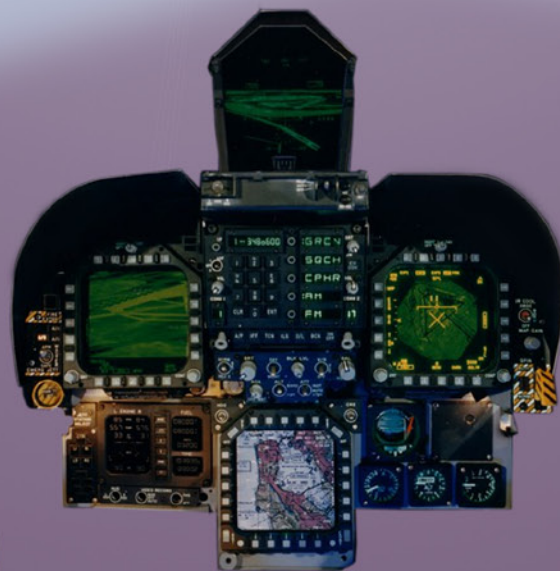




STIMSON'S Introduction to Airborne Radar

Third Edition



George W. Stimson
Hugh D. Griffiths
Chris J. Baker
Dave Adamy



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Contents

PART I Overview of Airborne Radar

Chapter 1 Basic Concepts	3
1.1 Echolocation	3
1.2 Radio Detection	4
1.3 Determining Target Position	6
1.4 The Doppler Effect	9
1.5 Imaging	10
1.6 Summary	12
Further Reading	13
Chapter 2 Approaches to Implementation	15
2.1 Noncoherent Pulsed Radar	15
2.2 Coherent Pulse-Doppler Radar	26
2.3 Exploiting Coherency	32
2.4 Summary	34
Chapter 3 Representative Applications	37
3.1 Weather Phenomena	37
3.2 Navigational Aids	39
3.3 Remote Sensing	42
3.4 Reconnaissance and Surveillance	44
3.5 Fighter/Interceptor Mission Support	46
3.6 Air-to-Ground Targeting	47
3.7 Proximity Fuses	49
3.8 Summary	49

PART II Essential Groundwork

Chapter 4 Radio Waves and Alternating Current Signals	53
4.1 Nature of Radio Waves	53
4.2 Characteristics of Radio Waves	56
4.3 Summary	61

Chapter 5 A Nonmathematical Approach to Radar 63

5.1 How a Phasor Represents a Signal	63
5.2 Combining Signals of Different Phase	65
5.3 Combining Signals of Different Frequency	66
5.4 Resolving Signals into In-Phase and Quadrature Components	70
5.5 Summary	74

Chapter 6 Preparatory Math for Radar 77

6.1 Signal Classification	77
6.2 Complex Numbers	78
6.3 Fourier Series	79
6.4 The Fourier Transform	82
6.5 Statistics and Probability	85
6.6 Convolution, Cross-Correlation, and Autocorrelation	90
6.7 Summary	93
Further Reading	94

PART III Fundamentals of Radar

Chapter 7 Choice of Radio Frequency 97

7.1 Frequencies Used for Radar	97
7.2 Frequency Bands	98
7.3 Influence of Frequency on Radar Performance	99
7.4 Selecting the Optimum Frequency	102
7.5 Summary	104
7.6 Further Reading	105

Chapter 8 Directivity and the Antenna Beam 107

8.1 Distribution of Radiated Energy in Angle	107
8.2 Characteristics of the Radiation Pattern	112

8.3	Electronic Beam Steering	116	12.5	Integration and Its Effect on Detection Range	173
8.4	Angular Resolution	117	12.6	Postdetection Integration	176
8.5	Angle Measurement	118	12.7	Summary	178
8.6	Antenna Beams for Ground Mapping	122	12.8	Some Relationships to Keep in Mind	179
8.7	Summary	122	Further Reading	179	
Further Reading	123				
Chapter 9	<i>Electronically Steered Array Antennas</i>	125	Chapter 13	<i>The Range Equation: What It Does and Doesn't Tell Us</i>	181
9.1	Basic Concepts	125	13.1	General Range Equation	181
9.2	Types of ESAs	126	13.2	Equation for Volume Search	185
9.3	Time Delay for Wideband Applications	127	13.3	Fluctuations in Radar Cross Section	187
9.4	Shared Advantages of Passive and Active ESAs	128	13.4	Detection Probability	188
9.5	Additional Advantages of the Active ESA	130	13.5	Cumulative Detection Probability	192
9.6	Key Limitations and Their Circumvention	131	13.6	Summary	193
9.7	Trend toward Digital Beamforming	132	Further Reading	196	
9.8	Summary	134	Chapter 14	<i>Radar Receivers and Digitization</i>	197
Further Reading	134		14.1	Basic Principles	198
Chapter 10	<i>Electronic Steered Array Design</i>	137	14.2	Low-Noise Amplification	198
10.1	Considerations Common to Passive and Active ESAs	137	14.3	Filtering	199
10.2	Design of Passive ESAs	141	14.4	Downconversion	200
10.3	Design of Active ESAs	143	14.5	Dynamic Range	203
10.4	Summary	148	14.6	Spurious Signals and Spectral Purity	204
Further Reading	149		14.7	Digitization	205
Chapter 11	<i>Pulsed Operation</i>	151	14.8	Radar Receiver Architectures	209
11.1	Advantages of Pulsed Transmission	151	14.9	Pulsed Noncoherent Receivers	209
11.2	Pulsed Waveform	152	14.10	Pulsed Coherent Receiver with Baseband Digitization	210
11.3	The Ambiguity Diagram	155	14.11	Pulsed Coherent Receiver with IF Digitization	211
11.4	Output Power and Transmitted Energy	155	14.12	Multichannel Receivers	212
11.5	Summary	158	14.13	Specialized Receivers	213
11.6	Some Relationships to Keep in Mind	159	14.14	Summary	214
Further Reading	159		Further Reading	215	
Chapter 12	<i>Detection Range</i>	161	Chapter 15	<i>Measuring Range and Resolving in Range</i>	217
12.1	What Determines Detection Range	161	15.1	Pulse-Delay Ranging	217
12.2	Electrical Background Noise	162	15.2	Range Ambiguities	219
12.3	Energy of the Target Echo	166	15.3	Eliminating Ambiguous Returns	221
12.4	Detection Process	171	15.4	Resolving Ambiguities	222
			15.5	How Many PRFs?	226

