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Chapter 2

Technical Issues in Ultra-Wideband Radar Systems

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CONTENTS

I. INTRODUCTION

CHAPTER OVERVIEW

This chapter introduces the basic concepts of ultra-wideband (UWB) radar waveforms and systems. It begins with a discussion of the capabilities afforded by conventional systems which, in turn, motivate the study of UWB waveforms. This set of existing capabilities also serves as a baseline for comparison of capabilities offered by UWB systems. Following that discussion is a brief review of key research areas which have brought UWB technology into focus. At the time this book was published, the appropriate areas for application of UWB technology were still fluid and to some extent uncertain; to aid the reader in discerning which applications may be appropriate, this chapter continues with more detailed material on fundamental radar principles and a discussion of the current technical issues concerning the use of UWB signals for radar.

COMMENT ON SIGNAL ANALYSIS METHODS

The common tools of waveform analysis and synthesis such as the Fourier and Laplace transforms carry some implicit assumptions about the nature of the system under consideration. One of these assumptions is that the system which generates and/or processes the signal of interest does not change within the duration of the signal. This property is generally called time-invariance. Another assumed property is that the system is linear (see Appendix 2C for a discussion of the concept of nonlinearity). Basically, linearity means that the system response does not change in response to changes in the magnitude or the complexity of the signal being processed. For narrowband systems, the requirement for linear time invariance (LTI) is usually met. It turns out that the LTI property is also met in most cases for UWB systems as well. However, due to the very high resolution properties of some UWB waveforms, there is an increased likelihood that the LTI assumption may not hold, particularly for time invariance. Hence, it is important to recheck these assumptions on a case-by-case basis when applying conventional analysis tools.

EXISTING TECHNOLOGY CAPABILITY

There are a number of fundamental physical principles that bound the performance of any radar system. The approach to, and the feasibility of, attaining improvements in radar performance must always be

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viewed in light of these boundary conditions. Some examples of radar performance dependencies are presented in Table 2.1. In many areas, current technology has already reached the limits of achievable performance. In light of these limitations, it is important to seek out new approaches which sidestep these limitations, or at least move them out of the way a little more. The UWB signaling approaches are being examined to see if they offer any unique advantages in this regard. While the conclusion on this issue is not yet definite, Section IV of this chapter provides some discussion of the possibilities which make continued investigation worthwhile.

Before proceeding, it is necessary to be more precise about the meaning of the word *ultra-wideband* (also *ultrawideband*). There are many similar sounding names in the radar and electronic countermeasures literature (e.g., ultrabroadband, wideband). Foreign literature adds more (e.g., superwideband). The key feature of a UWB waveform is its large instantaneous *percent* bandwidth.

The percentage bandwidth is defined as follows:

$$\%\mathrm{BW} = \frac{2(f_H - f_L)}{f_H + f_L} \times 100$$

where f_H and f_L are the upper and lower band edges of the signal, respectively. (The percentage bandwidth is also referred to as *fractional* bandwidth when not converted to percent.) Hence, it is also important to define the term band edge. In conventional narrowband systems, it is common to define the band edges as the frequencies at which the power spectral density is 3 dB below what it is at the center of the spectrum. This is convenient in the narrowband case because the spectrum is generally symmetrical about the center frequency and because the spectral region between the 3 dB points contains approximately 90% of the spectral energy. We will see later in this chapter that the spectrum of a UWB signal is not symmetric; in fact, in some cases the majority of the energy lies below what might be called the center frequency (see Section III).

A Defense Advanced Research Projects Agency (DARPA) panel on UWB technology published one definition for UWB which has come into somewhat common use. The panel's criterion for UWB was 25%.[9] However, a more supportable limit is 20%, because this is the percentage bandwidth above which the angle and time/frequency resolution properties become coupled (see also the section on Measurement Resolution and Reference 11, p. 2096).

HISTORICAL BACKGROUND

Conventional radar waveform design utilizes a small percent-bandwidth signal in order to take advantage of sinusoidal resonance effects. Resonance increases frequency selectivity and antenna efficiency. Strictly speaking, a circuit or an antenna is only resonant at a single frequency, but near-resonance conditions can persist in many designs at frequencies up to 10% from the resonant frequency. However, this percentage is still quite confining on the allowable combinations of center frequency and bandwidth. To avoid this percent bandwidth limitation, some radar waveform designs incorporate multiple center frequencies (or a frequency sweep). However, the *instantaneous* percent bandwidth used in these designs is still confined to obtain the benefit of sinusoidal resonance.

There are some radar applications that inherently require a larger percent bandwidth. The earliest known application was an Army need for detection of buried objects (a need that persists today). Research reports on this topic date back to the early 1960s.[1] A relatively long wavelength is required for propagation into the Earth's surface, and a relatively large bandwidth is needed to get acceptable resolution of the measured depth of the buried object. The nominal set of parameters for this application is a center frequency of 1 GHz with a bandwidth of 1 GHz for a percent bandwidth of 100. Other potential applications, which are discussed in more detail later in this chapter, are target imaging, foliage penetration and rejection of certain types of clutter, and the detection of low cross-section targets.

At the time that UWB requirements were emerging, there were (and still are) a number of research and development activities which made it possible to test some of the theories of UWB systems. One

Table 2.1 Examples of Current Performance Limits

Performance Parameter	Limitation
Detection range	Depends on average effective radiated power (ERP), target response, propagation medium, and clutter
Average ERP	Depends on antenna gain, transmit power, and duty cycle; current systems already approaching 100% duty cycle limit; transmit power limited by hardware technology; antenna gain limited by acceptable aperture size
Target response	Generally decreasing in size; can approach the size of a small bird
Target identification	Increasing need to perform this function; methods include imaging and natural resonances; resonance method requires low frequencies; imaging requires large (1 GHz) bandwidth; neither of these requirements is easily met with existing technology.
Propagation	Propagation is medium sensitive; need for transmission through earth or water requires low frequencies, while maintaining good resolution properties requires wide bandwidth
Clutter suppression	Achieved using moving target indicator (MTI) or Doppler processing methods, both of which measure target and clutter velocity; performance of both depends on the stability of signal phase; stability (short term coherence) of sinusoidal oscillators limits this; other clutter signals are received through the antenna sidelobes; tradeoffs among beamwidth, aperture size, and average sidelobe level limit the sidelobe clutter suppression
Velocity measurement	Resolution is dependent on the time available for target observation; since the observation time required to obtain the required accuracy generally exceeds one pulse width, ambiguities are introduced with conventional methods
Range measurement	Resolution is dependent on the bandwidth of the signal; for ordinary pulses, bandwidth is inversely proportional to pulse width; shorter pulses have less energy, leading to reduced detection range.

of these research activities is in the area of high power/short pulse radiation. Sources and radiating mechanisms for these pulses were developed for simulating nuclear electromagnetic pulse (EMP) effects to test the susceptibility of electronic components and systems.

Another related research area was concerned with the study of what has been called time-domain electromagnetics.[2-5] The main purpose of this work was to develop more complete methods of characterizing the reflection properties of radar targets. In the process, however, facilities were developed to test new theories, and these facilities included new methods for generation and radiation of UWB signals.

A third related research area originally started with the investigation of alternative kernels for waveform analysis. The fast Fourier transform (FFT) algorithm was published in 1965;[6] however, the capability of rapidly multiplying digital words of the required size was not generally available at that time. For this reason, other basis functions, particularly binary-valued basis functions such as Walsh functions, were being investigated for use in alternatives to the Fourier transform.[7] These binary-valued basis functions were orthogonal, of course, and recognizing that resonance can be viewed in terms of preferred dimensions in an orthogonal function space, a more generalized theory of resonance was defined (see Reference 8, pp. 319 to 320). This work eventually grew into the largest published volume of work on baseband radar.

As the results of these research efforts grew, so did the number of claims for potential improvements in radar performance. Recently, a DARPA panel conducted their own study of the potential value of several radar techniques in the general category of UWB radar. The results of the study panel were published[9] and a brief summary of their findings was also published by the IEEE.[10] The panel's report set aside many of the previous claims; however, there are some areas worthy of further exploration.[9] In the end, the decision to use a UWB waveform will be dependent on whether that approach is more cost-effective than a modified conventional approach.

II. FUNDAMENTAL RADAR PRINCIPLES

The current interest in UWB radar is based on the expectation that this type of radar will provide improvements in one or more of the following areas:

- Target detectability through radar cross-section RCS enhancements and improved clutter suppression
- Target identification through improved measurement resolution
- Reduced cost through employment of high power switches as transmitter sources

To be able to recognize when a UWB waveform does offer unique performance advantages, it is important to understand how radar performance capabilities are related to the radar waveform characteristics. This section presents a discussion of those relationships. Figure 2.1 shows a generic system block diagram for a radar, including the involvement of the target, antennas, and the propagation medium. This section presents the fundamental radar performance measures for each of these required blocks. Having these fundamentals in mind should make it easier to recognize when there is an advantage to using a UWB waveform.

MEASUREMENT RESOLUTION

It can be shown (see Appendix 2A) that the resolution of a range and Doppler measurement is dependent only on the signal bandwidth and duration, respectively, regardless of the design of the actual waveform. Appendix 2A also shows that ambiguities in the measurement of range and Doppler are dependent upon waveform periodicities in the frequency and time domain, respectively, regardless of the waveform design. Hence, questions concerning measurement precision can be addressed strictly in terms of signal bandwidth, duration, and periodicity. The choice of carrier frequency and, hence, the percent bandwidth are irrelevant to the resolution and ambiguity in measuring range and Doppler shift; further, the achievement of a desired resolution does not necessarily require a UWB waveform.

The coupling between the time- and frequency-domain resolutions is commonplace in conventional waveform design. (See Reference 28, p. 126, for a discussion of the radar uncertainty relation.) For example, a longer duration signal will generally have poorer range resolution but better Doppler resolution. Interestingly, when the percent bandwidth is large enough, a coupling arises between the time and frequency resolution capabilities and the angular resolution as well. In Reference 11, the expression for a four-dimensional ambiguity function is developed. The four dimensions are range, velocity (Doppler), azimuth, and elevation. When the signal percent bandwidth is sufficiently small (less than 20%), the four-dimensional function can be separated into the product of two two-dimensional functions: (1) range and Doppler, and (2) azimuth and elevation. Conversely, when the percent

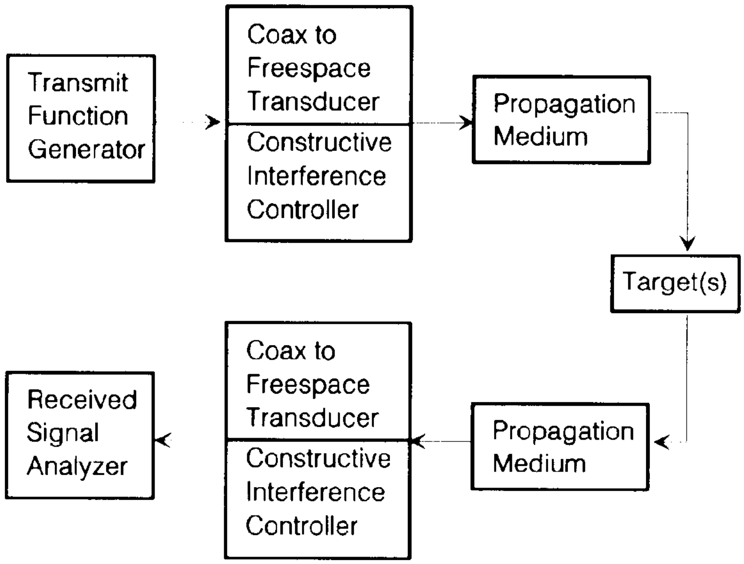


Figure 2.1 Generic radar system block diagram showing components that may be influenced by application of UWB waveforms.

bandwidth is large, there is a coupling among all four resolution coefficients. We will pursue this matter further in the discussion of antennas in Section IV where it will be shown how the time-domain characteristics of the signal have an impact on the angle measurement capabilities of the radar.

TARGET CHARACTERISTICS

Reflectivity Concepts

The reflection of an electromagnetic wave occurs when (1) a wave meets a discontinuity in the characteristic impedance of the propagation medium (e.g., a target object in free space) causing currents to be generated in the object, and (2) the currents flowing in the object cause a signal to be re-radiated. The overall reflection from a complex target is, therefore, dependent on several factors, including

- Electrical size (relative to wavelength) of all reflectors
- Spatial relationships among all reflectors
- Angle of reflection of re-radiated signal

A target can be considered to consist of a collection of standard geometric shapes such as spheres, disks (or plates), or ellipsoids. The reflectivity of each of these reflector types contains three reasonably distinct regions. For example, Figure 2.2 shows the RCS of a sphere. In the figure, we see the *Rayleigh* region where the signal wavelength is much longer than the size of the reflector. In this region, the reflector is very inefficient, and the reflected energy is inversely proportional to the fourth power of the wavelength. In the *resonance* region, the wavelength of the signal is of the same order of magnitude as the size of the reflector. At certain frequencies in this region, the reflector can be caused to resonate, which enhances the reflected signal strength relative to that at other nearby frequencies. In the *optical* region, the wavelength becomes considerably shorter than the reflector and the reflectivity becomes much less sensitive to changes in wavelength. For some target shapes, the reflectivity becomes invariant with frequency in this region.

The perceived reflectivity of a target is also heavily dependent on the angles of incidence and reflection of the radar signal. For example, the reflection from the fuselage of an aircraft can be quite large, but this high reflectivity is not available to the radar unless the receiver is positioned at the angle of reflection which results from the angle of incidence. For a monostatic system (where the transmit and receive antennas are collocated), this condition occurs only when the aircraft fuselage is nearly broadside to the radar beam. Under other conditions, this large return is reflected in a direction away from the receiver.

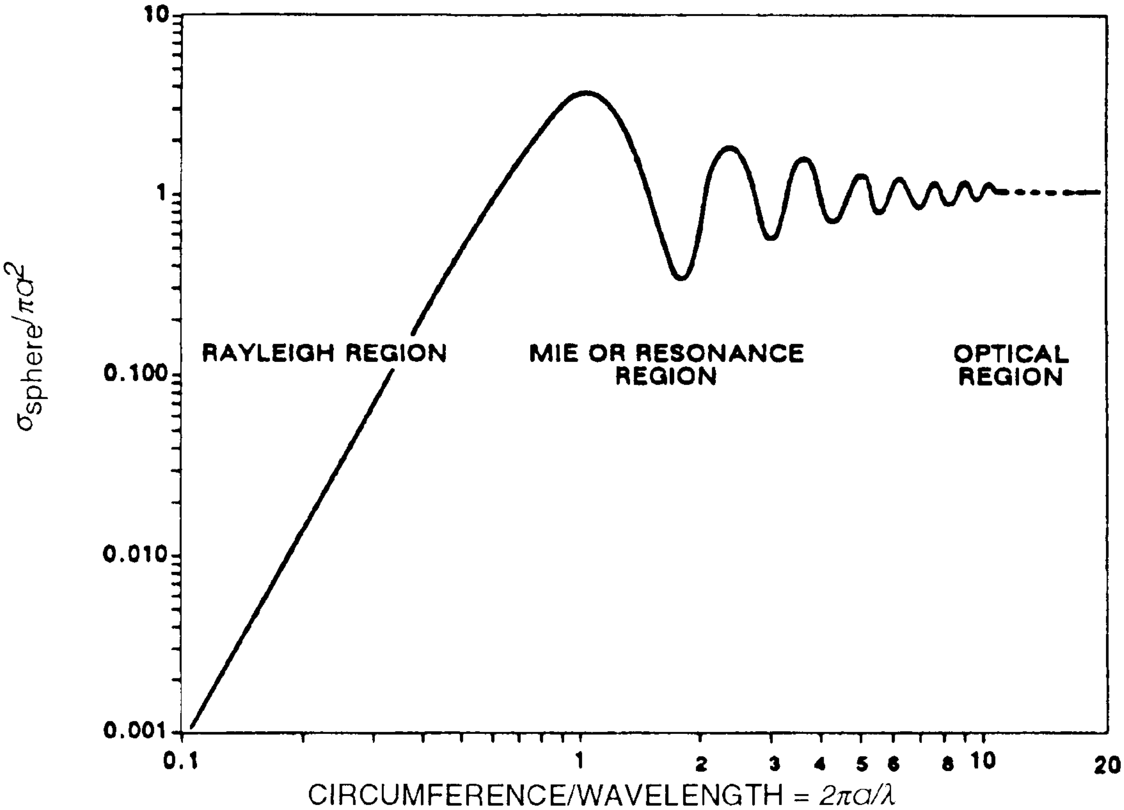


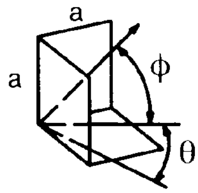
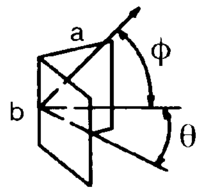
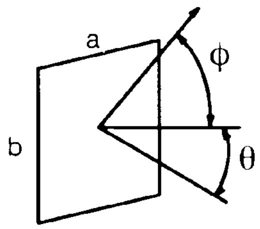
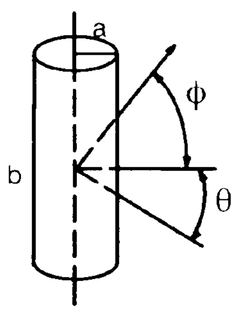
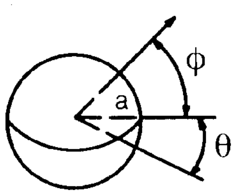
Figure 2.2 Target RCS for sphere vs. circumference/wavelength. (From Skolnik, M. I., *Radar Handbook,* McGraw-Hill, New York, 1970. Copyright 1970 by McGraw Hill. With permission.)

