain response usually exhibits "ringing" or a decaying oscillation in the signal. Fewer filter sections, typically three or four, provide a better response from a time-domain point of view, but the frequency restrictions are then worse; in the previous example, the undesired signals would have to be 1 or 1.5 GHz away from 10 GHz. No matter how many filter sections are used, as a practical matter the filter passband bandwidth must be somewhat less than the separation between the closest unwanted signals, by a factor of at least 1.5.

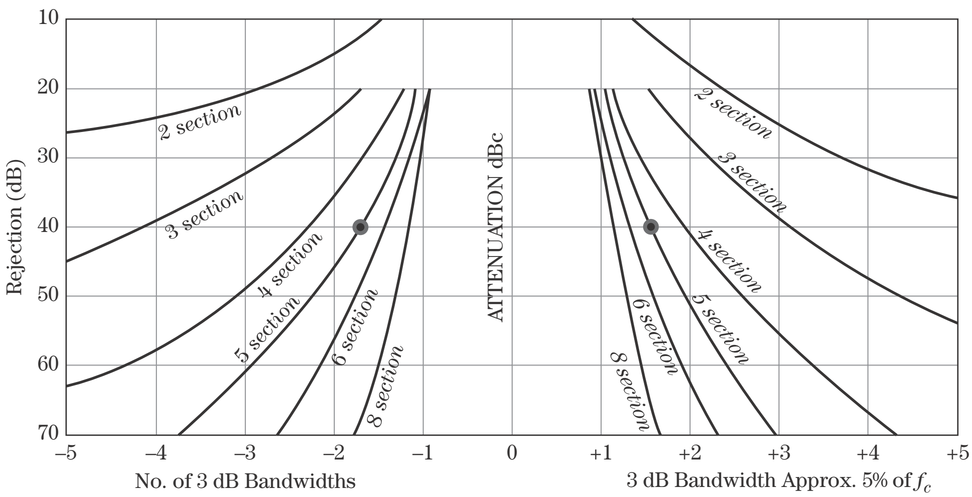

FIGURE 12-17 ■ Filter characteristics for a typical 5% bandwidth tubular filter. (Courtesy Lark Engineering [4]. Used with permission.)

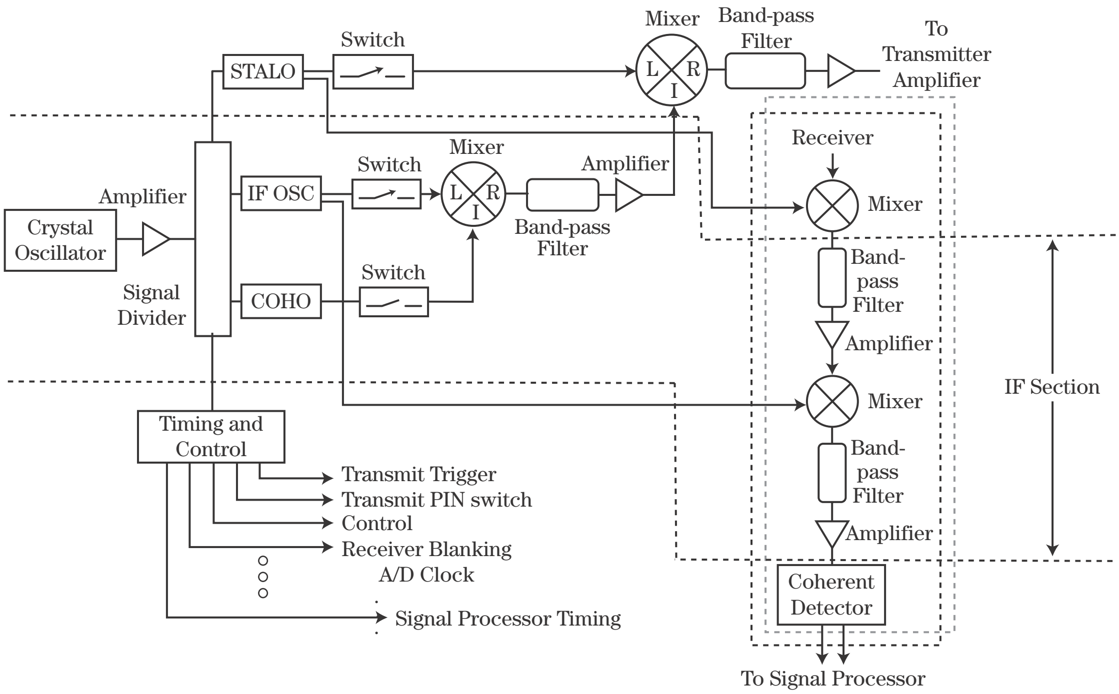

FIGURE 12-18 ■ Diagram of an exciter with two upconversion stages.