15.6	Enhanced Pulse Tagging (Range-Gated High PRF)	228	19.3	Line Width versus the Duration of the Pulse Train	282
15.7	Single-Target Tracking	228	19.4	Spectral Sidelobes	283
15.8	Electronically Scanned Radars	229	19.5	Summary	285
15.9	Summary	229	19.6	Some Relationships to Keep In Mind	285
	Further Reading	230		Further Reading	285
Chapter 16 Pulse Compression and High-Resolution Radar			Chapter 20 The Pulsed Spectrum Unveiled		
		233			287
16.1	Pulse Compression: A Beneficial Complication	233	20.1	Spectra	287
16.2	Linear Frequency Modulation (Chirp)	237	20.2	Spectrum Explained from a Filter's Point of View	294
16.3	Phase Modulation	242	20.3	Mathematical Explanation of the Pulsed Spectrum	298
16.4	Summary	248	20.4	Summary	304
	Further Reading	249		Further reading	305
Chapter 17 Frequency-Modulated Continuous Wave Ranging			Chapter 21 Doppler Sensing and Digital Filtering		
		251			307
17.1	Basic Principle	251	21.1	Doppler Filter Bank	307
17.2	Accounting for the Doppler Shift	252	21.2	Digital Filtering	310
17.3	Eliminating Ghosts	254	21.3	Inputs to the Filter	315
17.4	Performance	258	21.4	What the Digital Filter Does	317
17.5	Summary	259	21.5	Sidelobe Reduction	321
	Further Reading	260	21.6	Filtering Actual Signals	322
			21.7	Summary	325
				Further Reading	326
PART IV Pulse Doppler Radar			Chapter 22 Measuring Range-Rate		
					329
Chapter 18 The Doppler Effect			22.1 Range Differentiation		
		265			329
18.1	The Doppler Effect and Its Causes	265	22.2	Doppler Method	330
18.2	Where and How the Doppler Shift Takes Place	266	22.3	Potential Doppler Ambiguities	332
18.3	Magnitude of the Doppler Frequency	268	22.4	Resolving Doppler Ambiguities	334
18.4	Doppler Frequency of an Aircraft	270	22.5	Summary	336
18.5	Doppler Frequency of Ground Return	271		Further Reading	337
18.6	Doppler Frequency Seen by a Semiactive Missile	272	PART V Clutter		
18.7	Summary	273			
	Further Reading	274	Chapter 23 Sources and Spectra of Ground Return		
Chapter 19 Spectrum of Pulsed Signal					341
		277			
19.1	Bandwidth	277	23.1	The Amplitude of the Ground Return	342
19.2	Coherence	279	23.2	Doppler Spectra of Ground Clutter Returns	344

23.3	Relation of Clutter Spectrum to Target Frequencies	352
23.4	Return from Objects on the Terrain	353
23.5	Summary	355
	Further Reading	356

Chapter 24 *Effect of Range and Doppler Ambiguities on Ground Clutter* **359**

24.1	Dispersed Nature of the Clutter	360
24.2	Range Ambiguities	361
24.3	Doppler Profile	364
24.4	Summary	366
	Further Reading	366

Chapter 25 *Representing Clutter* **369**

25.1	Clutter as Noise	370
25.2	Limitations of the Noise Model for Ground Clutter	373
25.3	Improved Clutter Models	374
25.4	Other Characteristics of Ground Clutter	374
25.5	Discrete Scatterers	375
25.6	Predicting Detection Performance	376
25.7	Summary	380
	Further Reading	381

Chapter 26 *Separating Ground Moving Targets from Clutter* **383**

26.1	Introduction	383
26.2	Problem of Detecting "Slow" Moving Targets	384
26.3	Precise Angle Measurement	390
26.4	Summary	391
	Further Reading	392

PART VI Air-to-Air Operation

Chapter 27 *PRF and Ambiguities* **397**

27.1	Primary Consideration: Ambiguities	397
27.2	The Three Basic Categories of PRF	401
27.3	Summary	405
	Further Reading	406

Chapter 28 *Low PRF Operation* **407**

28.1	Differentiating between Targets and Clutter	407
28.2	Signal Processing	412
28.3	Advantages and Limitations of Low PRF Operation	417
28.4	Getting around the Limitations	417
28.5	Summary	422
	Further Reading	423

Chapter 29 *Medium PRF Operation* **425**

29.1	Differentiating between Targets and Clutter	425
29.2	Advantages and Limitations of Medium PRF Operation	428
29.3	Signal Processing	428
29.4	Rejecting Ground Moving Targets	430
29.5	Eliminating Blind Zones	431
29.6	Minimizing Sidelobe Clutter	433
29.7	Sidelobe Return from Targets of Large RCS	435
29.8	Summary	436
	Further Reading	437

Chapter 30 *High PRF Operation* **439**

30.1	High PRF Waveform	440
30.2	Advantages and Limitations of High PRF Operations	440
30.3	Isolating the Target Returns	440
30.4	Signal Processing	443
30.5	Ranging	445
30.6	Problem of Eclipsing	448
30.7	Improving Tail Aspect Performance	449
30.8	Summary	451
	Further Reading	452

Chapter 31 *Automatic Tracking* **453**

31.1	Single-Target Tracking	454
31.2	Track-While-Scan	458
31.3	Track Filtering	460
31.4	Summary	461
	Further Reading	462

PART VII Imaging Radar

Chapter 32	<i>Radar and Resolution</i>	465
32.1	How Resolution is Defined	466
32.2	Factors Influencing Choice of Resolution Cell Size	466
32.3	Achieving Fine Resolution	471
32.4	Summary	473
	Further Reading	473
Chapter 33	<i>Imaging Methods</i>	475
33.1	SAR	475
33.2	Spotlight SAR	480
33.3	Inverse SAR Imaging	481
33.4	Interferometric SAR	484
33.5	Polarimetric SAR	489
33.6	Tomographic SAR	490
33.7	Summary	491
	Further Reading	491
Chapter 34	<i>SAR Image Formation and Processing</i>	493
34.1	Unfocused SAR	493
34.2	Focused SAR	498
34.3	SAR Processing	501
34.4	Motion Compensation and Autofocus	508
34.5	SAR Image Interpretation	510
34.6	Summary	513
	Further Reading	513
Chapter 35	<i>Synthetic Aperture Radar System Design</i>	515
35.1	SAR Radar Range Equation	515
35.2	SAR Ambiguities	518
35.3	Bandwidth and Cross-Track Resolution	521
35.4	Beamwidth and Along-Track Resolution	522
35.5	Minimizing Sidelobes	522
35.6	SAR Design Examples	523
35.7	Summary	526
	Further Reading	526

PART VIII Radar and Electronic Warfare

Chapter 36	<i>Electronic Warfare Terms and Concepts</i>	531
36.1	EW Definitions	531
36.2	EW Subfields	531
36.3	Electronic Warfare Support	532
36.4	Electronic Attack	535
36.5	Electronic Protection	539
36.6	Decoys	539
36.7	Summary	540
	Further Reading	541
Chapter 37	<i>Electronic Warfare Support</i>	543
37.1	EW Antennas	543
37.2	EW Receivers	546
37.3	Receiver System Sensitivity and Dynamic Range	549
37.4	One-Way Radio Propagation	551
37.5	Passive Emitter Location	552
37.6	Search	559
37.7	Radar Warning Receivers	559
37.8	Summary	561
	Further Reading	562
Chapter 38	<i>Electronic Attack</i>	565
38.1	Jamming Geometry	565
38.2	Jamming Techniques	568
38.3	Deceptive Jamming Techniques Effective Against Monopulse Radars	574
38.4	Jamming Equations	578
38.5	Look through	581
38.6	Chaff	582
38.7	Anti-radiation missiles	583
38.8	High Power Lasers	584
38.9	High Power Microwave	584
38.10	Summary	584
	Further Reading	585
Chapter 39	<i>Electronic Protection</i>	587
39.1	Introduction	587
39.2	Ultra-low Sidelobes	587

PART IX Special Topics and Advanced Concepts

45.3	Categorization of Distributed Radar Systems	668	46.4	Waveform Diversity	688
45.4	The Distributed Radar Equation	672	46.5	Summary	691
45.5	Examples of Systems	675		Further Reading	691
45.6	MIMO Radar	677	Chapter 47	Target Classification	693
45.7	Summary	678	47.1	Introduction	693
			47.2	Classification Terminology	694
Chapter 46	Radar Waveforms: Advanced Concepts	681	47.3	Target Phenomenology	695
46.1	Practical Considerations	681	47.4	The Target Classification Processing Chain	697
46.2	Mismatch Filtering	684	47.5	Databases and Target Modeling	700
46.3	Nonlinear FM Waveforms	687	47.6	Summary	704
				Further Reading	705

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16

Pulse Compression and High-Resolution Radar



Original cavity magnetron, 1940

Ideally, to obtain both long detection range and fine range resolution, extremely narrow pulses (for fine resolution) of exceptionally high peak power (for long range) should be transmitted. However, there is a practical limit on the amount of peak power that subsequently limits the detection range. This peak power limit forces the use of long pulses at the expense of range resolution.