In general, the reflectivity of a target for monostatic systems can be reduced by manipulating the angle of reflection so that even this broadside reflection is not large in the monostatic direction. Two methods for doing this are (1) design in as many curved surfaces as possible and (2) construct the target surface from randomly oriented facets. Table 2.2 presents the mathematical relationships for the RCS of many common reflector shapes. Note that in the last five entries in the table the RCS increases with wavelength λ. The curvature of the target produces a reduction in the monostatic RCS as the frequency increases. Where it is not practical (e.g., for aerodynamic reasons) to employ one of these curved surfaces, it is also possible to reduce the reflectivity by faceting the surface. Faceting produces a surface roughness which makes the reflection more diffuse (i.e., scattered in many directions). An example RCS plot for a notional target that employs these RCS reduction techniques is shown in Figure 2.3. Note that, at high frequencies, the reflectivity is low for most aspects; however, there are also some narrow angular regions where the reflectivity is quite large. Note also that, as the frequency is reduced, the tall spikes become shorter but much wider; also, the low regions become higher.

The relative size and location of the reflectors that make up a complex target also influence its composite reflection. The physical size of a reflector (relative to the signal wavelength) controls the amplitude of the response. The position of a reflector (relative to the others) controls the relative time delay of its reflection. In conventional systems where the wavelengths are such that operation is all in the optical region of the target and the range of wavelengths is small (i.e., small percent bandwidth), the impact of these two target features is seen in the variation in amplitude and phase of the reflected signal. The amplitude fluctuation is generally referred to as *scintillation* and the phase fluctuation is referred to as *glint.*

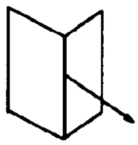
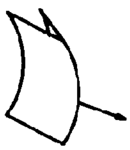
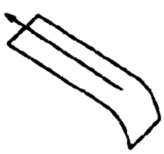
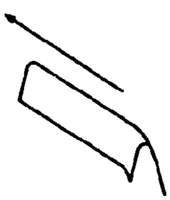
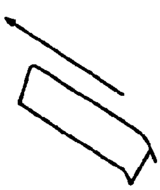
When the wavelength is increased such that operation in the Rayleigh and resonance regions is present and when the percent bandwidth is increased, the target can impose some additional alterations on the illuminating signal. For example, when the signal bandwidth includes the target resonant frequencies, the reflecting surfaces will return energy at the resonant frequencies more strongly than

Table 2.2 Radar Cross Section of Selected Geometrical Shapes

GEOMETRY	TYPE	FREQ. DEP.	SIZE DEP.	FORMULA	REMARKS
	SQUARE TRIHEDRAL CORNER RETRO–REFLECTOR	F2	L4	MAXIMUM $\sigma = \dfrac{12\pi a^4}{\lambda^2}$	STRONGEST RETURN; HIGH RCS DUE TO TRIPLE REFLECTION
	RIGHT DIHEDRAL CORNER REFLECTOR	F2	L4	MAXIMUM $\sigma = \dfrac{8\pi a^2 b^2}{\lambda^2}$	SECOND STRONGEST; HIGH RCS DUE TO DOUBLE REFLECTION, TAPERS OFF GRADUALLY FROM THE MAXIMUM WITH CHANGING θ AND SHARPLY WITH CHANGING φ.
	FLAT PLATE	F2	L4	MAXIMUM $\sigma = \dfrac{4\pi a^2 b^2}{\lambda^2}$ (NORMAL INCIDENCE)	THIRD STRONGEST; HIGH RCS DUE TO DIRECT REFLECTION, DROPS OFF SHARPLY AS INCIDENCE CHANGES FROM NORMAL.
	CYLINDER	F1	L3	MAXIMUM $\sigma = \dfrac{2\pi a b^2}{\lambda^2}$ (NORMAL INCIDENCE)	PREVELENT CAUSE OF STRONG, BROAD RCS OVER VARYING ASPECT (θ), DROPS OFF SHARPLY AS AZIMUTH (φ) CHANGES FROM NORMAL. CAN COMBINE WITH FLAT PLATE TO FORM DIHEDRAL CORNER REFLECTOR.
	SPHERE	F0	L2	MAXIMUM $\sigma = \pi a^2$ (NORMAL INCIDENCE)	PREVALENT CAUSE OF STRONG, BROAD RCS PEAKS OTHER THAN THOSE DUE TO LARGE OPENINGS IN TARGET BODY. ENERGY DEFOCUSED IN TWO DIRECTIONS.

Note: Adapted and revised from Knott, E.F. et al., *Radar Cross Section,* Artech House, Dedham, MA, 1985, 178–179.

Table 2.2 Radar Cross Section of Selected Geometrical Shapes (continued)

GEOMETRY	TYPE	FREQ. DEP.	SIZE DEP.	FORMULA	REMARKS
	STRAIGHT EDGE NORMAL INCIDENCE	F^0	L^2	$f(\theta, \theta_{int})L^2$ θ = ASPECT θ_{int} = INTERIOR DIHEDRAL ANGLE BETWEEN FACES MEETING AT EDGE	LIMITING CASE OF 2-DIMENSIONAL CURVED PLATE MECHANISM AS RADIUS SHRINKS TO 0. PREVALENT CAUSE OF STRONG, NARROW RCS PEAKS FROM SUPER-SONIC AIRCRAFT.
	CURVED EDGE NORMAL INCIDENCE	F^{-1}	L^1	$f(\theta, \theta_{int})\frac{a\lambda}{2}$ $a \geq \lambda$	LIMITING CASE OF 3-DIMENSIONAL CURVED PLATE MECHANISM AS PRINCIPAL RADIUS SHRINKS TO 0. THE FUNCTION f IS THE SAME AS FOR SHAPE ABOVE.
	APEX	F^{-2}	L^0	$\lambda^2 g(a, \beta, \theta, \phi)$ a, β = INTERIOR ANGLES OF TIP θ, ϕ = ASPECT ANGLES	LIMITING CASE OF PREVIOUS MECHANISM AS a SHRINKS TO 0. FOR a = β, THE TIP IS THAT OF A CONE. FOR a = 0, THE TIP IS THE CORNER OF A THIN SHEET, OR FIN.
	DISCONTINUITY OF CURVATURE ALONG A STRAIGHT LINE, NORMAL INCIDENCE	F^{-2}	L^0	$\frac{\lambda^2}{64\pi^3}\left(\frac{L}{a}\right)^2\left(1+\left(\frac{dy}{dx}\right)^2\right)^{-\frac{3}{2}}$ $a \geq \lambda$ $\frac{1}{a}$ = JUMP IN RECIPROCAL OF THE RADIUS OF CURVATURE $\frac{dy}{dx}$ = SLOPE OF SURFACE w.r.t. INCIDENT RAY	STRONGEST OF AN INFINITE SEQUENCE OF DISCONTINUITIES. VERY WEAK MECHANISM WHICH TOGETHER WITH THE SHAPE ABOVE SHARES DOMINANCE OF NOSE-ON RCS OF CONE SPHERE.
	DISCONTINUITY OF CURVATURE OF A CURVED EDGE	F^{-3}	L^{-1}	$f(\theta,\phi)\frac{\lambda^3 b}{a^2}\left(1+\left(\frac{dy}{dx}\right)^2\right)^{-\frac{3}{2}}$ $f(\theta,\phi)$ = FUNCTION OF ASPECT b = RADIUS OF EDGE > λ	IMPORTANT MECHANISM FOR TRAVELING WAVE BACKSCATTER WHERE RCS OF DISCONTINUITY IS AUGMENTED BY GAIN OF TRAVELING WAVE STRUCTURE. DEPENDENCES ARE BASED ON DIMENSIONAL CONSIDERATIONS.
	DISCONTINUITY OF CURVATURE ALONG AN EDGE	F^{-4}	L^{-2}	$g(\theta,\phi)\lambda\left(\frac{1}{a}\right)^2\left(1+\left(\frac{dy}{dx}\right)^2\right)^{-\frac{3}{2}}$ $g(\theta,\phi)$ = FUNCTION OF ASPECT	IMPORTANT MECHANISM FOR TRAVELING WAVE BACKSCATTER WHERE RCS OF DISCONTINUITY IS AUGMENTED BY GAIN OF TRAVELING WAVE STRUCTURE. DEPENDENCES ARE BASED ON DIMENSIONAL CONSIDERATIONS.

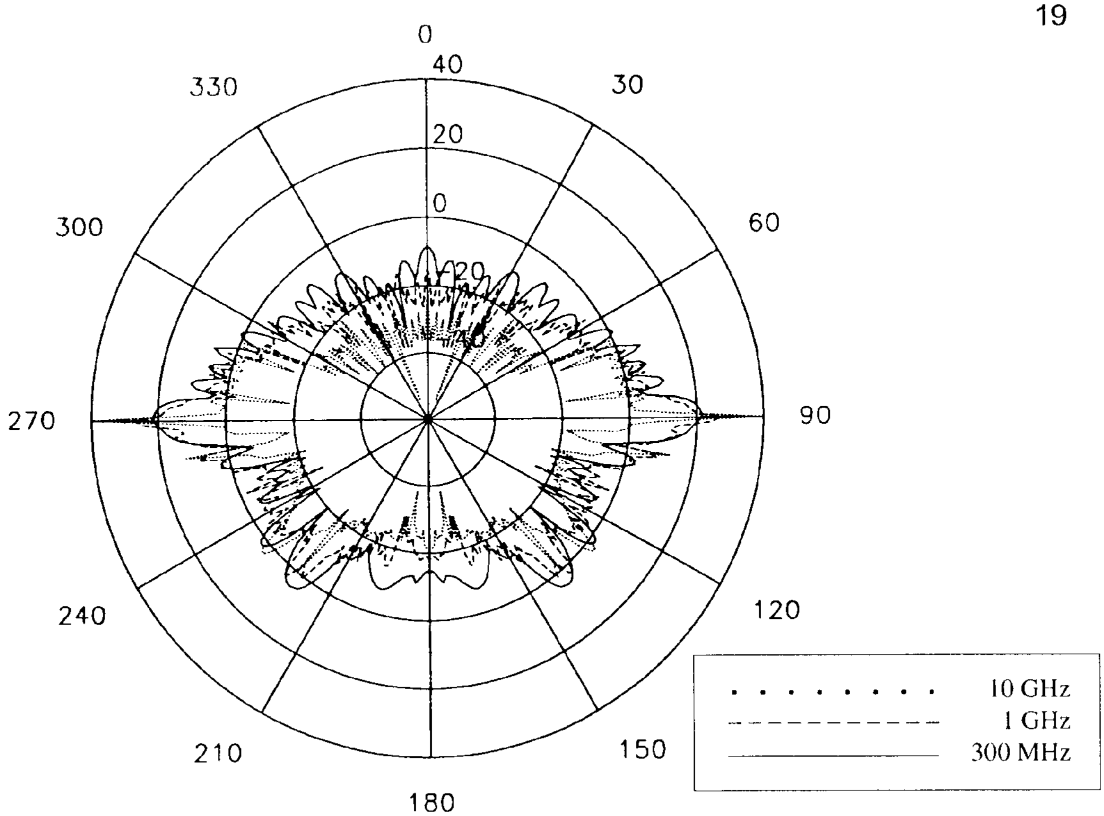


Figure 2.3 Example RCS from notional low-RCS target. At common radar wavelength, this target exhibits very strong reflectivity over very narrow angular sectors. Note, however, as the wavelength is increased that reflectivity is reduced, while the angular region is increased.

than at others. Also, since these resonant surfaces are likely to be at different physical locations on the target, each of the resonant frequency components will receive a different time delay before retransmission. This frequency-selective distribution in time delay means that the target becomes frequency dispersive. In general, dispersion in the signal frequency components produces distortion in the time domain characteristics of the signal. This distortion may be usable for identification of the target, but it may also make the signal less distinguishable from the accompanying clutter signal.

Target Identification

Target identification can be accomplished by radar in several ways. The two methods that may require the use of a UWB waveform are addressed in this section. The first method illuminates the target with a sufficiently broadband signal to be able to estimate its resonant frequencies. These estimates can be used as a template for identification purposes. The other approach is to use sufficiently precise range resolution so that all of the major scatterers of the target can be resolved individually, thereby generating something of an image of the target.

These two methods have very different demands on the waveform design. We will consider the target resonance method first, based on Reference 14. Because the natural body resonances of the target are related to its physical size, this method requires that the minimum frequency of the signal be such that its wavelength is 2 to 4 times the length of the resonant scatterer of the target. The entries in the center column of Table 2.3 show the implications of this requirement. Higher frequencies in the waveform will excite higher modes of these same resonances, but the fundamental resonant frequency generally provides the strongest response. To enhance the reliability of target identification, it is desirable to excite as many resonances as possible, and it is therefore necessary to use a signal of sufficient bandwidth to do so. From Table 2.3 it would seem unlikely that the upper frequency limit for this purpose would be much above 100 MHz, because the smallest resonators on a complex target will probably not be smaller than a small missile. Nonetheless, even if the signal must only cover 2 to 4 MHz, this is still a large (66%) percent bandwidth signal.

Table 2.3 Waveform Requirements for Target Identification[a]

Target Type	Minimum Frequency for Resonance (MHz)	Minimum Bandwidth for Imaging (MHz)
Bomber[b]	2–4	20
Fighter[c]	5–10	50
Small missile[c]	50–100	500

[a] All waveform requirements based on composite values taken from Table 2.4.

[b] Waveform requirements for the bomber are based upon the wingspan dimension.

[c] Waveform requirements for the fighter and the missile are based upon the fuselage dimension.

It is important to point out an operational limitation of this technique. It may not be feasible to produce the signal-to-noise ratio (SNR) necessary to obtain sufficiently accurate measurements for this technique to work.[15] The resonance extraction algorithms will always provide estimates of resonances based on the measured data, but these estimates may not be relevant to the target features.

There are some ways in which a UWB signal can be avoided. If the target responds in a linear manner, it does not have to be illuminated with all frequencies in the band of suspected resonances simultaneously; a swept waveform could be used. Also, if the approximate location of the resonant frequencies is known, a multifrequency waveform could be used. However, if good range resolution is also needed (e.g., to isolate the target response from clutter signals), the required bandwidth, in combination with the low center frequency needed for resonance excitation, will result in a UWB waveform. Since good range resolution would almost certainly be required, this method of target identification would, in general, always require a UWB waveform.

The second target identification method produces a one-dimensional "image" of the target (which is often referred to as *range profiling*). This technique requires the signal bandwidth to be at least equal to the value in the third column in Table 2.3. However, the spectral region in which the signal is located is not important to this target identification capability. Unless a low center frequency is needed for some other reason, then this target-imaging capability would not require a UWB waveform. Table 2.4 provides dimensions of typical targets.