Figure 12-18 is a block diagram of an exciter that has two upconversion stages. The additional coherent oscillator required for this architecture is often called the IF oscillator. This is the most typical exciter configuration because it allows for a somewhat wideband signal, such as that used in a pulse compression waveform. The selection of frequencies is based on the need to prevent unwanted mixing products from lying within the passband of the band-pass filters at each mixing stage. Which mixer ports are used for a given oscillator signal depends on the bandwidth specified for each of the mixer ports. Typically, the IF port will support a lower range of frequencies than either the LO or RF ports. The specific ports are chosen depending on the relative frequencies of the COHO and the IF oscillator. Note that there is a filter following every mixer in the diagram.

12.3.2 Waveform Generation

There is a wide variety of techniques for generating a wideband waveform for a coded pulse system. The most common modern radar techniques use either a wideband LFM within the pulse or a pseudorandom biphase code. The properties and application of these and other wideband waveforms are discussed in Chapter 20. The wideband waveform is usually developed in the IF section of the exciter in Figure 12-18. From the point of view of the exciter, the IF section is the set of components that operates at frequencies near those of the COHO rather than at frequencies near the RF. The block diagram of Figure 12-18 shows the section designated as the IF section.

12.3.2.1 Intrapulse, Pseudorandom Biphase Modulation

If pseudorandom biphase code pulse compression is used, the phase code is applied to the IF or COHO signal using a biphase modulator controlled by the waveform generation

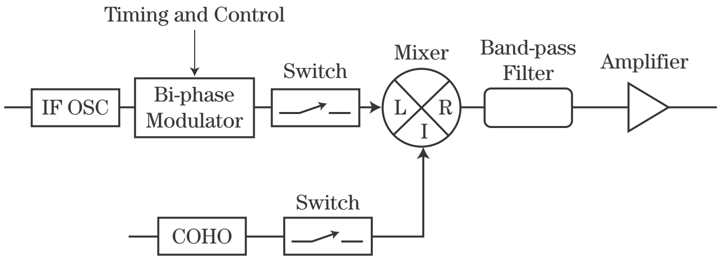

FIGURE 12-19 ■ Use of a biphase modulator to create pseudorandom phase modulation.

function of the radar computer. Figure 12-19 depicts the addition of a biphase modulator in the IF circuit. The phase either changes from one state (e.g., 0 degrees) to a complementary state (e.g., 180 degrees) or is left unchanged for each short time period consistent with the required bandwidth. Each such time period, which represents a fraction of the total pulse length, is called a *chip*. For example, a 200 MHz bandwidth would be consistent with a 5 nsec chip period. The choice to change the phase or leave it unchanged from one chip to the next would be made randomly every 5 nsec for, say, a 1 microsecond pulse. Since the computer processor develops the random sequence, the phase sequence is actually pseudorandom in nature. The computer stores the specific code sequence that was transmitted so it can perform the correlation process during the receive function. Chapter 20 provides a thorough mathematical description of the characteristics of the biphase modulated technique. Concerning implementation of the biphase modulation technique in an exciter, it is valuable to mention here that the spectrum of the biphase modulated waveform is a $\sin(x)/x$ shape, requiring a linear phase (vs. frequency) filter to properly maintain the integrity of the waveform. Bessel filters, having a "rounded" passband characteristic and linear phase, are often used in conjunction with biphase modulation techniques.

12.3.2.2 Intrapulse Linear Frequency Modulation (LFM)

If a frequency agility waveform or intrapulse LFM is required, then a synthesized signal generator is used in the IF circuit to produce the frequency modulated signal. The LFM waveform is characterized by a linear increase (or decrease) in frequency during the pulse. Often, several hundred megahertz of frequency modulation is imposed onto the pulse. Chapter 20 provides a thorough mathematical description of the operation of the LFM mode. The spectrum of the LFM waveform is flat, requiring a relatively flat filter response. Historically, an LFM waveform would be implemented using a surface acoustic wave (SAW) pulse expansion device. The received signal would use a complementary device called a SAW pulse compressor to produce a high-resolution pulse-compressed signal in the time domain. In modern systems, an LFM waveform is usually generated using a direct digital synthesizer (DDS). Figure 12-20 depicts where the DDS would be inserted into the exciter architecture. The baseband output of the DDS is then mixed with the IF oscillator. The DDS device has a random access memory in which samples of a sine wave are stored. The values are read out of the memory at a varying rate determined by the clock pulse rate (usually on the order of several hundred MHz), producing a sinusoidal signal at a controlled frequency. The sine wave is quantized in amplitude to an accuracy determined by the number of bits in the output digital-to-analog (D/A) converter. As with the biphase modulator, the DDS is controlled by the signal processor, which defines the waveform in terms of the start frequency, stop frequency, and rate of change of frequency.