The solution to this dilemma is pulse compression, in which coding is modulated onto long, peak power-constrained pulses during transmit, followed by “compression” of the received echoes by decoding their modulation. This provides the necessary average power for an achievable level of peak power. This chapter introduces the fundamental principles of pulse compression and the various classes of modulation coding, otherwise known as the radar *waveform*.

16.1 Pulse Compression: A Beneficial Complication

Pulse compression might appear to be an unnecessary complication to the notion of how radar operates. Narrow pulses can easily provide the desired range resolution by setting the pulse width. For relatively short-range operation this arrangement is acceptable. However, if one wishes to have long-range detection capability, it becomes clear from the radar range equation (see Chapter 13) that increasingly high peak powers are necessary. However, there are practical limits to what can be made available from a realistic radar transmitter. The necessary extension to longer pulses subsequently establishes a set of trade-offs to design the appropriate transmitted signal and



Figure 16-1. With a short pulse, closely spaced targets can be resolved. However, the limit on peak power likewise limits the maximum detectable range.



Figure 16-2. With the increased energy from a long pulse, the maximum detection range can be extended. Pulse compression is now required to separate the closely spaced targets.

receive filtering to perform the desired radar sensing function. It is also worth noting that echo-locating mammals seemed to have developed this capability long before radar engineers ever thought of it.

The Pulse Width Dilemma. Figure 16-1 illustrates an example of the transmission of a short pulse (at peak power) and the resulting echoes from two targets that are closely spaced in range. As long as these targets are separated by more than the pulse width it is possible to distinguish one from the other. However, because there is a limit to the amount of peak power the transmitter can achieve, this short pulse approach severely limits the maximum range from which targets can be reliably detected.

To extend the maximum detection range, more energy is required to be “put on the target.” Since the peak power is bounded, the pulse width must be increased (in time). Figure 16-2 shows an example of what occurs when the short pulse from Figure 16-1 is extended in time (pulse width) by a factor of 5. The energy that is incident onto, and thereby reflected from, a target also increases by a factor of 5, thus extending the maximum detection range. However, now there is overlap between the two closely spaced targets such that they cannot be distinguished from one another. The solution to this problem is pulse compression.

Waveforms. In radar, the waveform is simply the transmitted signal. This may be a continuous signal or it may be a pulse. The notion of a radar waveform is extended here to include a modulation imparted upon a pulse. In principle, this modulation could be in terms of frequency/phase, amplitude, or polarization, though the former is by far the most common. Taken as a whole, pulse compression involves the transmission of a modulated, pulsed waveform followed by filtering of the received echoes, where the filter is coherently matched to the waveform.

There are often numerous objectives to be considered when designing a waveform, including

- Total energy of the modulated pulse (this relates to the SNR of received echoes)
- Discrimination between delay-shifted versions of the waveform (for both range resolution and sensitivity)
- Impact of Doppler shift
- Low probability of intercept (by a potential adversary)

The pulse energy is maximized when the amplitude envelope of the pulse is constant. The delay and Doppler characteristics of a waveform are collectively referred to as the waveform *ambiguity function* (see Chapter 11). The intercept probability of a waveform is dependent upon whether it appears to be man-made or naturally occurring (noise radar is an example of the latter).

Most commonly, a waveform can be ascribed to one of the following classes: frequency modulated chirp (linear or nonlinear) or phase-coded waveform (biphase or polyphase).

Linear frequency modulation (LFM) chirp is the most widely used of all waveforms due to its simplicity of implementation on transmit, its robustness to Doppler shift, and the existence of a useful wideband receiver filtering structure known as *stretch processing*. However, due to relatively high time-delay (range) sidelobes resulting from LFM matched filtering, nonlinear frequency modulation (NLFM) and phase-coded waveforms have been devised as possible alternatives.

The Matched Filter. What actually happens when an echo passes through a filter that is matched to the transmitted waveform can be visualized if the echo is thought of as consisting of a sequence of subpulses, or *chips*, each with a distinct phase. As depicted in Figure 16-3 the matched filter is likewise a sequence of chips, though each possesses the conjugate phase (i.e., reflected about the real, or horizontal, axis). If the aligned sets of chips are piecewise multiplied, they all produce the same value (here, set to an arbitrary phase of $e^{j0} = 1$ for simplicity) such that they add constructively in phase.

Figure 16-3 depicts the precise point in time when an echo aligns with the matched filter, thus producing a gain on the echo. At other delay shifts a different phenomenon is observed. For example, Figure 16-4 illustrates what occurs when the echo is shifted in time by just one chip interval compared with the matched case of Figure 16-3. This time, when the aligned chips are piecewise multiplied, a set of phase values is produced that are out of phase with each other and thus combine destructively when added. The resulting summation will typically be much smaller than the matched case of Figure 16-3. For other delays, different sets of phases are produced by the matched filter, which subsequently yields different destructive combinations that vary as a function of delay in a way that is characteristic to each individual waveform.

In reality, a physical waveform must be continuous. For the discrete illustrations in Figures 16-3 and 16-4 the chips can be thought of as representing basic phase shapes that enable the adjacent chips to connect in a continuous manner over the extent of the waveform (and likewise the matched filter). When considered in this way, the matched filter concept extends to all types of frequency modulated and phase-coded waveforms.

It is becoming increasingly common to perform matched filtering digitally, thus requiring sampling of the received echoes and a digital representation of the filter. The determination of the sampling rate involves a trade-off between higher computational complexity and the acceptable degree of loss from *range straddling* (also known as *range cusping*) that occurs when an echo is not sampled precisely at its matched position.

The continuum of delay shifts comprising the matched filter response to a single echo (with no Doppler) is actually the

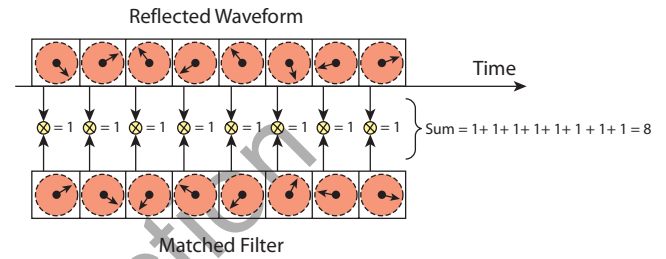


Figure 16-3. For a waveform represented as a sequence of 8 chips, the matched filter constructively combines the segments to yield a processing gain, also known as the *pulse compression ratio*, of 8.