Clutter Rejection

Part of the original interest in UWB waveforms was the anticipation that they might provide some improved capability to separate small targets from clutter. This section describes three methods for clutter rejection/target enhancement to show where a UWB waveform may offer unique advantages: resolution cell size reduction, increased wavelength, and waveform polarity discrimination.

The usual method for achieving clutter rejection is to isolate the clutter reflections from the target in one or more of the four measurement domains (range, Doppler, azimuth, and elevation). However, reducing the center frequency (as is done in some UWB waveforms) may also reduce clutter reflection from foliage. Also, a baseband type of UWB waveform may provide a unique capability of isolating a certain class of targets (high permeability) from clutter. The method for doing this is described later in this section.

Clutter Cell Size Reduction

Generally, any steps taken to reduce the size of the resolution cell in any of the measurement domains will provide an improvement in clutter rejection. Note that if the radar makes ambiguous measurements of range or velocity, the ambiguous clutter returns may diminish the utility of a reduced resolution cell.

Table 2.4 Dimensions of Typical Targets[16,17]

Target Type	Wingspan (ft)	Fuselage Length (ft)
B-52 bomber	185	160
F-15 fighter	42	63
F-16 fighter	32	49
737-400 transport	93	100
Tomahawk missile	9	21
Standard ARM	3	15

Also, note that if the cell size is made so small that the target return is distributed across several resolution cells, a degradation in detection sensitivity will occur. In any case, range or Doppler resolution is only dependent on signal bandwidth or duration, respectively (see Appendix 2A).

Angle resolution can be used to suppress clutter by reducing the angle cell size. Figure 2.4A shows the relationship between the illuminated clutter cell size and the radar antenna beamwidth. In Figure 2.4B the resolution provided by the waveform exceeds that provided by the antenna. For small percent bandwidth waveforms, angle resolution is completely determined by the antenna design and the center frequency. Ultra-wideband waveforms may offer unique advantages here because the resolution properties of range and velocity are coupled to the resolution in angle (see Section II and also Reference 11, p. 2096). With this coupling it may be possible, for example, to use excess range resolution to improve azimuth or elevation resolution.

The rejection of clutter outside the resolution cell is not perfect — this is true for all of the measurement dimensions: range, Doppler, and angle. This imperfection is due to additional smaller cells, called sidelobes, which are displaced from the desired cell. Examples of sidelobes are shown in the ambiguity function in Figure 2.5. The center lobe in the function represents the resolving capability of the waveform. The additional lobes away from the center are sidelobes: regions of imperfect clutter suppression. In small percent bandwidth systems, the size and location of the range and Doppler sidelobes are controlled by the waveform parameters; the angle sidelobes are controlled by the antenna design. In UWB systems, the angle sidelobes are influenced by both the antenna design and the waveform design. Hence, UWB waveforms may offer some unique advantages for sidelobe clutter rejection.

Increased Wavelength

Clutter rejection also may be improved by taking advantage of the fact that longer wavelengths are better able to penetrate foliage and natural vegetation.[18] A fairly extensive source of measured clutter reflectivity for narrowband signals is provided in Reference 19. These data show that for some types of clutter the reflectivity is inversely proportional to wavelength. For other types, it is actually directly proportional to wavelength, and in others there is no monotonic behavior with wavelength. Hence, this data would not suggest an optimal frequency for clutter rejection.

However, some other data obtained from measurements made with a UWB waveform (Reference 20, Appendix F) show that the reflection at these frequencies for heavily forested terrain can be dominated by the reflection from the trunks of trees, not the foliage. It could also turn out that the dominant clutter source may be from the ground beneath the foliage, rather than the foliage itself. If this turns out to be the case, then whatever attenuation is provided by the foliage will actually tend to suppress the return from the ground reflection.

From these data, it seems that the performance of a system which must detect targets within or above foliage could probably be improved by moving the signal to a lower center frequency. Choosing a low center frequency (e.g., at UHF or below) for good foliage rejection while maintaining good range resolution requires a UWB signal.

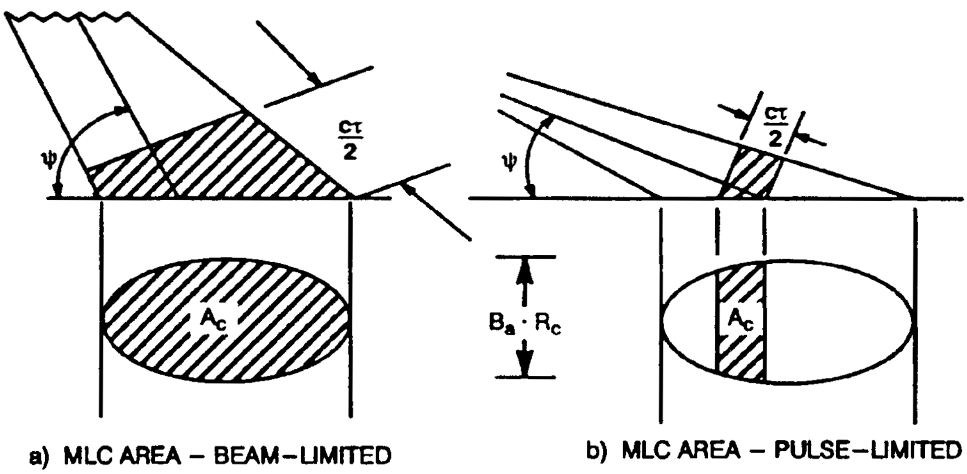


Figure 2.4 Radar resolution effects on clutter cell size. In (A) the cell size in the elevation dimension is limited by the beamwidth of the antenna. In (B) the cell size is limited by the resolution of the waveform.

Waveform Polarity Discrimination

Another possible method for clutter suppression which may be possible for baseband UWB waveforms (see Baseband Waveform in Section III) exploits the reflectivity property of the target itself. The reflection coefficient at a boundary between two media, ρ, is (Reference 21, p. 151)

$$\rho = \frac{E_r}{E_i} = \frac{\eta_2 - \eta_1}{\eta_2 + \eta_1} \tag{2.1}$$

where (Reference 21, p. 128)

$$\eta_i = \sqrt{\frac{j\omega\mu_i}{\sigma_i + j\omega\varepsilon_i}}\ , \quad i = 1, 2 \tag{2.2}$$

and

η_i = characteristic impedance in ohms
σ = the medium conductivity in mho/meter
μ = magnetic permeability, in Henry/meter
ε = electric permittivity or dielectric constant, in Farad/meter

When the reflecting medium is a conductor, σ is large. In a perfect conductor it is infinite, and then η for that medium is zero. Suppose that in Equation 2.1, medium 1 is free space and medium 2 is a perfect conductor. Under these conditions, ρ becomes –1, which means that the orientation of the E-field vector is reversed on reflection.

However, note from Equation 2.1, that it is possible to prevent the orientation of the E-field vector from reversing by making η_2 larger than η_1. Recognizing that

$$\eta_1 = \sqrt{\frac{\mu_0}{\varepsilon_0}}$$

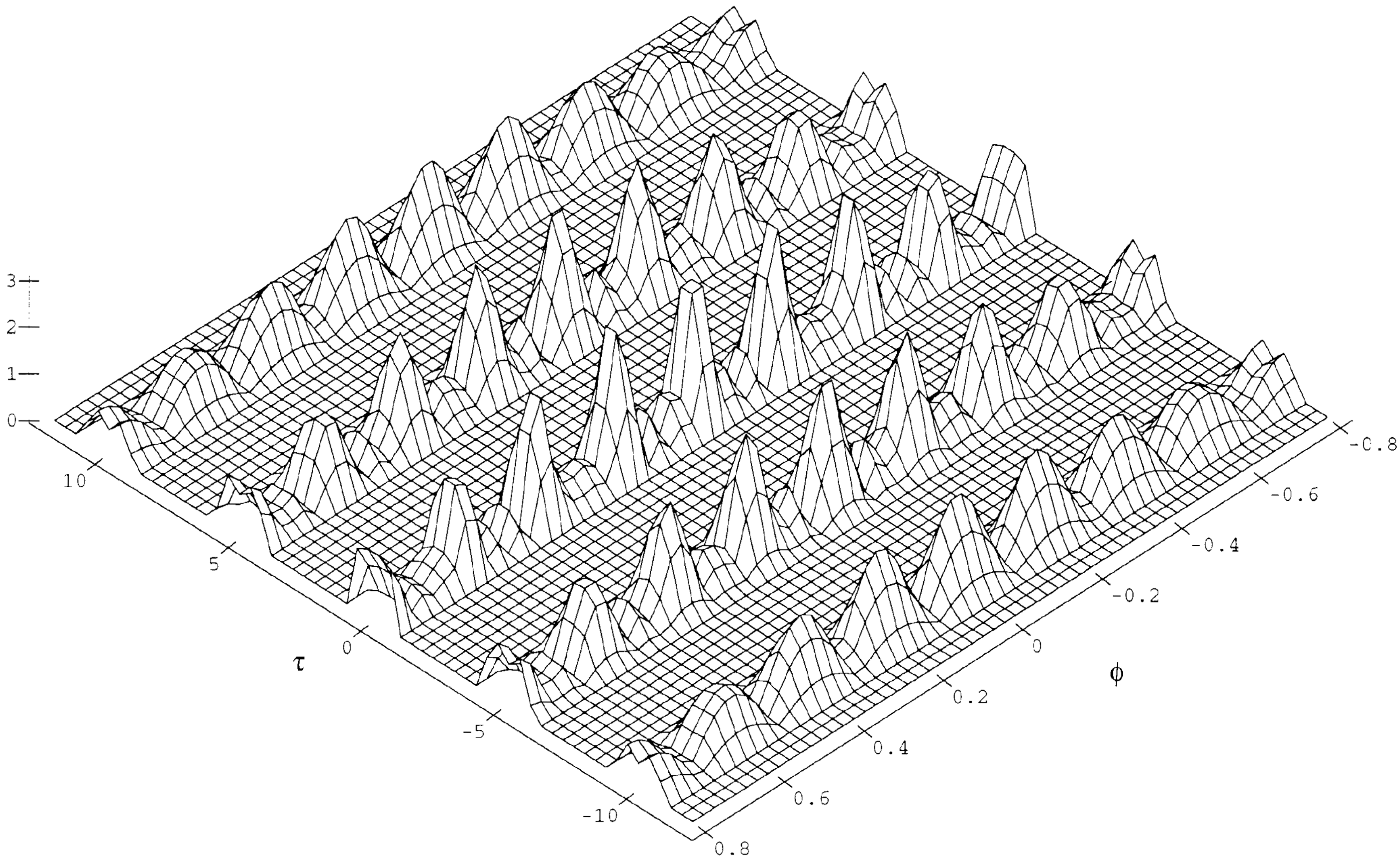


Figure 2.5 Example of ambiguity function showing time and frequency sidelobes. The center lobe in the function represents the resolving capability of the waveform. The additional lobes away from the center are sidelobes — regions of imperfect clutter suppression. In small percent-bandwidth systems, the size and location of the range and Doppler sidelobes are controlled by the waveform parameters, and the angle sidelobes are controlled by the antenna design. In UWB systems, the angle sidelobes are influenced by both the antenna design and the waveform design. Hence, UWB waveforms may offer some unique advantages for sidelobe clutter rejection.

we see by comparing to Equation 2.2 that for $\eta_2 > \eta_1$ requires that

$$\sqrt{\frac{\mu_2/\varepsilon_2}{\mu_0/\varepsilon_0}} \gg 1 \tag{2.3}$$

In most natural nonmagnetic materials, $\mu \approx \mu_0$ and therefore for Equation 2.3 to hold would require the material to have a relative dielectric constant less than one (which is impossible). However, for high permeability materials, it may be possible for Equation 2.3 to hold and therefore produce a reflection without reversing the orientation of the E-field vector.

The type of material with these properties is referred to as magnetic RAM (radar-absorbing material) and has the advantage of a high index of refraction $\sqrt{\mu\varepsilon}$ at lower frequencies. This property causes the material to exhibit an electrical thickness which is many times the actual thickness. For example, one nickel-zinc ferrite has an index of refraction above 50 at 100 MHz, making the material appear to be 50 times as thick electrically as it actually is (Reference 13, p. 255). This type of material, then, makes an effective radar signal absorber for longer wavelength signals. Conversely, it provides the capability for using much thinner coatings to absorb higher frequencies. However, the interesting feature of this type of material in the present discussion is that it may not produce a reversal of the E-field vector, even if σ_2 is appreciable.

The significance of this property for baseband signals is shown in Figure 2.6. The waveforms shown there are idealized examples of the signals in the transmitter and receiver. The actual appearance of the propagating signals will depend on how well the antenna(s) match the propagation medium and the transmitter/receiver and on the dispersiveness of the target. The idealized waveforms are used here to simplify the discussion of the clutter suppression effect. Corruption of the waveform due to antenna and/or target effects may diminish the utility of this technique. On the other hand, however, it may be possible to calibrate out the effect of the antenna(s), and the distortion caused by the target might be usable as a target ID method. More experimentation is needed to fully evaluate these possibilities.

From Figure 2.6, we see that the orientation of the E-field vector is affected in the same way for both carrier-based and baseband signals. When a propagating wave strikes a reflector with low permeability (which is representative of most targets and clutter), the polarity of the wave is inverted as shown in Figure 2.6C. When a wave strikes a highly permeable material, the E-field orientation of the reflected wave is unchanged. For the carrier-based waveform, it is equivalent to say that the carrier experiences a 180° phase shift upon reflection from low permeability targets. In certain radar systems, this phase shift (or lack thereof) is detectable. However, it is not possible to determine whether the phase shift was caused by a reflection from a low permeability target or by a half-wavelength change in the round trip target distance. With the baseband waveform, there is no ambiguity between polarity reversal and time delay. Hence, if a threshold detector was placed at an appropriate voltage as shown in Figure 2.6B, it may be possible to detect returns from high permeability targets without interference from those of other targets or clutter with the opposite polarity.

The discussion above contains the implicit assumption of a point target which is clear of any multipath. A complex target may produce multiple internal reflections before re-radiating the signal. If the high permeability target produces no reversal of E-field polarity, the same E-field orientation will exist in the reflected signal, no matter how many internal reflections occur. However, it is possible for a low permeability target, through multiple internal reflections, to also produce a return signal whose E-field orientation is unchanged from that on transmit. Hence, for complex targets, the situation is that multiple internal reflections would not cause the high permeability target to be missed, but may improve the detectability of a low permeability target and cause it to be mistaken for a high permeability one.