Next Page

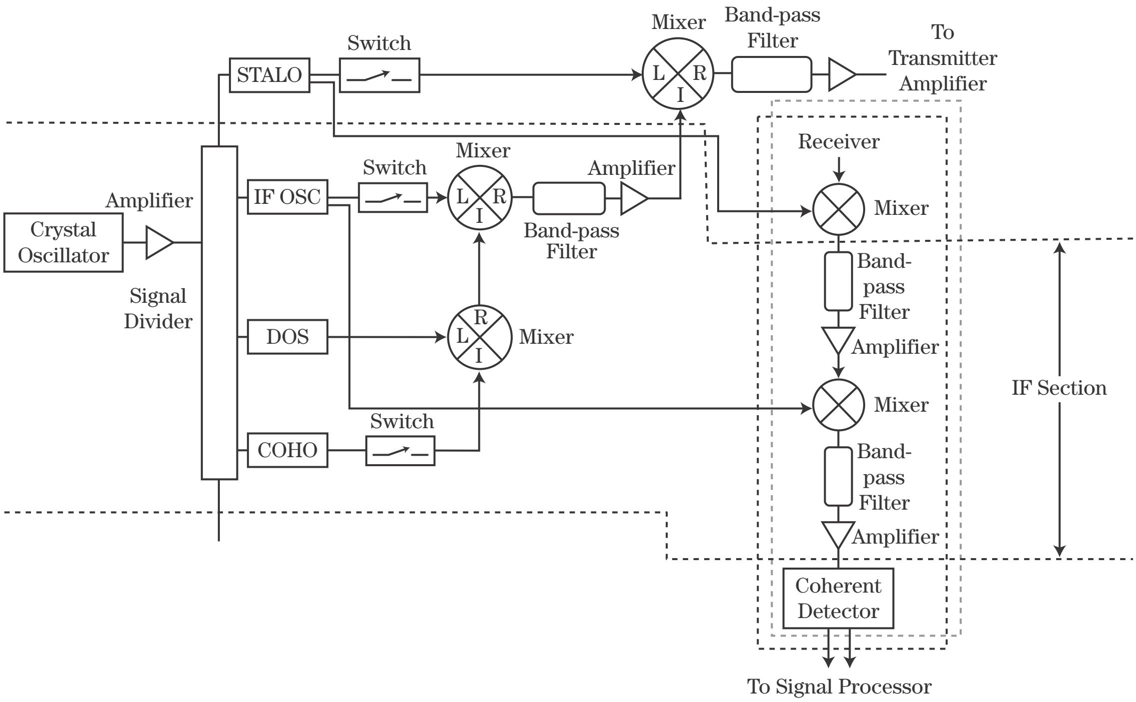

FIGURE 12-20 ■ Exciter drawing showing incorporation of a DDS.

Current DDS technology exhibits phase noise significantly worse than that required for detection of a long-range target in severe clutter. Therefore, the wideband waveform is used only after detection is made and target *imaging* or *range-profiling* is required. For example, in the search mode, when detection performance needs to be optimized the radar will generally not use a wideband waveform but rather will use either an uncoded pulse or modest frequency modulation bandwidth. In this case, high range resolution is not required, but low phase noise is. When the mode is switched to one requiring high range resolution—for instance, in a target identification application—then the wideband waveform generator is employed. Phase noise will be increased; however, low phase noise is not required at this stage. If the radar has an operational mode where low phase noise is required at the same time as high range resolution imaging, then a low-noise method for waveform generation will be required.

12.3.2.3 Dwell-to-Dwell Frequency Change

Better phase noise performance is attainable using a more sophisticated frequency synthesizer. There exist a large number of commercial synthesizers operating at frequencies from hundreds of MHz to 18 GHz and above. A typical high-performance synthesizer will change frequency in as little as a microsecond to as much as several hundred milliseconds. The phase noise performance of a frequency synthesizer is better than a DDS-based synthesizer but not as good as that of a fixed-frequency oscillator. Synthesizers do not switch frequency fast enough to implement intrapulse frequency modulation but can provide the IF signal for dwell-to-dwell frequency hopping. This class of synthesizer is large, usually rack-mounted, or on the order of the size of a shoe box. Figure 12-21 depicts an exciter architecture with a stepped frequency synthesizer, capable of changing transmit frequency