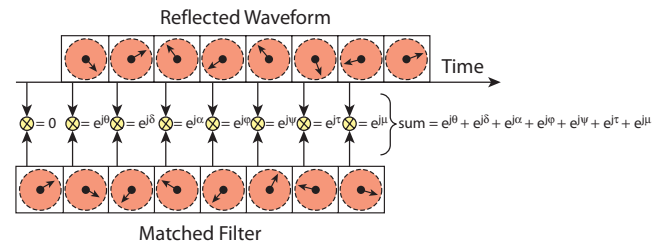


Figure 16-4. For delays different from the match point, the segments of the echo do not match the phase sequence of the matched filter, thereby combining destructively to produce a smaller value (here, much less than 8).

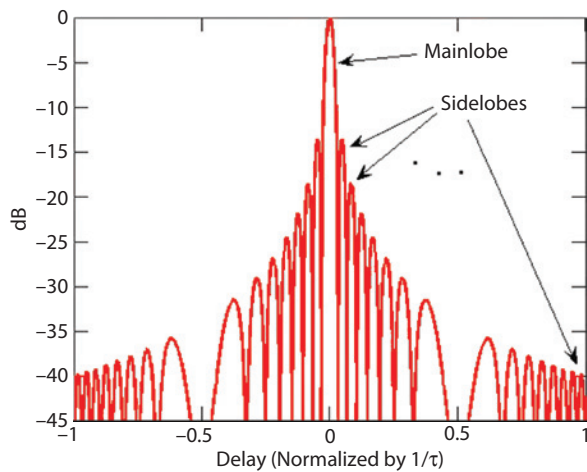


Figure 16-5. The matched filter response (waveform autocorrelation) for an LFM chirp with uncompressed pulse width τ illustrates the mainlobe and sidelobes in delay that would result from a single target echo.

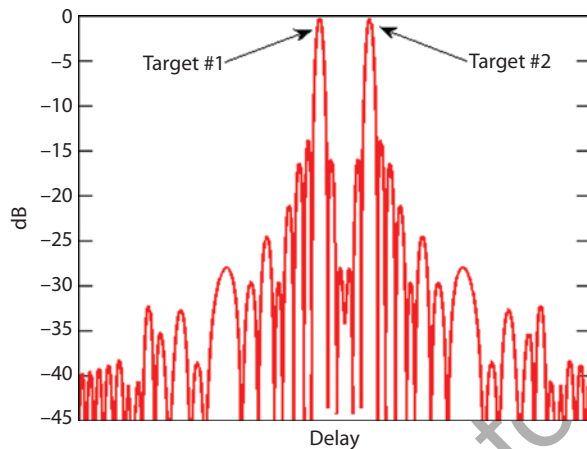


Figure 16-6. The echoes from two closely spaced targets may be resolved if they have similar receive powers and are not too close together.

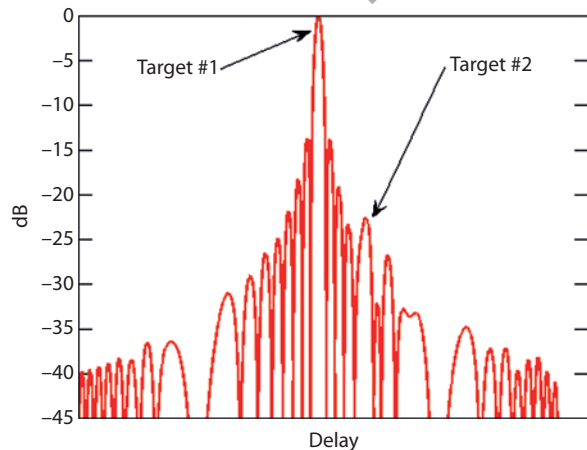


Figure 16-7. If the receive echoes from two closely spaced targets have sufficiently disparate receive powers, then the smaller target may be lost among the range sidelobes of the larger target.

autocorrelation of the transmitted waveform. For example, Figure 16-5 illustrates the autocorrelation for an LFM chirp that is normalized so the match point is at 0 dB. Figure 16-5 shows the ideal pulse compressed response having -13 dB peak sidelobes (these may be reduced using weighting; see the section “Amplitude Weighting”).

Resolution and Range Sidelobes. Similar to an antenna radiation pattern, the matched filter mainlobe is the delay region immediately surrounding the matched position. Using the LFM matched filter response in Figure 16-5 as an example, it is predominantly the width of the mainlobe that determines if two closely spaced targets in range can be resolved. Therefore, if the matched filter is applied to the echoes generated by the two targets of Figure 16-2 (assuming the pulse was modulated with an LFM waveform), the pulse compressed output would look like the result shown in Figure 16-6.

As it is much shorter than the pulse width, the width of the mainlobe enables improved range resolution. The range resolution is now inversely proportional to the bandwidth of the waveform. A convenient point of reference is that a range resolution of 30 cm corresponds to a waveform bandwidth of approximately 500 MHz.

Referring again to Figure 16-5, the smaller peaks surrounding the mainlobe are known as *range sidelobes*. For the LFM chirp, the largest sidelobe is approximately 13 dB lower than the value at the matched position and defines the *peak sidelobe level* (PSL). Range sidelobes are one of the performance trade-offs of pulse compression, as they limit the sensitivity of the radar. For example, if the received power of the two target echoes depicted in Figure 16-6 were very different, the matched filter response would instead look like the result in Figure 16-7, in which the range sidelobes induced by the higher-power target can actually mask the mainlobe of the lower-power target.

Doppler Effects and the Ambiguity Function. The discussion thus far has been limited to the case where no Doppler effects are present. Doppler is a shift in frequency that is induced by radial motion between the radar and the subject of the radar illumination (see Chapter 18 for a detailed discussion). For example, a police radar measures the amount of frequency shift of the echo from a moving vehicle to measure its speed relative to the position of the radar. Relative motion towards the radar causes a positive frequency shift (i.e., a higher frequency echo), while relative motion away from the radar causes a negative shift (i.e., a lower frequency echo).

With regard to pulse compression, the impact of motion-induced Doppler frequency shift is an altering of the phase progression of the waveform echo. As a result, the gain from constructive combining at the matched position (see Figure 16-3) can be degraded or even completely lost depending on the degree of Doppler shift and the nature of the waveform.

A plot of the matched filter response versus Doppler frequency shift is shown in Figure 16-8. This is defined as the *ambiguity function* (see Chapter 11).

The matched position is located where both delay and Doppler are zero. The zero Doppler cut (horizontally across Doppler = 0 Hz) reveals the waveform autocorrelation (and is the same result shown in Fig. 16-5). In the Doppler dimension the mainlobe width is inversely proportional to the pulse width. Away from the mainlobe, range-Doppler sidelobes can be observed.

In current fielded radar systems, the two most commonly employed waveforms are the LFM chirp and the biphase (or binary phase)-coded waveform. The following sections outline the benefits and deficiencies of each.

16.2 Linear Frequency Modulation (Chirp)

Because of its similarity to the chirping of a bird, its inventors called this form of modulation a “chirp.” Since it was the first pulse compression technique, the term chirp is still in common usage and is synonymous with pulse compression.

For LFM chirp coding, the frequency of the transmitted pulse is increased (an “up-chirp”) or decreased (a “down-chirp”) at a constant rate throughout its length (see Figure 16-9), thus every echo has the same linear increase/decrease in frequency.

LFM Implementation. A major benefit of LFM chirp is the ease with which it can be implemented. The transmitter needs only to sweep linearly from some starting frequency at the beginning of the pulse to some ending frequency at the tail of the pulse, which can be accomplished in many different ways in both analog and digital hardware.