It is also true that the orientation of the reflected E-field is generally affected by the type of material, the signal wavelength, and the signal angle of incidence. Also, at RF frequencies the values of ϵ and μ may become complex (as opposed to purely real), in which case, the polarity of the reflected

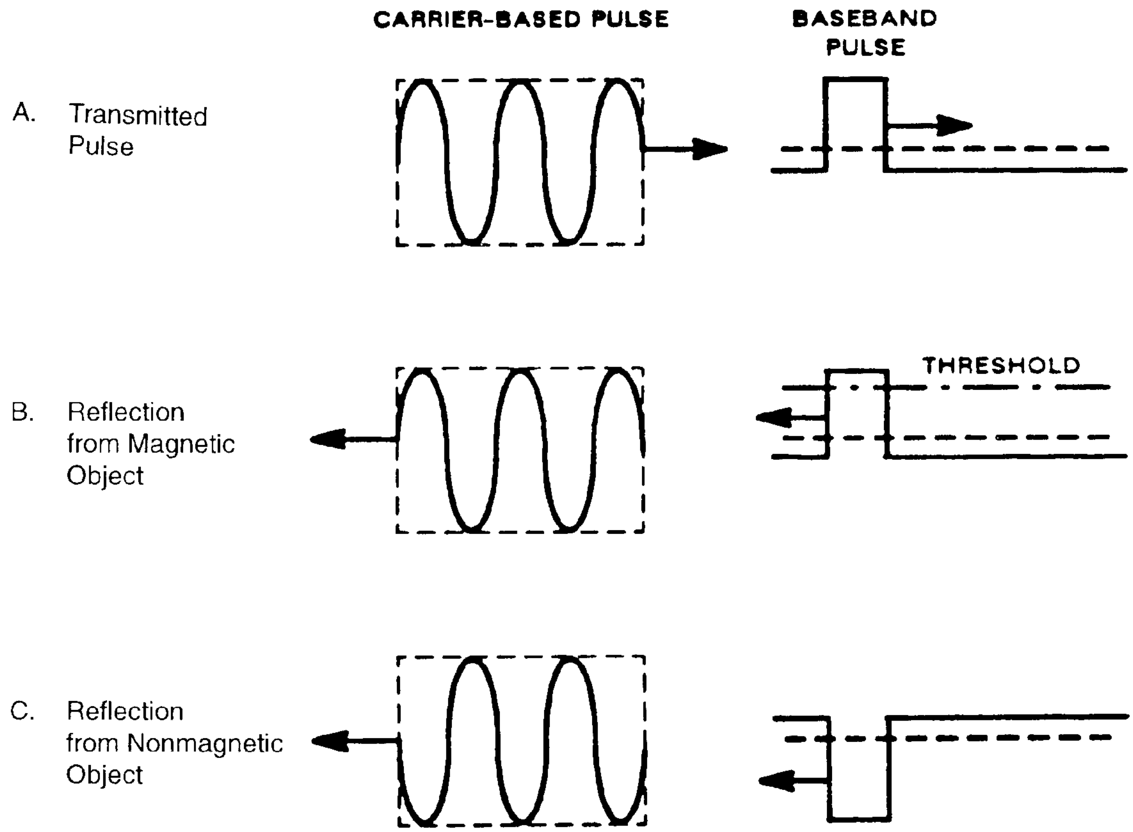


Figure 2.6 Effect of reflection on E-field orientation angle.

wave would become more difficult to predict. Hence, further work is needed to determine if this clutter rejection technique can be made reliable.

Generally, sources of multipath will also be low permeability objects and therefore will produce an E-field orientation reversal for horizontally polarized waveforms. Hence, an odd-bounce multipath reflection will cause an orientation reversal from both types of targets. If the multipath signal arrives coincidentally with the direct return, it will subtract from the direct return and can reduce the detectability of the target. If it arrives after the direct return, it will be ignored by the threshold setting.

Summary on Clutter Suppression Techniques

We have examined three fundamental methods for clutter suppression. Reduction of the resolution cell size is possible by both conventional and UWB systems. However, exploitation of the coupling between angle and time/frequency resolution is unique to UWB waveforms. Exploitation of this coupling may offer some unique improvement in clutter suppression. Low center frequency is another method for not only reducing clutter reflection, but also for improving foliage penetration. However, a lower center frequency does not automatically result in a UWB waveform. We also looked at the method of waveform polarity discrimination for improving the detection of high permeability targets in clutter. This method is unique to UWB waveforms because it requires the ability to distinguish the alteration in waveform polarity arising from a half-wavelength difference in the time of arrival from one produced by the reversal upon reflection.

PROPAGATION MEDIUM

Depending on the frequency, the propagation medium contributes significant noise, attenuation/absorption, and dispersion to the signal. In radar design, a tradeoff must be performed to decide where in the frequency spectrum to locate the signal. In conventional design, this decision is driven by the required bandwidth and the requirement to keep a small percent bandwidth to take advantage of sinusoidal resonance effects. A bandwidth of 1% is typical for this type of design (the figure could go as high as 10%); for this reason, high precision radars are always found at higher microwave frequencies (above 1 GHz).

Figure 2.7 shows a combination of figures from Reference 22, p. 126, and Reference 23, p. 472. The radiation injected into the receiver by galactic events is represented as a noise and is characterized by a temperature so that the analysis of its effect can be handled in the same way as receiver noise. Note that for frequencies below 2 GHz, the galactic noise temperature, and hence the noise power, is not constant over frequency. If a radar signal with significant bandwidth was located in this frequency range, it would experience a nonwhite noise spectrum. Because classical detection theory is based on the assumption of *white* Gaussian noise, the common expectations for required SNR may not be valid for a system operating in this range. There are, of course, systems which do operate in this region, but their bandwidth is small enough so that the noise is essentially flat over the bandwidth. The issues concerning noise and detection statistics will be discussed in more detail in Chapter 10.

On the upper end of the frequency spectrum are curves showing the additional attenuation due to moisture content in the propagation medium. This is an important consideration for systems which must operate in all weather conditions. For example, if the radar system had a path loss margin of 10 dB and was expected to detect targets at ranges up to 100 km in heavy rain, then the spectrum of the signal should not be allowed to extend beyond 4 GHz.

Overall, it would appear that the more desirable region in which to operate is between 500 MHz and 10 GHz. However, if this is not possible for some reason (e.g., if resonance excitation is desired), additional transmit energy would be required. When sufficient energy can be produced, the additional noise and/or attenuation from the propagation medium could aid in hiding the radar signal from intercept receivers.

INTERFERENCE SUSCEPTIBILITY

We will be concerned here with intentional interference to the UWB system; the considerations for natural interference were discussed in the previous section. In general, there will be a limitation on the total power transmitted by the interfering source. If the radar can force the jammer to spread its signal over a wider bandwidth through the use of a wideband signal, then the power density of the interfering signal will be reduced. The radar can make use of this property to diminish the susceptibility to noise inputs.

In the spread-spectrum community, this noise-spreading property of broadband signals was welcomed until the downside was pointed out: the processing performed by the receiver to recapture the desired signal is such that any undesired signal appearing in the receiver passband will be spread in bandwidth until it is noiselike; therefore, a CW tone will have as much capability of disrupting the transmission as a noise signal matched to the radar signal bandwidth. Thus, the wideband radar system exhibits no more resistance to noise interference than a narrowband system, unless the radar is able to use a part of the spectrum that is not accessible to the jammer.

In many instances, though, the jammer tries to produce false replicas of the desired signal. Here, the more complicated the radar signal is, the more difficult it will be for the jammer to be successful. In many cases, it is necessary for the jammer to obtain a sample of the radar signal in order to produce a reasonably faithful replica. A wideband radar system will force the intercept receiver (which is used to collect this sample) to also be wideband. In general, the intercept receiver is unable to process the signal as efficiently as the radar and, as a result, it is more likely that the wideband radar signal may go undetected by the interceptor.

Currently, wideband systems using more conventional approaches, such as spread-spectrum modulation and frequency hopping, pose particular problems for intercept receivers, especially in dense signal environments where other signals may interfere with the detection of the broadband signals. The UWB waveforms have the potential to impose the same (or worse) problem on the interceptor. In addition to the impact of wide bandwidth, the use of a low carrier frequency may further degrade interceptability. This degradation occurs due to the narrowband design of intercept receivers for this frequency region. This degradation in interceptability will persist until the interceptor has implemented a receiver similar to that used in the radar itself.

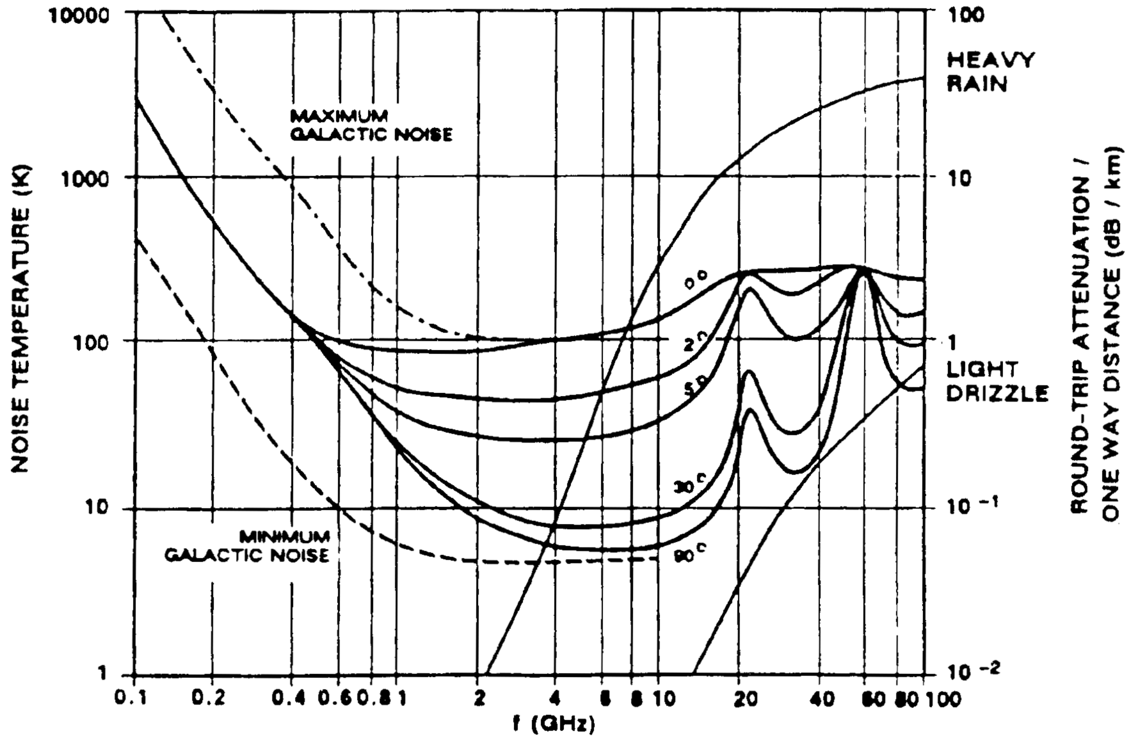


Figure 2.7 Variation in external noise and attenuation with signal frequency. The family of curves bounded by the dashed lines are curves of noise temperature. Each curve corresponds to a different elevation angle. The curves advancing to the right from the abscissa scale are boundary curves for rain attenuation for the two rain conditions indicated at the right of the graph. (Adapted from References 22 and 23.)

It is also important to consider the question of interference to other systems caused by the operation of the UWB system. In the frequency range we have been discussing (i.e., 10 GHz and below) there are no unused regions of the EM spectrum. The question of how compatible the UWB system (which uses much more bandwidth than any other user in the spectrum nearby) will be with other current users is one that must be addressed before any UWB systems will be permitted to operate. This question is complex, however, because

- the radiation will be very directive
- polarization could be selected to minimize interference to and from other users
- the duty cycle of the UWB signal for the simple impulse design is very low
- the narrowband nature of systems operating at these frequencies may suppress most of the broadband EMI from the UWB system

The issue of electromagnetic compatibility will be discussed in more detail in Chapter 10; however, it is reasonable to expect that each case must be assessed and tested individually.

DETECTION

Detection in radar depends only upon the amount of transmitted signal energy reflected from the desired target that is available to the receiver, relative to that which is received from all other sources. For all of the waveform types described in Section III, it is possible in principle for each to be designed so that all have the same amount of transmitted energy (although currently the energy available from impulse-generating devices is limited due to the limitations on transmit waveform duty cycle). However, even under the condition of equal energy, there may be differences in the detection capability of each waveform type due to (1) differences in the effects of the propagation medium and clutter and (2) the distortions produced by the radiation and reception processes of the antenna(s) and the target.

In narrowband systems, the effects of the propagation medium, clutter, antennas, and the target are essentially constant for all frequencies in the radar waveform. In a UWB system, the behavior of these factors may vary across the larger operating bandwidth. For example, as seen in the discussion on Propagation Medium earlier in this section, the effects of the propagation medium and galactic noise may vary considerably. In both conventional and UWB systems, the signal bandwidth affects the clutter and interference rejection capability. However, the bandwidth in a UWB system also influences the angle resolution capability, which may provide additional amounts of clutter and interference rejection capability. Thus the assumption that the background is stationary, white, and Gaussian may not be supportable.

The optimal receiver captures all of the available energy reflected from the target and uses it to produce the measure used to make the detection decision. The UWB signal will experience changes upon transmission, reflection, and reception. From the conventional perspective, these changes would be looked upon as distortions—distortions, which would require additional complexity in the receiver if it were to collect all of the available energy; but, as mentioned earlier, these distortions may be valuable for target identification. Hence, as is the case for conventional systems, it may not be practical to obtain maximum detection range and target identification capability at the same time. Thus, in general, the optimal receiver will be application-dependent and will likely be inseparable from the antenna. Detection will be accomplished by integrating over the time and spatial domains simultaneously. For these reasons, it is important to verify the applicability of existing detection theories before applying them to UWB problems.

III. CLASSIFICATION OF RADAR WAVEFORMS

Ultra-wideband waveforms may, at first, appear to be fundamentally different from more conventional radar and communication waveforms. In fact, however, they are easily grouped into a continuum of waveform classes, which include signals from continuous CW signals through impulse waveforms. This section presents an overview of the various waveform classes to show the location of UWB waveforms in this continuum.

In general, the continuum of waveform types from baseband (largest percent bandwidth) to narrowband (small percent bandwidth) contains three recognizable classes: baseband, monocycle, and polycycle as shown in Figure 2.8. Each of these types will be discussed in more detail below. The frequency spectrum for each of these pulses is presented in Figure 2.9, and the cumulative power spectrum for each of these waveforms is presented in Figure 2.10. The pulse width for the baseband and monocycle pulses was set to 3.3 ns. The widths of the polycycle pulses are 13.2 and 16.5 ns, while the center frequency remained at 300 MHz. These pulse widths correspond to 25 and 20% bandwidth, respectively.

BASEBAND WAVEFORM

The first class of waveform shown in Figure 2.8 is the baseband waveform. It has no carrier but has average value zero to satisfy the conditions for radiation. The effect of setting the average value to zero is visible in Figure 2.9. The spectrum of the baseband pulse is very similar to a unipolar pulse except that the spectral amplitude falls rapidly near 0 Hz. The ripples on top of the spectral lobes are due to the duty cycle of the waveform, which for this spectrum was set to 0.1. Note from Figure 2.10 that 90% of the energy of this waveform falls below 200 MHz.

The "negative" level between pulses seen in Figure 2.8 is a nuisance, because it can suppress other weaker returns interspersed between the larger returns. A way to suppress this negative level is to use a pulse type, which itself has a zero average value. This can be approximated by using coding schemes similar to those used in spread-spectrum signaling.[25] The result is a concatenation of several pulses with opposite polarity in pseudorandom order. With a large number of concatenated pulses, this composite pulse can be made to have an average value arbitrarily close to zero, thus eliminating the need for a DC offset between pulses to produce an average value of zero.

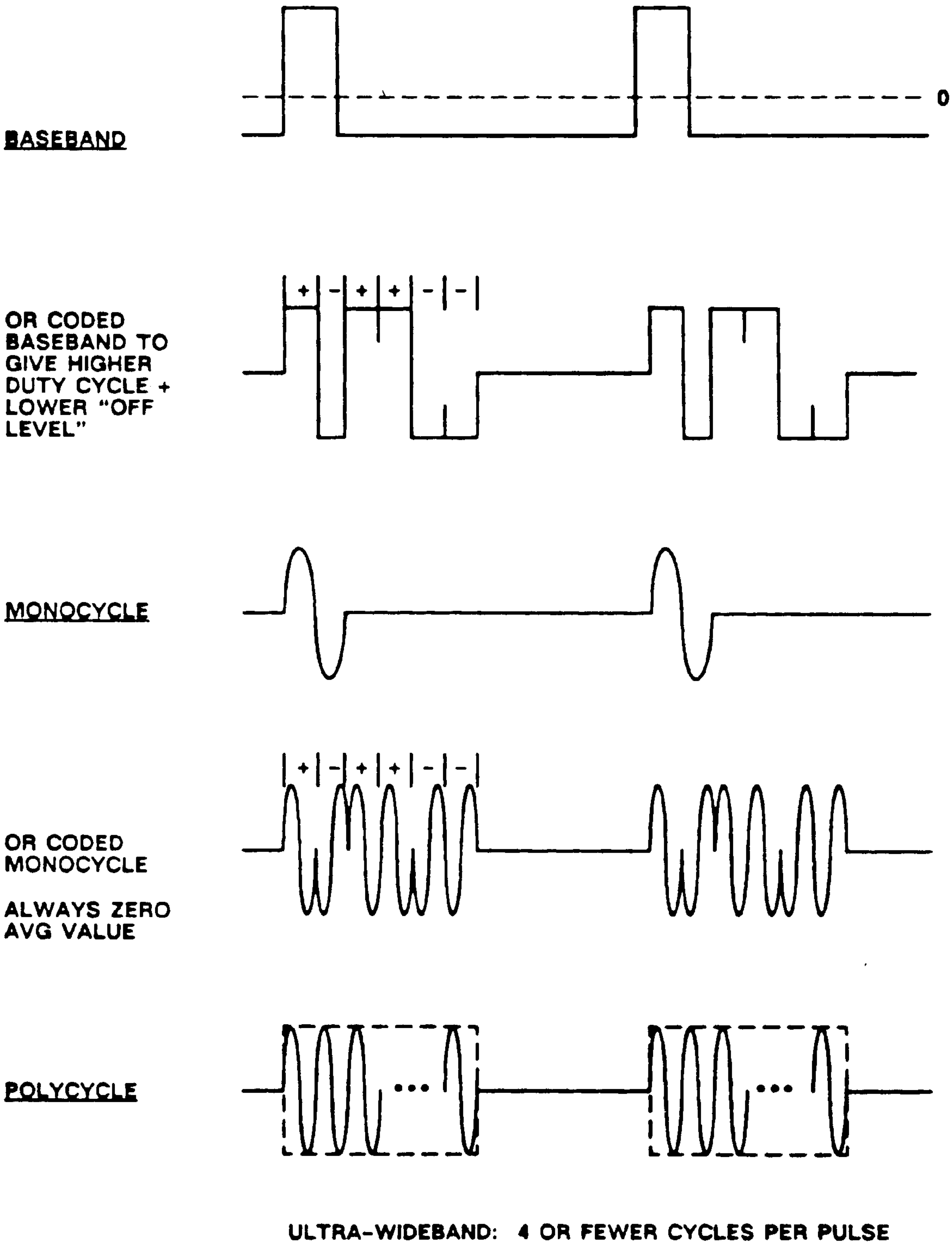


Figure 2.8 Continuum of waveform types.