Filtering may be done with an analog device—such as an acoustical delay line—or, more common in modern systems, digitally. For a narrow range swath the LFM chirp can be decoded using a technique called *stretch processing*, which can accommodate a very large waveform bandwidth, thus enabling very fine range resolution.

For stretch processing (described in detail in the accompanying panel, the echo delay time (range) is converted to frequency. As a result, the return from any one range corresponds to a constant frequency, and the returns from different ranges may be separated with a bank of narrowband filters implemented with the efficient fast Fourier transform (see Chapter 21). Range is determined by measuring the instantaneous difference between the frequencies of the transmitted and received signals.

Incidentally, stretch processing is similar to the FM ranging technique used by continuous wave (CW) radars (see Chapter 17). The principal differences are that instead of transmitting pulses, the CW radar transmits continuously, and the period over which the transmitter’s frequency changes in any one direction is many times the round-trip ranging time.

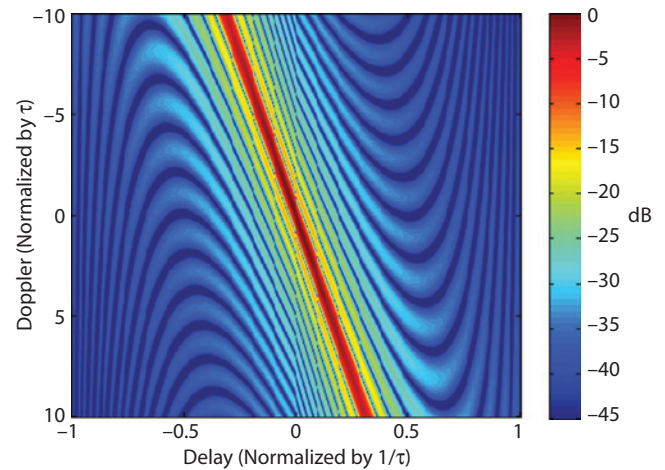


Figure 16-8. Delay/Doppler ambiguity function for the LFM chirp (brightness scale in decibels).

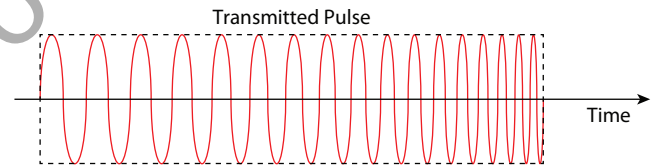
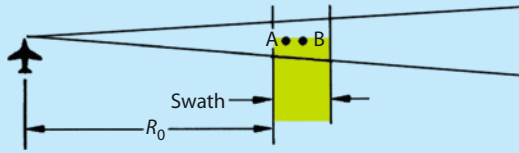


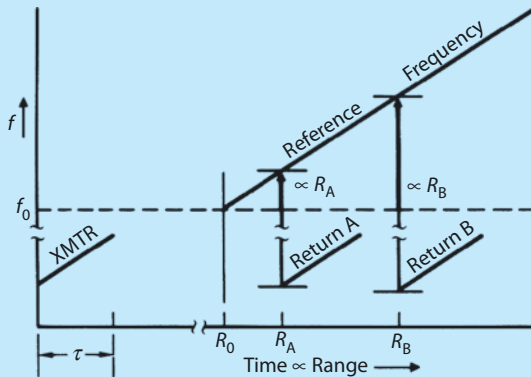
Figure 16-9. An LFM up-chirp.

Stretch Processing of LFM Chirp

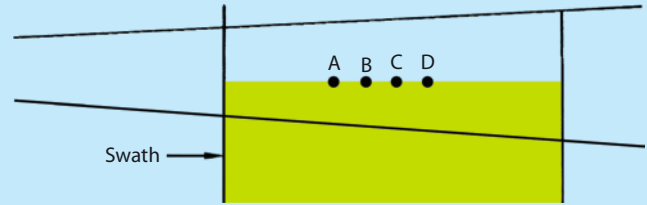
FOR A NARROW RANGE SWATH, SUCH AS IS MAPPED BY A SYNTHETIC aperture radar (see Chapter 33), LFM chirp modulation is commonly decoded by a technique called *stretch processing* or *deramping*.



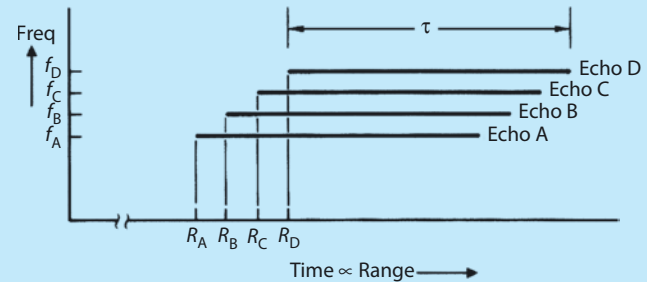
For the up-chirp example, as the return from the swath is received, its frequency is subtracted from a reference frequency that increases at the same rate as the transmitter frequency. However, the reference frequency increases continuously throughout the entire interval over which echoes are received.



Consequently, the difference between the reference frequency and the frequency of the return from any particular point on the ground is constant. Moreover, as can be seen from the above figure, if we subtract the reference frequency's initial offset, f_0 , from the difference already obtained, the result is proportional to the range of the point from the near edge of the swath, R_0 . Range is thus converted to frequency.



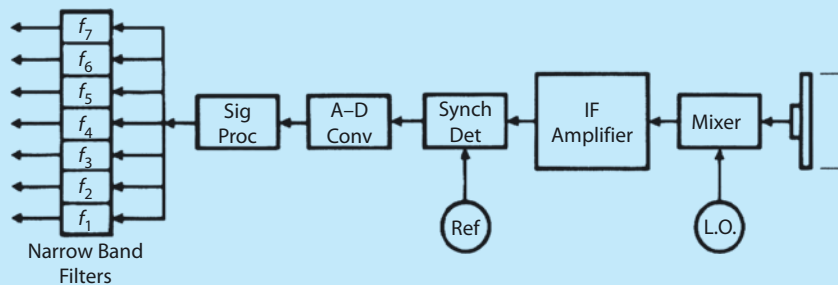
To see how fine resolution is achieved, consider the returns from four closely spaced points after the subtraction has been performed.



Although the returns were received such that their pulse echoes almost completely overlap, the slight stagger in their arrival times results in clearly discernible differences in frequency.

As indicated in the figure below, the continuously changing reference frequency may be subtracted at one of three points in the receiving system. One is the mixer, which converts the radar returns to the receiver's intermediate frequency (IF). The second point is the synchronous detector, which converts the output of the IF amplifier to video frequencies. And the third point is in the signal processor, after the video has been digitized.

To sort the difference frequencies, the video output of the synchronous detector is applied to a bank of narrowband filters, implemented with the fast Fourier transform.



Pulse Compression Ratio. The straightforward nature of the LFM provides a convenient framework with which to better explain the processing gain and range resolution enhancement provided by pulse compression. The pulse compression ratio, represented by the factor 8 for the example in Figure 16-3, is the ratio of the uncompressed pulse width τ to the compressed width τ_{comp} . Whereas the previous example explained the phenomenon in terms of phase-modulated subpulses, LFM chirp allows us to consider it in terms of frequency sensitivity.

If returns received simultaneously from two slightly different ranges are to be separated on the basis of the difference in their frequencies, besides providing a delay proportional to frequency (refer to the panel on stretch processing), a second requirement must also be satisfied. The frequency difference must be large enough for the signals to be resolved by the filter.