In addition to providing a zero average value within one pulse, coding can also be used to provide a longer pulse, which will increase the duty cycle and therefore the average power in the waveform. The ability to get more energy into the signal in this manner can be very important in applications where the required peak power cannot be produced, or is undesirable for reasons of electromagnetic interference (EMI), equipment reliability, or safety.

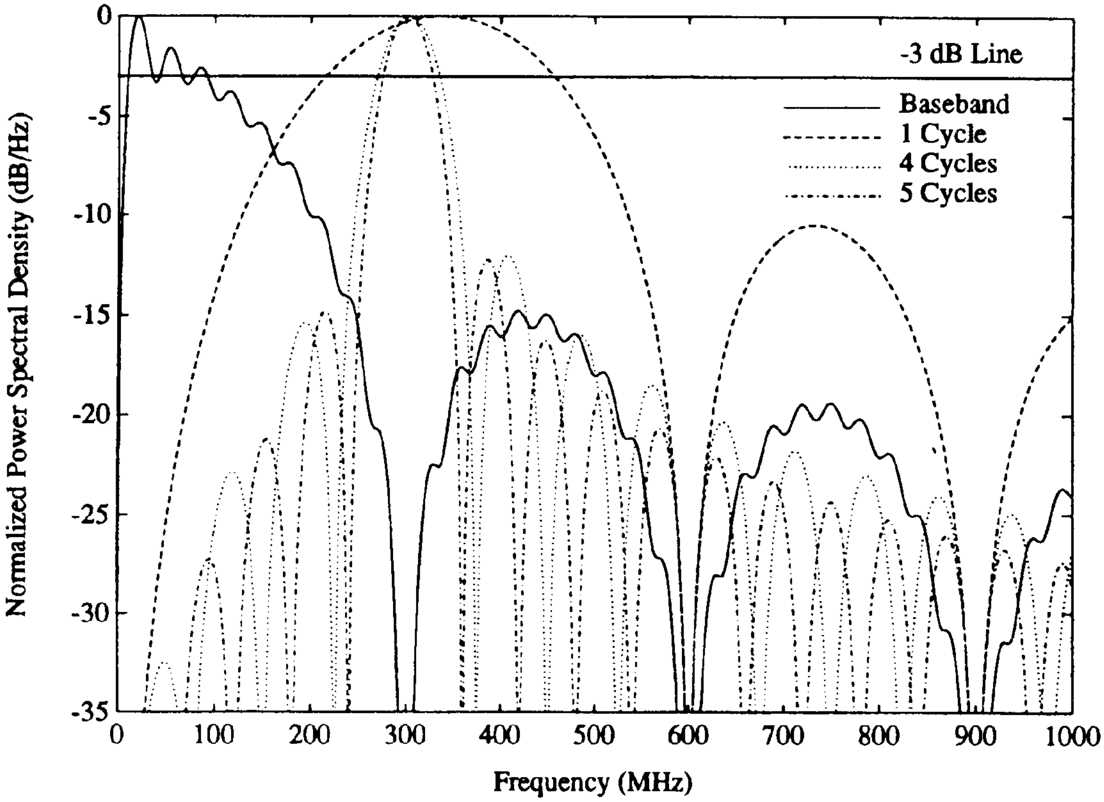


Figure 2.9 Energy spectra for signals shown in Figure 2.8. The baseband and monocycle signal have a 3.3-ns pulse width. The monocycle and polycycle signals have a 300-MHz center frequency. The polycycle pulses have 13.2 and 16.5 ns widths, corresponding to a 25 and 20% bandwidth, respectively.

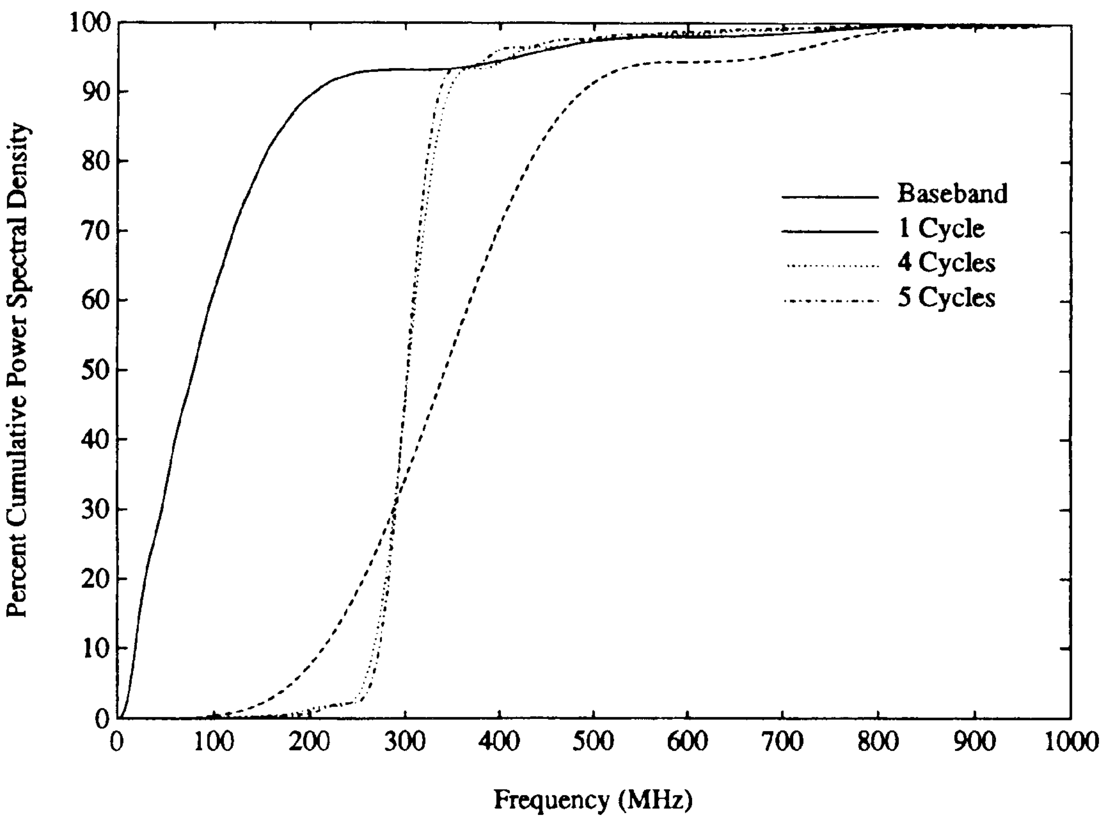


Figure 2.10 Cumulative power spectra for the signals shown in Figure 2.9.

MONOCYCLE WAVEFORM

The next class of waveform shown in Figure 2.8 is a monocycle waveform. This waveform can be considered carrier-based with a center frequency equal to the frequency of the single sinusoidal cycle and a pulse width equal to the period of the sinusoid. A potential advantage of this waveform over the baseband signal is that the peak of the energy spectrum of a single pulse is not at zero frequency. Figure 2.9 shows that the spectral peak of this waveform is near the frequency of the sinusoid, and Figure 2.10 shows that 90% of the energy is contained between 200 and 450 MHz.

As in the case of the baseband signal, the monocycle waveform can also be coded to increase the duty cycle, but the single cycle of the sinusoid has an inherent average value of zero, thus coding is not necessary to avoid the need for a DC offset between pulses. The disadvantage of the monocycle waveform is that the polarity of the carrier is ambiguous with the time delay, which reduces its ability to suppress certain types of clutter. (See the discussion on Clutter Rejection in Section II.)

POLYCYCLE WAVEFORM

A polycycle waveform is the last class in the continuum. This waveform is definitely recognizable as carrier-based because there are multiple cycles of the carrier within the pulse envelope (thus the name polycycle). Suppose a narrowband signal has 1% bandwidth, which means that the bandwidth is no more than 1% of the center frequency. Since the width of a pulse is approximately the reciprocal of its bandwidth, a 1% bandwidth is equivalent to saying that the duration of the pulse is 100 times the period of the carrier, i.e., a 1% bandwidth waveform will have 100 cycles of the carrier within it. Using the 20% bandwidth definition for UWB, a UWB waveform would have, at most, five carrier cycles in one pulse. Hence, the class of polycycle waveforms includes some of the UWB waveforms. The spectra for the polycycle pulses shown in Figure 2.9 are for 13.2 and 16.5 ns pulses impressed upon a 300-MHz carrier frequency, giving them a 25 and 20% bandwidth, respectively. Note from Figure 2.10 that the energy of these pulses is confined to a bandwidth of approximately 60 MHz about a nominal center frequency of 300 MHz.

A taxonomy of the relationships of all of the various waveform types is shown in Figure 2.11. The columns in the figure are arranged in order of decreasing percent bandwidth from left to right. The columns have titles corresponding to the categories discussed above. Other waveform names that have been used to describe these types of signals are inserted beneath these general headings to indicate the category where they best fit. Note that the bracket at the bottom of the figure extends partially into the polycycle column. This is meant to indicate graphically that some of the waveforms in this category fit the criteria for UWB.

IV. TECHNICAL ISSUES IN UWB RADAR SYSTEM DESIGN

Once the requirement has been established that a UWB system is the best choice for a particular function, the system developer is then faced with the problem of selecting the most appropriate technology for system implementation. The topics in this section are meant to provide an overview of factors to consider in subsystem selection/design. At this time, UWB technology should be considered still to be in its infancy. Hence, a wide variety of suitable components is not generally available. For this reason, some of the discussion below addresses the suitability of conventional components and design techniques for application to UWB system design.

TRANSMITTERS AND SOURCES

Most of the current issues surrounding transmitters for UWB radars are related to the capabilities of, and the need for, high power switching devices. One of the expectations in UWB technology is that very fast, but inexpensive, switches which can hold back very high voltages will bring forth an inexpensive high power transmitter. Some of the switch technologies have continuously controllable resistance to provide some control over pulse shape, but many do not. Other switch modules use a cascade of switches along a transmission line for pulse shaping, but the pulse shape is not easily altered after the switch is manufactured.

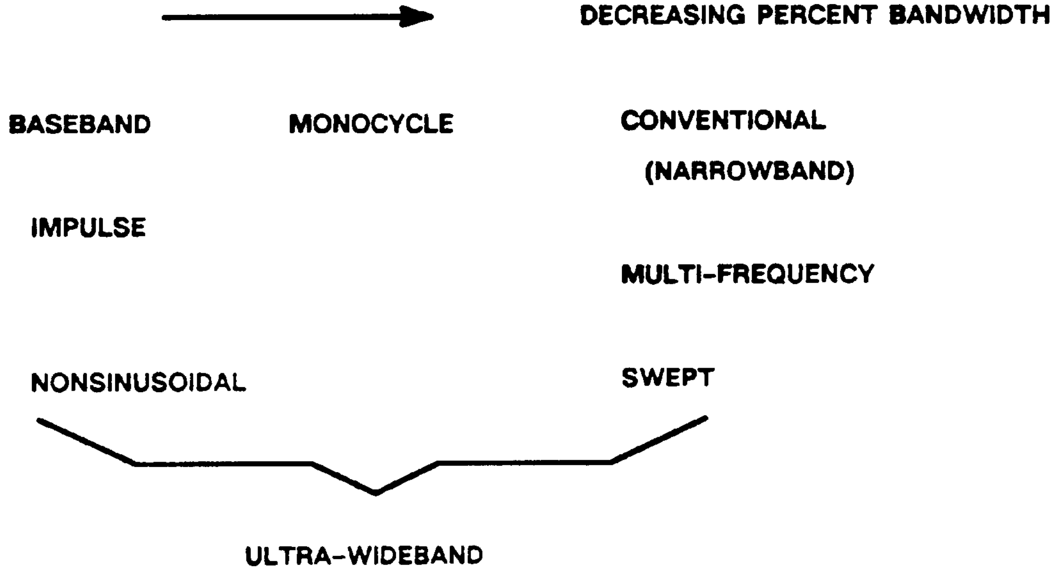


Figure 2.11 Classification of radar waveforms.

There are some applications where the character of the waveform which is input to the antenna is not extremely critical. For example, for buried object detection where radiated beam control is not critical, it is practical to use the signal directly from a high power switch. In other cases, the waveform design may be crucial not only to the range and velocity measurement, but also to the angle measurement. Then, too, even for the ground penetration application, clutter rejection capability would be improved if the radiated beam shape were better controlled. Would it be worth the added expense to improve the waveform generator and antenna to obtain this added capability?

Switching sources available when this book was published control extremely high powers (in the range of gigawatts). Also, the duration of the signal from such devices is extremely short (around one ns), but the recharge time is fairly long (several milliseconds). On the one hand, the available peak power is much higher than anything available from more conventional transmitter technology, but the *average* power is considerably lower. Since detection range is dependent on signal energy and recognizing that energy is equal to the time integral of average power, it is clear that currently available switching devices would be useful only for short range applications. (Chapters 4, Part 2, and 6, Part 1, address switching technology and UWB waveform generation.)

It has been concluded (Reference 9, p. IX-9) that it will not be possible to produce enough power to cause target materials to exhibit nonlinear properties (see Appendix 2C for a discussion on the definition of nonlinearity). Barring that requirement, the required power for UWB radar is the same as for conventional radar purposes. In conventional radar design, when enough average power cannot be obtained with the available transmitter, the waveform is redesigned to provide a higher duty cycle without allowing any reduction in range resolution or unambiguous range. This approach also could be taken for UWB systems. The question is: Will the added complexity of the waveform generator (for the transmitter) and the decoder (for the receiver) be more expensive than using a conventional approach?

Alternatively, the individual radiated pulses could remain simple and the pulse repetition frequency (PRF) could be increased to provide more average radiated power. For this approach to be effective, the PRF would need to be so high that the unambiguous measurement range of the radar would be considerably shorter than its maximum detection range. The radar receiver itself could remain simple, but a more complicated processor would be required to sort out the ambiguous range measurements. Also, increased clutter level due to ambiguous clutter returns may defeat much of the improvement resulting from the higher average power returned from the target.

Another issue concerning transmitter sources is related to the antenna issue. If an efficient and controllable transducer for radiating the signal from a switched power source is not available, what will we do with the switch. Keeping in mind that the transducer is going to perform at least one differentiation on the input signal (see the section Antennas and also Reference 24), it may be necessary to perform some signal shaping on the output of the switch before it is input to the antenna. This signal shaping circuitry will have to operate at high power levels from the switch. It remains to be seen whether this combination of components is less complex and expensive than one which uses a low power generator to produce the desired waveform shape, followed by a high power amplifier to produce the required output power.