As will be made clear in Chapter 20, the frequency resolution of the matched filter response increases (becomes narrower) as the uncompressed pulse width increases (see Figure 16-10). Specifically, the frequency resolution Δf is related to the uncompressed pulse width as

$$\Delta f = \frac{1}{\tau}.$$

In other words, as illustrated by Figure 16-11, for the LFM matched filter to resolve two closely spaced echoes, the instantaneous difference in their delay-shifted frequencies must meet or exceed the inverse of the uncompressed pulse width τ .

Furthermore, the compressed pulse width τ_{comp} is the period of time over which the frequency of the uncompressed LFM pulse changes by Δf (see Figure 16-12). By extension, if the frequency of the uncompressed LFM pulse changes at a rate of $\Delta f/\tau_{\text{comp}}$ (in hertz per second), then the total change in frequency, ΔF , over

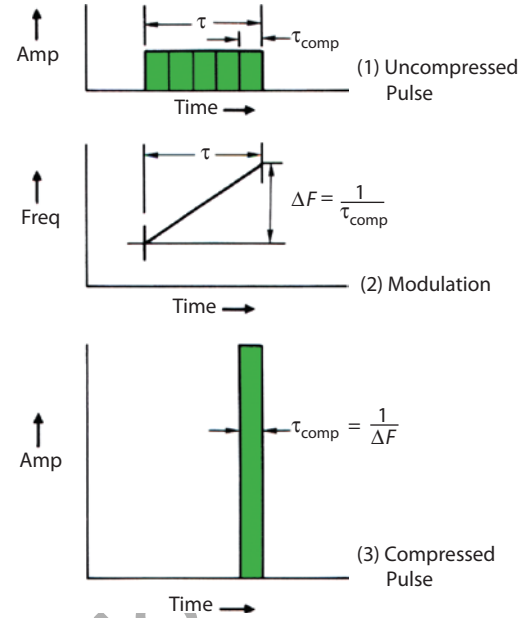


Figure 16-10. Conceptual relationship between uncompressed pulse width, chirp modulation bandwidth ΔF , and compressed pulse width for an LFM waveform. The compressed pulse width corresponds to the mainlobe from Figure 16-5.

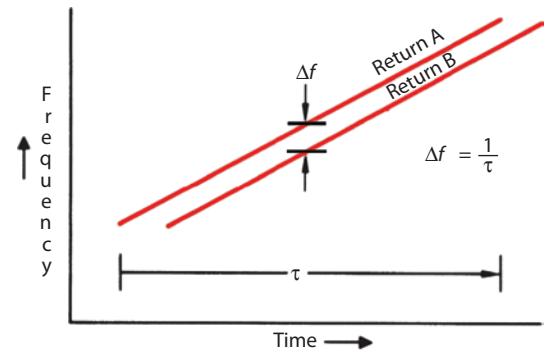


Figure 16-11. For a filter to resolve two concurrently received LFM returns, the instantaneous difference in their frequencies (Δf) must equal at least $1/\tau$.

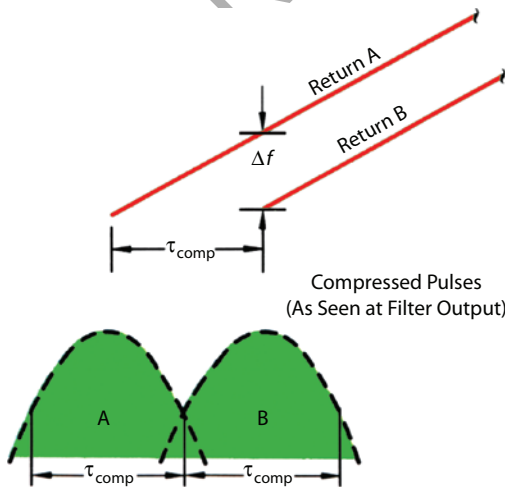


Figure 16-12. If the minimum resolvable frequency difference is Δf , the time in which the frequency of the uncompressed LFM pulse changes by Δf is the width of the compressed pulse, τ_{comp} .

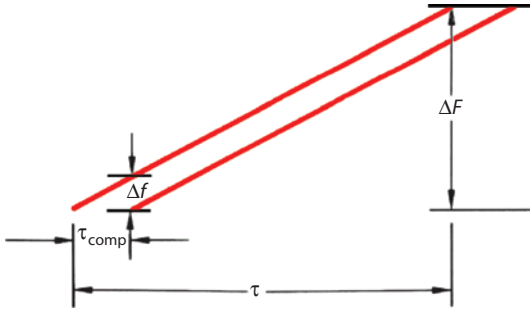


Figure 16-13. The ratio of uncompressed pulse width, τ , to compressed pulse width, τ_{comp} , equals the ratio of the total change in frequency over the pulse width, ΔF , to minimum resolvable frequency difference, Δf .

the duration of the uncompressed pulse will be this rate times the uncompressed pulse width, τ .

The rate of frequency change $\Delta f/\tau_{\text{comp}}$ is known as the *chirp rate*:

$$\text{chirp rate} = \frac{\Delta f}{\tau_{\text{comp}}} \text{ (in hertz per second).}$$

The total change in frequency, ΔF , is the *bandwidth* for the LFM chirp:

$$\text{bandwidth} = \Delta F = \left(\frac{\Delta f}{\tau_{\text{comp}}} \right) \tau = \text{chirp rate} \times \tau.$$

As is apparent from the geometry of Figure 16-13, the pulse compression ratio, τ/τ_{comp} , equals the ratio of ΔF to Δf :

$$\text{Pulse compression ratio} = \frac{\tau}{\tau_{\text{comp}}} = \frac{\Delta F}{\Delta f}.$$

Substituting $1/\tau$ for Δf , the pulse compression ratio equals the uncompressed pulse width times ΔF :

$$\text{Pulse compression ratio} = \tau \Delta F.$$

The quantity $\tau \Delta F$ is also called the *time–bandwidth product*.

This simple relationship—pulse compression ratio equals time–bandwidth product—tells us a lot about the LFM chirp. To begin with, for a given uncompressed pulse width τ , the compression ratio increases directly with an increase in bandwidth ΔF . Conversely, for a given bandwidth ΔF , the compression ratio increases directly with an increase in the uncompressed pulse width τ .

If the time–bandwidth product is set equal to τ/τ_{comp} as

$$\tau \Delta F = \frac{\tau}{\tau_{\text{comp}}},$$

τ cancels out so that

$$\tau_{\text{comp}} = \frac{1}{\Delta F}.$$

In other words, the width of the compressed pulse is determined entirely by the bandwidth ΔF of the transmitted pulse; that is, the greater the frequency change, the narrower the compressed pulse width. Rearranging this last equation tells us that the total change in transmitter frequency (the LFM bandwidth) must be

$$\Delta F = \frac{1}{\tau_{\text{comp}}}.$$

This relationship provides a useful benchmark for the transmitter bandwidth necessary to achieve a desired bandwidth (and therefore range resolution) for arbitrary waveforms. It should be noted, however, that the equality only holds for the LFM chirp, which spends an equal amount of time (and thus power) in each of the frequencies due to its linear frequency sweep. Different waveforms that occupy some frequencies longer than

others or employ a weighting across frequencies may require a higher bandwidth to achieve the same compressed pulse width as the LFM.

To get a feel for the relative values involved for LFM chirp, consider a couple of representative examples.