ANTENNAS

Underlying the discussion of antenna behavior in most classical treatments is the assumption that the signal being radiated is (or is approximately) sinusoidal. Hence, theoretical results predicting antenna performance are very accurate when the percent bandwidth is small. Predicting (and controlling) the radiation pattern for larger percent bandwidth signals such as UWB signals is more problematic. The reason for this problem is that no antenna transmits a signal with the same form as its input. All antennas transmit a signal whose form is of the derivative of the input. The number of differentiations is dependent upon the type of antenna.[24]

A simple way to see that this is true is to recognize that the radiated field strength is always proportional to the acceleration of charge in the antenna. To see this relationship, we begin with the source-free form of one of Maxwell's equations. This equation makes no assumption about the form of the solution and is, therefore, equally valid for any type of waveform.

$$\nabla \times E = -\mu \dot{H} \tag{2.4}$$

where:

E = The electric field intensity in volts/meter
$\dot{H}$ = the time derivative of the magnetic field intensity in amperes/meter-second
μ = the magnetic permeability of the medium (in this case free space) in Henrys/meter

The relationship between the magnetic field and the vector potential is:

$$\mu H = \nabla \times A \tag{2.5}$$

where:

A = the vector magnetic potential in Webers/meter

The first time derivative of Equation 2.5 is

$$\mu \dot{H} = \frac{\partial}{\partial t}(\nabla \times A) \tag{2.6}$$

Substituting Equation 2.6 into Equation 2.4 we have

$$\nabla \times E = -\frac{\partial}{\partial t}(\nabla \times A) \tag{2.7}$$

which means that

$$\nabla \times E = -\nabla \times \left(\frac{\partial A}{\partial t} \right)$$

$$E = \frac{\partial A}{\partial t} \tag{2.8}$$

Now, since

$$A \propto J \propto i$$

where:

J = current density in amperes/meter^2
i = transmitter drive current in amperes

we can also write

$$E \propto \frac{\partial i}{\partial t} \tag{2.9}$$

Hence, the E-field strength has a temporal variation which is proportional to the time derivative of the antenna forcing function.

Looking at a simple example of the implications of this condition, suppose an envelope signal, $a(t)$ is impressed upon a carrier signal $\cos\omega_c t$:

$$s(t) = a(t) \cos \omega_c t \tag{2.10}$$

The first derivative of this function has two terms

$$\dot{s}(t) = -\omega_c a(t) \sin \omega_c t + \dot{a}(t) \cos \omega_c t \tag{2.11}$$

The first term in Equation 2.11 contains the original form of the information-bearing signal while the second term contains its derivative. The carrier in the first term has experienced a 90° phase shift and has also been multiplied by a "scale factor" ω_c, but these modifications are of no consequence to the information transmission.

To simplify the process of extracting the information from this signal at the receiver, it would be desirable to make the second term as small as possible relative to the first. This means we should strive for

$$\omega_c a(t) \gg \dot{a}(t) \tag{2.12}$$

The magnitude of the frequency will be much larger than the magnitude of the envelope; hence, the frequency is the dominant term on the left side of the inequality in Equation 2.12. Thus, an equivalent requirement would be for $\omega_c \gg \dot{a}(t)$. The bandwidth of the envelope is proportional to its rate of change (i.e., its time derivative). Hence, the requirement in Equation 2.12 is equivalent to requiring that the percentage bandwidth be small. Therefore, minimizing the ever present effect of differentiation caused by the antenna requires a small percent-bandwidth signal.

There is some work ongoing[26-27] that begins with the initial intent of transmitting a large percent bandwidth signal. These efforts are producing some interesting results that suggest there may be ways to take advantage of the coupling across measurement domains suggested in Reference 11 and produce small beams of large percent bandwidth signals without enormous physical apertures.

In the remainder of this section, we will look at the behavior of conventional antennas radiating UWB signals to expose some of the shortcomings of this approach. This section is divided according to major classes of UWB waveforms and the performance of each suitable type of antenna is discussed within the subsection.

Array Antennas

Baseband and Monocycle Waveforms

The formulation of a *beam* by an antenna occurs for a baseband waveform in the same way that it occurs for carrier-based waveforms, which is through the creation of a spatial region where constructive interference occurs among signals emitted from different regions on the antenna. Constructive interference is the result of the addition of multiple signals with like polarity. (Note that *in-phase* sinusoidal signals have like polarity.) The process of beam formation from array antennas will be examined with the aid of Figure 2.12, which shows a linear array with element spacing d and length L. Each element is radiating a stylized baseband pulse. The waveform actually radiated may be different from that shown in the figure, depending on the ability of the radiating element to match the transmit source and free space. The stylized pulses are used here to simplify the presentation of the beamforming concepts.

The ordinary way to evaluate beamforming capabilities is to allow all elemental radiation to occur omnidirectionally and to evaluate the composite summation for various angles away from the array normal. An equivalent method for a baseband waveform is shown in Figure 2.12. Here, a line representing the orientation for a wavefront is superimposed on the elemental radiation patterns. The orientation of this wavefront line is orthogonal to the propagation direction under consideration. When the wavefront is oriented at an angle greater than θ to the array, the radiation from all of the elements does not arrive simultaneously and hence θ can be defined as the edge of the antenna beam.

Similarly, Figure 2.13 shows the situation for grating lobes. When the angle relative to the array is ϕ_0, the pulse from one pulse repetition interval (PRI) aligns with pulses from other PRIs. This alignment produces a constructive interference level which is as strong as for the direction of propagation at zero degrees relative to the array normal. Hence, the alignment ϕ_0 represents a grating lobe situation. Note from Figure 2.13 that this situation begins to develop for angle ϕ_1. At this angle, the radiation from two elements is aligned. As ϕ increases, the number of elements in alignment will increase until the angle ϕ_0 is reached.

From Figures 2.12 and 2.13, it is clear that the angular behavior of the radiation pattern is dependent on the time-domain properties of the radiated waveform. This is an example of the coupling which occurs across these measurement dimensions for UWB waveforms. (Note that in narrowband systems the angular properties of the radiation pattern are a function of the duration of the carrier waveform, rather than duration of the pulse.)

Some sample calculations using the formula in Figure 2.12 provide other insights concerning the requirements for producing a desirable beam shape. The formula was used to produce the graphs shown in Figure 2.14. From the figure we see that with a 100-MHz bandwidth, a 10-m aperture would produce a 35° beamwidth. To reduce the beamwidth to a more acceptable size, for example, 3.5°, would require a bandwidth ten times larger (or equivalently, an aperture ten times larger).

A sample calculation for grating lobes using the relationship in Figure 2.13 shows that the problem there is not so difficult. Grating lobes do not exist when the argument of the arcsin function is greater than one. Even using a PRF as high as 100 MHz, the element spacing would need to be only less than 3 m to prevent grating lobes. For the minimal grating lobe condition indicated by ϕ_1 in Figure 2.13, a PRF up to 30 MHz could be used with apertures up to 10 m.

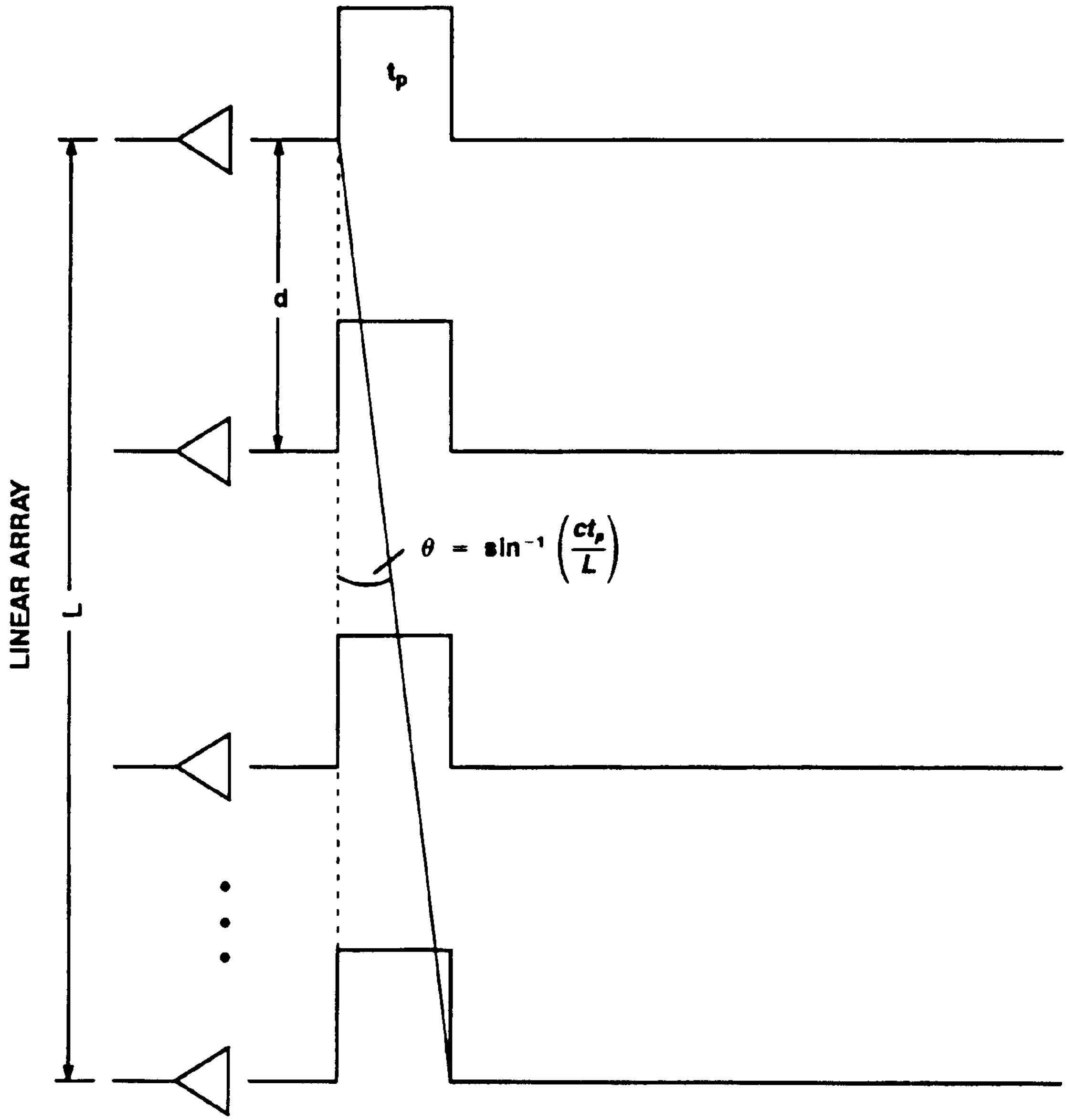


Figure 2.12 Stylized example of a baseband waveform emitted from a linear array showing relationships for beamwidth. The array extent is L. The element spacing is d. The pulse duration is t_p.

The results above show that the beamwidth and grating lobes are dependent on the duration of the pulse. This is in contrast to the narrowband situation where the grating lobes are related to the duration of a period of the carrier (i.e., the wavelength). This relationship is important to recognize here for two reasons. First, even if the UWB radar does not experience any grating lobes from its own signal, if there are any narrowband signals impinging on the antenna they can excite grating lobes in the usual way. Notice though that if the UWB pulses were made short enough to contain only one cycle of the carrier, the grating lobe behavior would be the same as for the baseband pulse. It is the multiple carrier cycles of the polycycle pulse which give rise to the normal grating lobe structure.

Several summary observations can be made from the discussion above. First, the time domain and spatial domain properties of a UWB radar interact; therefore, the antenna and the waveform must be designed together in order to obtain desirable performance in both domains. Second, in addition to the size and element spacing requirements, there are also the requirements to control the relative timing of the signals at the array elements using time delay control rather than phase shifters. Third, to obtain the radiation properties suggested in References 26 and 27 it may also be necessary to control the shape as well as the delay of the signal to each element in order to control beamsteering.

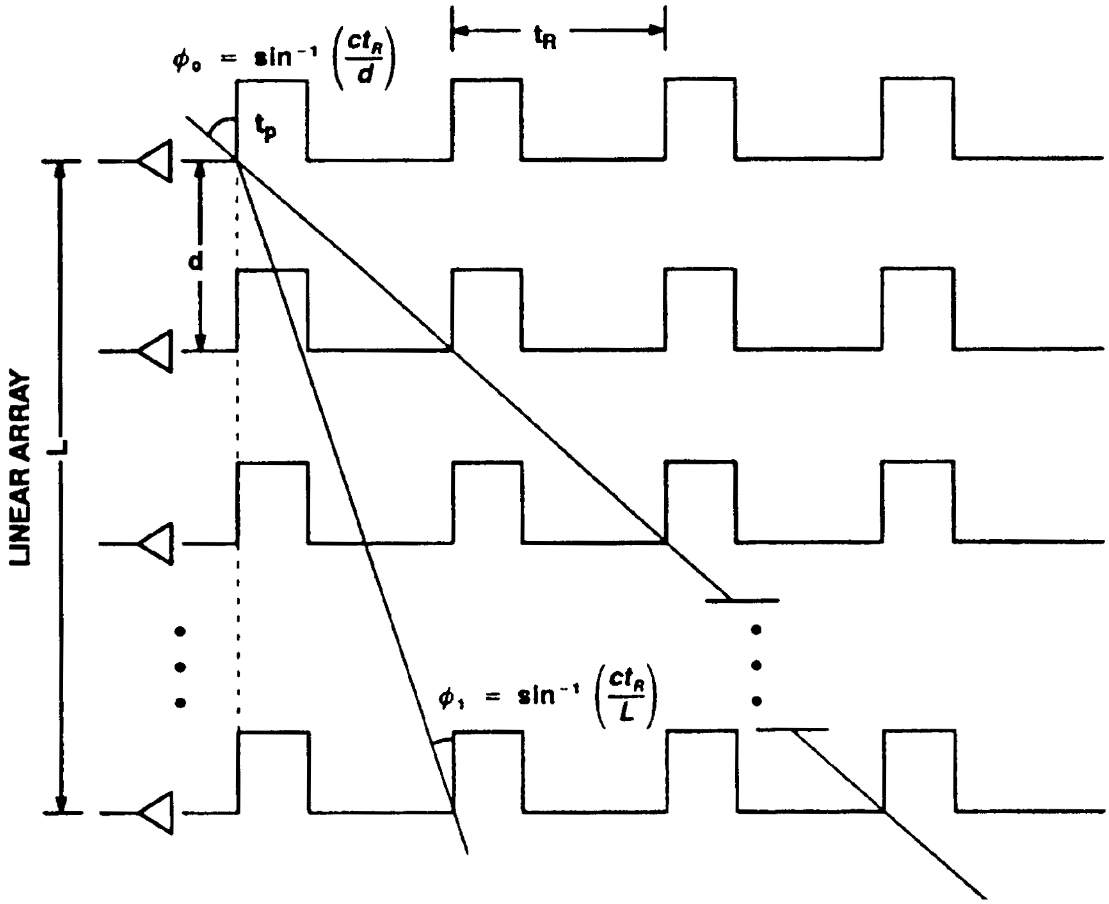


Figure 2.13 Stylized example of a baseband waveform emitted from a linear array showing relationships for grating lobes. The interpulse period is T_R.

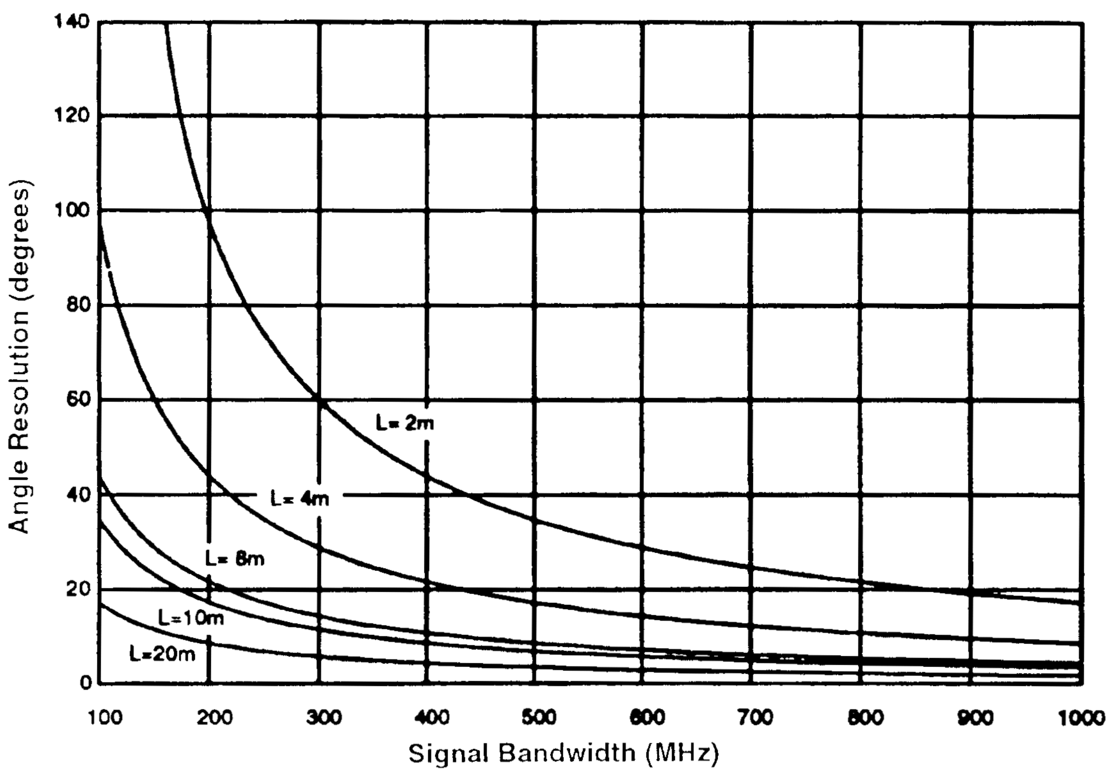


Figure 2.14 Antenna beamwidth resulting from various signal bandwidths and antenna aperture sizes. The bandwidth is computed as $1/t_p$ from the results shown in Figure 2.12.

Control of interelement coupling may also present unique problems. Because in narrowband design, the pulse is longer than the array, the entire aperture is illuminated simultaneously and the contribution of all elements can be brought to bear in controlling coupling. Ultra-wideband pulses can be much shorter than the aperture, so these coupling control techniques may not work.

Polycycle Waveforms

For polycycle sinusoidal illumination, an array with regularly spaced elements has a convenient closed-form representation for the directivity (see Reference 12, pp. 11-19):

$$D(\theta) = \frac{\sin\left(\frac{N\pi d}{\lambda}\sin(\theta)\right)}{\sin\left(\frac{\pi d}{\lambda}\sin(\theta)\right)}$$

where:

N is the number of elements
d is the spacing between the elements
λ is the wavelength of the input signal.

The array is designed for a particular wavelength, λ_c; $d = \lambda_c/2$.

As long as the radiation resistance of the individual elements does not vary significantly when the signal frequency is moved away from the design frequency, the gain at the exact center of the beam will not change. (Constant gain should persist until the wavelength gets so large that coupling between elements becomes significant.) However, because the gain at locations at any separation from the exact beam center will change, frequency offset has more of an effect on beamwidth than gain. Frequencies below the design frequency will produce a broader beam; frequencies above the design frequency will produce a narrower beam. This behavior is analogous to that of the baseband waveform described above where the beamwidth was inversely proportional to pulse width.

For frequencies above the design frequency the radiation pattern will exhibit grating lobes. For polycycle waveforms the existence and location of grating lobes is dependent on the period of the carrier (as is the case for narrowband systems), rather than the duration of the pulse, thus it would appear that in any application of an array antenna for a polycycle UWB signal, one would want to put the upper edge of the signal spectrum at the design frequency of the antenna. The signal frequencies below the design frequency will have an effect on the sidelobe structure, but they will not produce grating lobes.

For a uniform array antenna the beamwidth is inversely related to the frequency offset from the design frequency. If the lower band limit of the signal was 25% below the upper band limit, the antenna beamwidth for the frequency components near the lower band limit would be approximately 25% wider than that for the frequency components at the upper band edge.

Reflector Antennas

A reflector antenna is a continuous aperture and does not exhibit the grating lobe phenomenon. (Some reflectors are made of wire mesh to reduce weight and wind resistance. These are effectively discrete apertures that can exhibit grating lobes for frequencies above the design frequency.) As with arrays, there is also the possibility of a variation in gain as a function of frequency. There are two parts of the antenna that can influence the gain: the reflector aperture itself and the feed. The gain of both of these components determines the overall gain of the antenna. The gain of the reflector should remain fairly constant until the beamwidth of the feed becomes so large that the spillover loss becomes significant. The gain of the feed will, in general, vary directly with frequency while the beamwidth of the feed will vary inversely with frequency, thereby causing the lower frequencies to illuminate more of the aperture.

These two variations in the feed behavior tend to keep the gain of the overall antenna fairly constant until the spillover losses become significant.

To minimize the effect of spillover loss, it would be best to put the *lower* band limit of the signal at the design frequency of the antenna. The higher frequencies would then cause the beamwidth of the feed to decrease (which will cause less of the aperture to be illuminated), but the gain of the feed should increase. One drawback to this approach arises when the beamwidth of the feed becomes small enough that its sidelobes also illuminate the reflector. The sidelobe radiation is 180° out of phase with that from its mainlobe and will cause the reflector to exhibit reduced mainlobe gain and higher sidelobe levels when the frequency is high enough to cause sidelobe radiation from the feed to illuminate the reflector.

Summary on Antennas

In general, existing antennas should be capable of radiating mono- and polycycle waveforms. The only real difference in performance of these two antenna classes is the possibility of generating grating lobes; the array antenna may generate grating lobes, but the continuous reflector antenna will not. Both types exhibit the same coupling between frequency and beamwidth; obtaining a usable beamwidth and managing the effects of a frequency-dependent beamwidth will present a problem for either type of antenna.

A baseband waveform may contain extremely low frequency components (see Figures 2.9 and 2.10). Conventional antenna designs will have particular problems maintaining beamwidth control for this type of waveform; preserving gain will become a problem as well. The gain of a reflector will begin to degrade due to spillover losses, while the array gain will degrade due to coupling between elements at the lower frequencies. For these reasons, and due to the interaction between the time domain measurements and the spatial domain measurements in an UWB system, it appears that an antenna specifically designed for baseband use will be required to radiate this type of waveform with a desirable beam shape.

Ultimately, the radar application of UWB signals is dependent upon the availability of a reasonable size antenna which is capable of directing a useful beam size in a desired direction. At present this is a critical requirement. New approaches to design are needed such as those described in References 26 and 27, which begin the antenna design with the assumption of a UWB waveform rather than beginning with a narrowband antenna and attempting to make it radiate the UWB signal. Chapter 5 discusses UWB antennas.

RECEIVERS

Introduction

Whether the radar system is UWB or not, there are several functions which must be performed by the receiver. These functions include: (1) extraction of the desired signal from the background interference, (2) integration of the desired signal to increase the SNR, (3) the initial detection function, including any logic needed to keep the false alarm rate at a specified level, and (4) extraction of certain parameters from the received signal which are necessary for target reporting and/or tracking. Because of the coupling of the four measurement domains (range, Doppler, azimuth, and elevation) which occurs for UWB signals, the receiver will also play a role in the angle measurement process. This section addresses each of these requirements and focuses on the unique differences which arise due to UWB waveforms.

Coherent Signal Processing

We want to be able to optimize the process of extraction of the signal from the background interference. It has been shown theoretically that the greatest improvement in SNR is achieved when coherent signal processing is performed on the input signal. Linear coherent signal processing can be performed in two complementary ways: correlation and matched filtering (Reference 28, pp. 53–54). Correlation produces an SNR improvement by bandwidth compression; matched filtering, by pulse compression.

A Doppler processor is an example of correlator processing. Note that because of the bandwidth compression, the target range is no longer available at the output of the Doppler processor. To overcome this limitation, Doppler processing is applied to several unique range bin samples in parallel. The target range is inferred from the Doppler processor with the largest output.

In contrast to Doppler processing, matched filtering is usually applied to one pulse, rather than a train of pulses. It is conceivable to have a tapped-delay line type of filter structure which would coherently sum a finite train of pulses. In a very limited sense, this is the processing method used in moving target indicator (MTI) processing. This type of filter, however, is sensitive to any Doppler effect on the length of the pulse train, and in practice there would need to be a different pulse train matched filter for all possible target velocities. Having such a bank of filters would provide the capability of estimating the target range and velocity, just as in the more conventional pulse-Doppler processor.

Note that in each of the coherent signal processing examples above, the processes of signal extraction, interference rejection, and parameter estimation become inseparable.

Coherent processing also can be used to estimate other combinations of parameters, for example, azimuth and center frequency. The same type of tradeoffs occur here as well, because high resolution estimates of the target parameters are not available in both of these dimensions simultaneously. One could obtain high resolution angle estimates while having to infer center frequency or vice versa. It is also possible to devise a processor that would provide estimates on all four target parameters at once. However, this is of no particular value in the narrowband case, because the range-Doppler pair of measurements is independent from the azimuth-elevation pair of measurements.

UWB Receiver Processing

What would be the difference now if we remove the narrowband characteristic from the signal definition? Three fundamental issues arise. The first issue has to do with the fact that coherent processing has been proven to be the optimum *linear* method for SNR enhancement. Since we are no longer dealing with narrowband waveforms, is it still true that the best receiver is linear? In particular, for the baseband video signal structure which conveys no information by its amplitude, is it best to process this waveform with a linear receiver? The second issue which arises is the technology available to process the UWB signal. For signals with bandwidths less than 100 MHz, digital implementations for matched filters are possible. Beyond this, the limited number of bits available from A/D converters implies that only analog processing can currently provide the required bandwidth. If analog processing is used, the integration time of the matched filter is limited to the order of 10 μs; otherwise, insertion loss becomes excessive. For correlators, digital processing has the same bandwidth limit, but analog correlators do not perform well for very low duty cycle waveforms. The third issue concerns the parameter estimation requirements necessary to support the target angle estimate. Since the four measurement domains are now coupled, the transmitted waveform, the antennas, and the receiver must all be designed interactively.

The performance demands placed on the receiver must be balanced with the performance of the rest of the components in the overall system equation. How high will the transmit power be? Will there be several of these transmitters in an array, or just one feeding a reflector? How high can the transmit power go before arcing occurs in front of the aperture? How will the propagation medium affect the transmitted waveform? What will be the width of the transmitted pulse? How will the differentiations performed by the transmit and receive antennas and the target itself affect the characteristics of the received pulse? Would certain waveforms that are less technologically demanding be good near-term choices? Depending on how these questions are answered for a particular system application, the receiver may be as simple as a comparator followed by a digital counter or something much more complex such as an array of large-word-size, high speed A/D converters followed by complex digital signal processors.

Perhaps the most intriguing issue of all concerns the overall coupling of the various measurement domains. For example, these couplings indicate that with sufficient range resolution a required angle resolution can be achieved with little dependence on the antenna aperture size. Another example, which is discussed in Appendix 2B, is that with sufficient range resolution, target velocity estimation is possible without measuring the Doppler effect on the carrier frequency. The techniques for joint design of waveforms, antennas, and receivers are open research areas — ones with exciting possibilities for future system designs.

V. SUMMARY

This chapter focused on fundamental principles to show the differences and similarities of conventional and UWB radar. The fundamental principles involved include the following:

- Measurement accuracy and resolution
- The relationship of signal bandwidth and duration to measurement performance, independent of signal center frequency
- Signal energy as a determinant of detection range
- Time delay as the fundamental quantity, in contrast to phase shift which is an equivalent method for *narrowband* signals
- Target reflectivity depends on (1) electrical size of scatterers, (2) relative position of scatterers on complex target, and (3) the incident and reflected angles

In this discussion, we saw that there is no inherent uniqueness in UWB signals in terms of measurement capability. The range resolution is dependent on the absolute bandwidth of the signal and the velocity (Doppler) resolution is dependent on the duration of the signal. A UWB signal may be required if fine range measurement resolution requirements are combined with a requirement for a low center frequency. Low center frequency is required for such applications as ground penetration (and also sea penetration) and may be helpful for penetration of foliage and for detection of reduced cross-section targets. As shown in Table 2.3, low center frequency is also needed when it is necessary to excite the resonant frequencies of a target to aid in identifying the target. Table 2.5 provides a summary of the applications where wide bandwidth and low center frequency are needed. Any application that requires a low center frequency and a wide bandwidth is a candidate for a UWB waveform.

Table 2.5 is not meant to be a collectively exhaustive set of conditions for UWB waveform usage, nor is it always the case that a combination that fits within the specifications of Table 2.5 will require a UWB waveform. The purpose of this table is to convey the general circumstances which may require that type of waveform.

Supposing now that a UWB waveform is required, what special considerations must be addressed to develop this type of system? From the discussion in this chapter we know that target reflectivity and antenna behavior will be different and that the time, frequency, and angle measurement resolutions will be coupled. Perhaps the most crucial concern that must be addressed in the design of the UWB system is the repercussions of explicitly violating the narrowband constraint. By doing so, we have (as shown in the Antennas section) increased the energy content of the derivative terms of the signal. Ordinarily, in narrowband systems, these terms are intentionally made small and then neglected. In the UWB system design, these terms will not be small and if neglected will cause undesirable responses from the system. Accounting for all of the differentiations performed by the antennas and the target will probably require the inclusion of the compensating number of integrations in the transmitter or receiver, or both, so that the signal at the receiver output is of usable form.

Because of the coupling of time and angle resolutions brought about by the UWB waveform, it will be necessary to design the antenna (aperture illumination function) and the waveform simultaneously. Ambiguity functions are currently used in waveform design to aid in assessing the resolution and ambiguity performance offered by a candidate waveform (for the time and frequency dimensions). A similar process will now be required for the joint assessment of the angle and time resolution of the UWB system (see Chapter 5).

The physical antenna design itself will require some unique properties. Figure 2.14 shows that with a uniform array it is possible to get approximately 10° beamwidth from a reasonable bandwidth signal (<1 GHz). However, to get the beamwidth much smaller requires a large increase in aperture size or

Table 2.5 Requirement Combinations that May Require UWB Waveform

Low Center Frequency	Wide Bandwidth
• Foliage penetration • Ground penetration • Sea penetration • Target resonance excitation • Low cross-section targets	• Range profiling/target imaging • Clutter suppression

signal bandwidth or both. A beamwidth of about 10° seems to be the point of diminishing returns for radiation of UWB signals from conventional arrays. Also, there is the implicit assumption in the results of Figure 2.14 that all of the beam formation is performed using time delays in the individual antenna elements, not phase shifts.

From the last two paragraphs there seems to be a need for, and an opportunity to, employ a new approach to antenna design for UWB signals. Some research which seems aimed at this need is currently ongoing. The results being reported by Ziolkowski[26,27] are very interesting because his technique begins with the assumption of a UWB signal and proceeds from there to produce an antenna design rather than attempting to stretch an existing narrowband antenna design for a UWB applications. His results demonstrate that it is possible to design a baseband signal (which takes into account the differentiations of the antennas and the target) and radiate it in such a way that it stays focused for unusually long distances. The effect is equivalent to extending the nearfield region by factors upwards of six to ten.

The last issue that is unique to UWB systems is the dispersive nature of the target and the propagation medium. Dispersion means that the time delay imposed on a signal is frequency dependent. Dispersion is not normally a problem in narrowband systems because (1) the dispersive nature of the propagation medium is too gradual to have an impact, (2) the range resolution of the radar is too coarse to detect the relative time shifts caused by the target (the only indicator of target dispersion available to a conventional radar is the phase fluctuation known as glint), and (3) the pulse is usually much larger than the antenna, thus any antenna dispersion is wiped out. A UWB system will have to compensate for this signal distortion in some manner. It should be possible to calibrate the radar for antenna dispersion, and the propagation medium dispersion may still be negligible. The dispersion of the target, however, will be unpredictable. From a detection point of view, the dispersion will hurt sensitivity. However, there may be something target-unique about the dispersion which could be used as a target identification cue.