- Using LFM to provide the same compression as in the 8-chip matched filter discussed earlier, $\tau/\tau_{\text{comp}} = 8$. If the original pulse is $1 \mu\text{s}$, the range resolution has now improved to 18.75 m. This would separate aircraft targets except for very tight formations of small planes.
- It is assumed that in order to provide adequate “energy on target,” the width of a radar’s transmitted pulse must be $\tau = 10 \mu\text{s}$. To provide the desired range resolution of 1.5 m, a compressed pulse width of $\tau_{\text{comp}} = 0.01 \mu\text{s}$ is required. Therefore the pulse compression ratio must be

$$\frac{\tau}{\tau_{\text{comp}}} = \frac{10}{0.01} = 1000$$

To achieve a compressed pulse width of $0.01 \mu\text{s}$ (10^{-8} s), the change in transmitter frequency, ΔF , over the duration of each transmitted pulse must be $1/10^{-8} = 10^8$ Hz, or a bandwidth of 100 MHz.

Since the duration of the uncompressed pulse is $10 \mu\text{s}$ (10^{-5} s), the rate of change of the transmitter frequency (the chirp rate) will be $10^8/10^{-5} = 10^{13}$ Hz/s, or 10,000 GHz/s. This arrangement equates to a total linear frequency modulation excursion of 100 MHz over the duration of the pulsed waveform.

Incidentally, these values explain why stretch processing is practical only for relatively narrow range intervals. The ranging time for an interval of 100 km, for instance, is $13.3 \times 50 = 665 \mu\text{s}$. If the receiver local-oscillator frequency is shifted at a rate of 10 GHz/s throughout that time (see Figure 16-14), the total frequency shift would be $10,000 \times 620 \times 10^{-6} = 6.65$ GHz. Such a large shift was deemed impractical in previous editions of this book and is only now beginning to enter the realm of possibility.

LFM Ambiguity Function. LFM allows very large compression ratios to be achieved with a relatively simple implementation. To assess the performance capability of LFM, consider its delay and Doppler characteristics that were illustrated via the ambiguity function in Figures 16-5 and 16-8.

One disadvantage of LFM is the high sidelobes that occur in the range dimension. These high sidelobes have driven the development of alternative waveforms and filtering strategies.

Another possible disadvantage is the ambiguity that occurs between range and Doppler (the range–Doppler ridge), shown in Figure 16-8. If an echo possesses a sufficient Doppler shift, it will also appear to be shifted in range, thereby limiting the ultimate accuracy for which true range may be determined. However, this *Doppler tolerance* also allows for simpler receiver

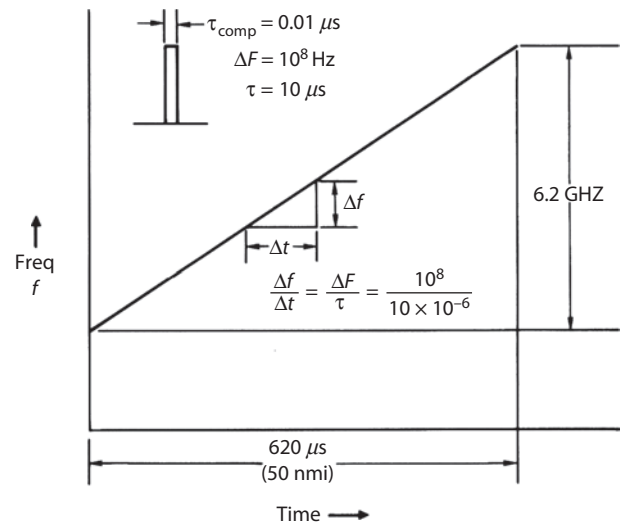


Figure 16-14. If stretch processing is used over a 100 km range interval to decode a $10 \mu\text{s}$ pulse modulated for a 1000:1 compression ratio, the receiver local oscillator would have to be swept over 6.65 GHz.

processing by precluding the need for a bank of matched filters tuned to different possible Doppler frequencies.

Amplitude Weighting. A well-known approach to reduce range sidelobes for LFM is to apply an amplitude weighting to the waveform that reduces the power in the regions nearer the ends of the pulse. Due to the frequency-swept nature of LFM, this weighting results in a deemphasizing of the frequencies near the extremities of the bandwidth, which results in lower sidelobes in the time domain due to the relationship between time and frequency (from Chapter 6).

The trade-off for this significant reduction in range sidelobes is reduced transmit power, which directly impacts detection sensitivity. This weighting can also cause a degradation in range resolution due to reduced power in the outer frequencies, which is essentially a reduction in bandwidth. A typical compromise is to allow the resolution size to increase by approximately 50%, which enables sidelobe levels to be around -35 to -40 dB or less.

From an implementation standpoint, weighting the transmitted pulse may also be prohibitive if high-efficiency, nonlinear amplification is required.

A common compromise is to transmit the standard LFM waveform while applying a receive filter that is weighted. This form of *mismatched filtering* has the advantage of still enabling the maximum power on transmit as well as power-efficient nonlinear amplification. The trade-off is a small mismatch loss between the waveform and filter that is acceptable for many radar applications.

16.3 Phase Modulation

In this type of coding the waveform is represented as a discrete sequence of increments, with each increment corresponding to one from a set of phase values modulated onto a *subpulse* (or *chip*). The set of possible phase values is often referred to as the phase *constellation*. For practical reasons, it is often desirable for the nature of the subpulse shape to provide continuous transitions between adjacent subpulses.

Binary Phase Modulation. The simplest form of phase modulation employs a constellation of two opposite phase values (usually 0° and 180°) that are modulated onto a subpulse. The radio frequency phase of certain subpulse segments is shifted by 180° (or -1), according to a predetermined binary code. The subpulse is comprised of a multiple number of wavelengths of the carrier frequency.

Figure 16-15 illustrates an exemplary three-segment code. (So you can readily discern the phases, the wavelength has been arbitrarily increased to the point where each segment contains only one cycle.)

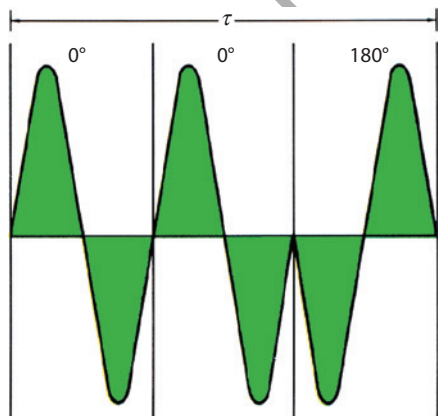


Figure 16-15. Binary phase coding of a transmitted pulse. The pulse is marked off into segments and the phases of certain segments (here, the third) are reversed.

A common shorthand method of indicating the coding is to represent the segments with + and – signs. An unshifted segment (0°) is represented by a + sign and a shifted segment (180°) by a – sign. The signs making up the code are referred to as digits. The number of digits indicates the pulse compression ratio of the code.

The received echoes are passed through a tapped delay line (Figure 16-16) that provides a time delay exactly equal to the duration of the uncompressed pulse, τ . The delay line may be implemented either with an analog device or digitally. Clearly the tapped delay line for the binary-coded waveform is an implementation of the matched filter previously shown in Figures 16-3 and 16-4.

Like the transmitted pulse, the delay line is divided into segments. An output tap is provided for each segment. The taps are all tied to a single output terminal. At any instant, the signal at the output terminal corresponds to the sum of whatever segments of a received pulse currently occupy the individual segments of the line.