Although not a unique problem to UWB radar, producing the signal energy needed for reliable detection may be an important issue in some designs. For example, there currently is considerable interest in designs that use a high power switch for the transmitter. The expectation is that such a transmitter would be beneficial from a cost standpoint. However, while these devices have tremendous peak power capacity, they currently have lower average power than conventional devices. There will need to be further advances in this technology to support long-range applications (greater than a few tens of kilometers). In contrast, other devices which may be used for transmit waveform generators have essentially continuous duty cycle, but do not have near the power output of the switch. While these devices offer highly flexible waveform control, they would need to be accompanied by a power amplifier suitable for the application and, again, energy will be limited.

In summary, there appear to be new applications where UWB systems are best suited. Practical working implementations will require both new design methods (for waveforms and antennas) and new technology to produce the high power/high duty cycle/wide bandwidth waveforms needed by these designs.

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APPENDIX 2A: SIGNAL CHARACTERISTICS GOVERNING RANGE AND VELOCITY MEASUREMENT RESOLUTION RADAR SIGNAL FUNDAMENTALS

It will be shown in this appendix that the velocity and range resolution properties of a radar waveform are dependent only on its duration and bandwidth, respectively. The measurement of velocity is discussed first because the relationships used in that discussion are more intuitive than those for range. The relationships necessary to the discussion of range measurement are obtained from those for velocity by recognizing the appropriate dualities between time and frequency. These are not the usual relationships encountered in the literature;[28,29] they are much simpler, and the results obtained are in full agreement with those obtained by other means.

Velocity Measurement

Motion detection and measurement are accomplished by observing changes in phase of the received signal relative to that of the transmitted signal. A moving target causes the wavelength(s) of the illuminating signal to be compressed or expanded due to decreasing or increasing target range. For most cases, this rescaling of wavelengths can be accurately characterized as a frequency shift. The frequency shift is directly proportional to the product of target velocity and transmit frequency ($f_d = 2\nu f_T/c = 2\nu/\lambda$). Hence, velocity *resolution* improves with increasing frequency.

Under the condition of a fixed frequency shift (constant target velocity) the velocity *accuracy* improves with increasing observation time, as can be seen from the following argument. An estimate of the target velocity can be obtained from the equation

$$\hat{v} = \frac{\lambda}{2}\left[\frac{\hat{\phi}(t) - \hat{\phi}(t - \hat{T})}{2\pi \hat{T}}\right] \tag{2A.1}$$

where

$\hat{\nu}$	=	estimate of target velocity,
	=	estimate of phase at time t
	=	estimate of T, the time interval between phase samples, and
λ	=	free-space wavelength

The error in estimating ϕ is a function of the signal energy-to-noise power density ratio. The error in estimating T is defined by the accuracy of the timing function. Alternately, the measurement of T can be considered perfect by redefining the error process associated with the measurement of ϕ. If this is done, Equation 2A.1 may be rewritten as

$$\hat{v} = \frac{\lambda}{4\pi T}[\phi(t) + \varepsilon(t) - \phi(t-T) - \varepsilon(t-T)]$$

$$\hat{v} = \left\{\frac{\lambda}{4\pi T}[\phi(t) - \phi(t-T)]\right\} + \left\{\frac{\lambda}{4\pi T}[\varepsilon(t) - \varepsilon(t-T)]\right\} \tag{2A.2}$$

where $\varepsilon(t)$ is a composite measurement error process due to errors in measurement of the signal phase and errors in measurement of the time at which the phase was measured. The first term on the right side of Equation 2A.2 is the actual velocity ν. The second is the total measurement error which can be made arbitrarily small choosing a sufficiently large value for T.

The observation interval is limited in practical monostatic systems (systems that employ a common transmit and receive antenna) in order to operate the transmitter and receiver on a time-shared basis. The isolation achievable between the transmitter and receiver is such that, if the receiver had to operate at the same time as the transmitter, a significant reduction of receiver sensitivity would result. Hence, in a time-shared operating mode a contiguous observation interval would be limited to the round-trip propagation time to the closest target of interest.

Typical minimum times are a few tens of microseconds. Thus, a velocity estimate (based on measurement of frequency shift) cannot be made in a single observation with acceptable accuracy, but, continuous observation of the changing phase is unnecessary since the expression for the velocity estimate based on a sequence of N discrete pulses can be written as

$$\hat{v} = \frac{\lambda}{4\pi T}\sum_{n=1}^{N}[\phi(t_n) - \phi(t_n - T_s) + \varepsilon(t_n) - \varepsilon(t_n - T_s)] \tag{2A.3}$$

where T_s is the interval between successive phase measurements $= T/N$. Since $\phi(t_n - T_s) = \phi(t_{n-1})$ and $\varepsilon(t_n - T_s) = \varepsilon(t_{n-1})$, the estimate collapses to

$$\hat{v} = \frac{\lambda}{4\pi T}[\phi(t_N) - \phi(t_1) + \varepsilon(t_N) - \varepsilon(t_1)] \tag{2A.4}$$

which indicates that only the samples at each end of the interval NT_s are required.

However, since the phase is no longer being observed continuously, it is necessary to make T_s small enough to detect every rotation of ϕ through 2π so that the difference $\phi(t_N) - \phi(t_1)$ may be determined *unambiguously*. To do so requires that $\phi(t)$ be sampled at least twice for each rotation through 2π radians. For example, a Doppler frequency shift of 50 kHz would rotate through 2π in 20 μs. Therefore, the maximum value for T_s would be 10 μs, or the sampling rate (which is usually referred to as the PRF) would be 100 kHz.

Range Measurement

A relationship analogous to Equation 2A.2 can be written by making the following substitutions

measurement time t	$\leftarrow$	measurement frequency f
observation time span T	$\leftarrow$	observation frequency span B
free space wavelength λ	$\leftarrow$	free space velocity c

$$\hat{R} = \frac{c}{4\pi B}[\phi(f) - \phi(f-B) + \varepsilon(f) - \varepsilon(f-B)] \tag{2A.5}$$

The meaning of this relationship is, perhaps, not intuitive. Ignoring the error term for the moment (which can be made arbitrarily small by appropriate choice of B), Equation 2A.5 is an approximation to the derivative of the phase with respect to frequency. This derivative is commonly referred to as the envelope delay or group delay (Reference 30, p. 18). It is used to measure the propagation delay characteristics (as a function of frequency) by measuring the phase shift imposed on each frequency. In an ideal channel the phase shift vs. frequency is a linear function and the derivative is a constant across the entire signal spectrum.

The measurement of phase *shift* to measure time delay through the channel implies that the phase at the input to the channel is known and is subtracted from the signal upon reception in order to determine the delay. In its present form, there is an implicit assumption in Equation 2A.5 that the phase of each frequency component is zero at transmission time. To make Equation 2A.5 usable with signals of arbitrary initial phase, it is modified as follows

$$\hat{R} = \frac{c}{4\pi}\left[\frac{\phi(f) - \phi_0(f) - \phi(f-B) + \phi_0(f-B)}{B} + \frac{\varepsilon(f) - \varepsilon(f-B)}{B}\right] \tag{2A.6}$$

As an example, let the signal be composed of two sinusoids which at transmission time, t_0, are

$$s_1(t_0) = A\sin\omega_1 t_0 + \phi_1$$

$$s_2(t_0) = A\sin\omega_2 t_0 + \phi_2$$

At the time of reception, they are

$$s_1(t_r) = A\sin\omega_1 t_r + \phi_1$$

$$s_2(t_r) = A\sin\omega_2 t_r + \phi_2$$

Using Equation 2A.6, the range to the target which caused this reflected signal (ignoring the error term) is

$$\hat{R} = \frac{c}{2}\left[\frac{\omega_2 t_r + \phi_2 - (\omega_2 t_0 + \phi_2) - (\omega_1 t_r + \phi_1) + \omega_1 t_0 + \phi_1}{\omega_2 - \omega_1}\right] \tag{2A.7}$$

$$\hat{R} = \frac{c}{2}(t_r - t_0)$$

Note that in this expression the difference between ω_1 and ω_2 has replaced the bandwidth term, B; it is this frequency difference which controls the accuracy of the range measurement. However, analogous to the velocity measurement, the phase of these two signals must be sampled often enough in frequency so that every rotation of ϕ through 2π will be detected. Minimizing the gaps in the spectrum will minimize the range ambiguities.

SIGNAL REQUIREMENTS FOR DESIRED RANGE AND VELOCITY MEASUREMENT PERFORMANCE

Velocity Measurements

The velocity error term from Equation 2A.4 is

$$e_v = \frac{\lambda}{4\pi T}[\varepsilon(t) - \varepsilon(t-T)]$$

The process $\varepsilon(t)$ is the combined error due to error in measurement of signal phase and error in measurement of the time the sample is taken. Because the error in measurement of phase varies inversely with the signal energy-to-noise power density ratio, this composite error varies in similar fashion. Defining the standard deviation of $\varepsilon(t)$ as $\sigma_\phi(r)$, the standard deviation of the velocity estimate is determined from

$$\sigma_{\hat{v}} = \frac{\lambda}{4\pi T}\sqrt{2}\,\sigma_\phi(r) \quad \text{m/s} \tag{2A.8}$$

To prevent ambiguities, the sample interval T_s must be such that the phase shift between samples is less than π radians:

$$|\phi(t) - \phi(t-T_s)| < \pi \tag{2A.9}$$

Dividing both sides of Equation 2A.9 by $2\pi T_s$

$$\frac{1}{2\pi T_s}|\phi(t) - \phi(t-T_s)| < \frac{1}{2T_s} \tag{2A.10}$$

the left side becomes the estimate of the Doppler frequency. Hence, multiplying both sides by $\lambda/2$ yields

$$|v_{mu}| = \frac{\lambda}{4\pi T_s}|\phi(t) - \phi(t-T_s)| < \frac{\lambda}{4T_s} \tag{2A.11}$$

from which the maximum value of T_s is determined as

$$T_s < \frac{\lambda}{4|v_{mu}|} \quad \text{seconds} \tag{2A.12}$$

where $|v_\mu|$ is the minimum velocity for which ambiguous reports can be tolerated. Any target which is moving at a velocity less than $|v_\mu|$ will be reported unambiguously. Any target moving faster than $|v_\mu|$ will have its velocity reported as being somewhere in the range $0 < v < v_\mu$.

Range Measurements

A similar development for range accuracy yields

$$\sigma_{\dot{R}} = \frac{c}{4\pi B}\sqrt{2}\,\sigma_{\phi}(r) \tag{2A.13}$$

for the standard deviation of the range measurement in terms of the frequency span B. Likewise, the maximum spacing B_s of frequency samples to prevent range ambiguities is

$$B_s < \frac{c}{2R_{\text{mu}}} \quad \text{Hertz} \tag{2A.14}$$

where R_μ is the minimum range for which ambiguous reports can be tolerated. Any target at a range less than R_μ will be reported unambiguously. Any target which is at a range greater than R_μ will be reported in the range $0 < R < R_\mu$.

Signal Parameter Conflicts

In a periodic pulse train, the spacing in the time domain is the reciprocal of the spacing in the frequency domain. Hence, Equation 2A.14 can also be written as

$$T_s > 2\frac{R_{\text{mu}}}{c} \quad \text{seconds} \tag{2A.15}$$

Comparison of Equations 2A.14 and 2A.15 shows there is a potential conflict between the requirement for tolerable velocity ambiguity and tolerable range ambiguity.

For a single pulse, the time bandwidth product is approximately equal to one: $T = 1/B$. Comparison of Equations 2A.8 and 2A.13 shows there is, therefore, a potential conflict between required levels of velocity accuracy and range accuracy. There are signal structures (such as spread-spectrum or pulse-compression signals) which decouple T and B. However, they do so by inserting gaps in one domain or the other. Thus, the result is a trade in accuracy for ambiguity (Reference 28, Chapters 4 to 7).

One final conflict exists within the velocity measurement itself. Equation 2A.8 indicates that a shorter transmitted wavelength leads to lower velocity inaccuracy, while Equation 2A.12 indicates a longer transmitted wavelength results in reduced velocity ambiguity.

The Doppler effect is used to measure velocity because the measurement of the Doppler shift has been less of a technological challenge than the measurement of range with sufficient accuracy to estimate range derivatives from range directly. However, if this could be done, all of the conflicts in simultaneous measurement of range and velocity would vanish. (It is shown in Appendix 2B that the frequency span needed to achieve the required range accuracy is of the order of 100 MHz.) Signal design would then be reduced to making the frequency spread sufficient for range resolution and the frequency gaps small enough for tolerable range ambiguity.

APPENDIX 2B: RANGE ACCURACY REQUIREMENTS FOR VELOCITY ESTIMATION FROM DIFFERENTIAL TIME DELAY RADIAL VELOCITY ESTIMATION

The radial velocity of a target can be estimated from its rate of change in range. This is normally not done because very precise range measurements are required to detect the change in range. Since we are now considering the use of very large bandwidths, it is of interest to determine if the velocity estimate can be made from time domain measurements. This is of particular interest, because if this type of velocity estimation is practical, it will eliminate the conflicting requirements for range and Doppler accuracy which in turn give rise to ambiguities in conventional waveform design.

The radial velocity is estimated using the following relationship

$$\hat{v} = \frac{r(t) - r(t+T)}{T} \tag{2B.1}$$

where $r(t)$, $r(t + T)$ are range measurements made at times separated by the time interval T. There are two sources of error is this expression. The first is caused by errors in the measurement of the range (actually the time of arrival) of the target. The second source results from errors in measuring the time interval T. This second source is due to timing jitter in the measurement system, and because the error does not normally increase when the value of T is increased, this error can be made arbitrarily small by making T sufficiently large. We will assume here that this condition occurs when the value of T is greater than a few PRIs, for example, 10 ms. This is comparable to the time required to make a velocity measurement using frequency measurements. With $T = 2$ ms, Equation 2B.1 shows that a range measurement accuracy of 1 ns is required to get a velocity accuracy of 20 m/s, or 2.5 ns for 50 m/s accuracy.

SNR REQUIREMENTS FOR TOA ESTIMATION

From Reference 31, p. 656, the Cramer-Rao bound for the standard deviation for the estimate of the time delay between two pulses is given by

$$\sigma_{\hat{\tau}} \geq \frac{1}{\sqrt{8\pi^2 d D \beta^2}} \tag{2B.2}$$

where

d = single-pulse predetection SNR
D = signal time-bandwidth product
β = receiver equivalent noise bandwidth

Now, this estimator gives the lower bound on the performance of all possible estimators and does not indicate that any realizable estimator will achieve this performance. Nonetheless, the result is useful in examining the possibility for measuring the velocity in this manner.

Evaluating Equation 2B.2 for a range accuracy requirement of 2.5 ns and $D = 1$ results in a bandwidth requirement of 14 MHz when the SNR is 10 dB. Even if a practical estimator required ten times more bandwidth to obtain the same measurement accuracy, the required bandwidth is still within the realm of UWB waveforms.

APPENDIX 2C: THE CONCEPT OF NONLINEARITY

The meaning of the term nonlinear is sometimes misunderstood. This appendix provides the precise meaning intended for Chapter 2.

To begin the definition of *nonlinear* we will first state what is meant by *linear*. Suppose the output from a system is $ax_1(t)$ when the input is $x_1(t)$ and its output is $bx_2(t)$ when the input is $x_2(t)$. The system is linear if, when the input is $x_1(t) + x_2(t)$, the output is $ax_1(t) + bx_2(t)$. This definition is expandable to the general case where the input may consist of the set of functions $x_i(t)$, i:$\{1, \infty\}$. Suppose now that the output from the system is $a_i\ x_i(t)$ when the input is $x_i(t)$ i:$\{1, \infty\}$ when the functions are input individually. If the output from the system is

$$\sum_{i=1}^{\infty} a_i x_i(t) \tag{2C.1}$$

then the system is linear.

In general, the input-output relationship of a system can be represented in polynomial form $y = \Sigma_{i=0}^{N}\ c_i x^i$. If the values of the coefficients for c_i for $i > 1$ are not equal to zero, then the system is nonlinear. Substituting the sum of inputs from the examples above into this polynomial function will show that the terms for $i \geq 2$ will cause cross products of the terms in the input sum of functions.

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