Now, in certain taps, 180° phase reversals are inserted. Their positions correspond to the positions of the phase-shifted segments in the transmitted pulse. Thus when a received echo has progressed to the point where it completely fills the line, the outputs from all of the taps will be in phase (Figure 16-17). Their sum will then equal the amplitude of the pulse times the number of segments it contains.

To see step by step how the binary-coded pulse is compressed, consider a simple three-segment delay line and the three-digit code, illustrated in Figure 16-15.

Suppose the echo from a single-point target is received. Initially the output from the delay line is zero. When segment 1 of the echo has entered the line, the signal at the output terminal corresponds to the amplitude of this segment (Figure 16-18). Since its phase is 180° , the output is negative: -1 .

An instant later, segment 2 has entered the line. Now the output signal equals the sum of segments 1 and 2. Since the segments are 180° out of phase, however, they cancel: the output is 0.

When segment 3 has entered the line, the output signal is the sum of all three segments. At this point segment 1 has reached the tap containing the phase reversal. The output from this tap is, therefore, in phase with the unshifted segments 2 and 3 such that the combined output of the three taps is three times the amplitude of the individual segments: $+3$.

As segments 2 and 3 pass through the line, this process continues. The output drops to 0, then becomes -1 , and finally returns to 0 again.

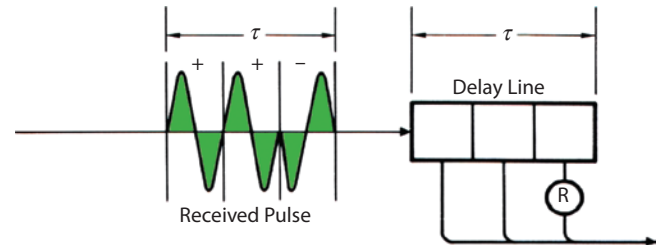


Figure 16-16. Received pulse echoes are passed through a tapped delay line filter. A separate tap is provided for each segment of the pulse. Here, the third tap is reversed R to represent a 180° phase shift.

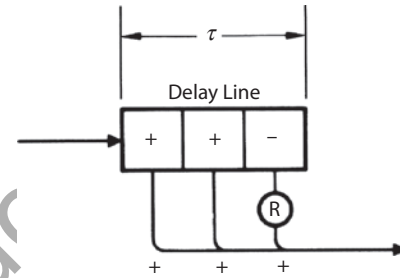


Figure 16-17. The phase reversal \oplus is placed so that when a pulse completely fills the delay line, outputs from all taps will be in phase.

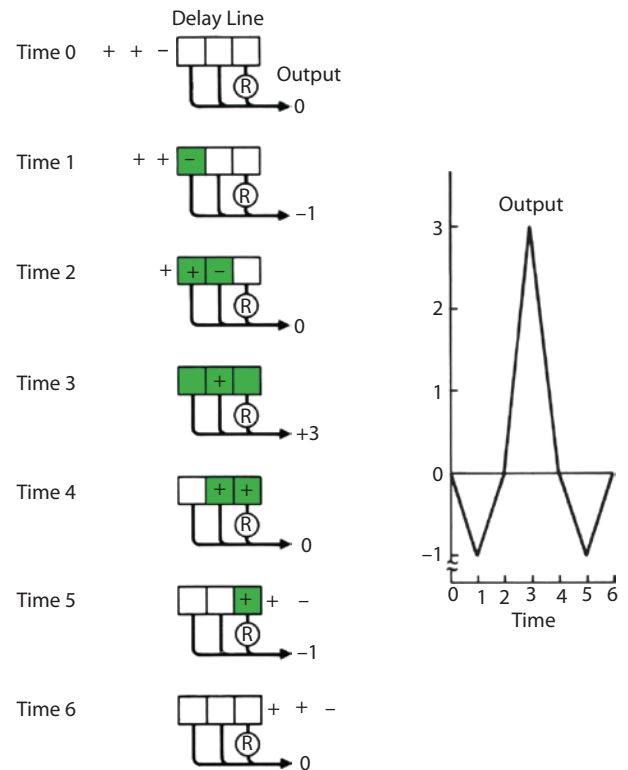


Figure 16-18. Step-by-step progress of a three-digit binary phase modulated pulse through a tapped delay line.

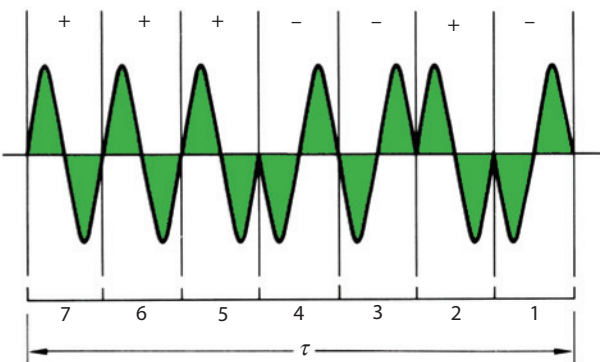


Figure 16-19. A seven-digit binary phase code.

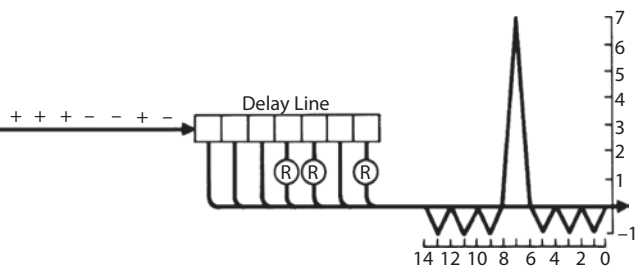


Figure 16-20. Output produced when a seven-digit binary phase code is passed through a tapped delay line with phase reversals in the appropriate taps.

A somewhat more practical example is shown in Figure 16-19. This code has seven digits. Assuming no losses, the peak amplitude of the compressed pulse is seven times that of the uncompressed pulse, and the compressed pulse is only one-seventh as wide.

To see why the code produces the output it does, transfer the code to a sheet of paper and slide it across the delay line plotted in Figure 16-20, digit by digit, noting the sum of the outputs for each position. (A minus sign, $-$, over a tap with a reversal \oplus in it becomes a $+$, while conversely the reversal of a $+$ becomes a $-$.) You should obtain the output shown in the figure.

Barker Codes. Ideally, for all positions of the echo in the line—except the central one—the collection of 0° or 180° outputs would cancel and there would be no range sidelobes.

One set of codes, called the Barker codes, comes very close to meeting this goal (Figure 16-21). Two of these codes have been used in the above examples. As has been seen, they produce sidelobes whose amplitudes are no greater than the amplitude of the individual code segments. Consequently the ratio of mainlobe amplitude to sidelobe amplitude, as well as the pulse compression ratio, increases with the number of segments into which the pulses are divided—that is, the number of digits in the binary code.

Unfortunately, the longest Barker code contains only 13 digits. Arbitrary binary codes can be made practically any length, but their sidelobe characteristics, though reasonably good, do not possess this desirable property of the Barker codes. Such codes require an exhaustive computer search and are called *minimum peak sidelobe* codes.

Complementary Codes. It turns out that the four-digit Barker code has a special feature that enables us to build codes of greater length and even eliminate the sidelobes altogether (under certain conditions). This feature arises because the four-digit code, as well as the two-digit code, has a complementary form. The sidelobe structures produced by the complementary

N	Barker Codes
2	+ - Or (+ +)
3	+ + -
4	+ - + - Or (+ - - -)
5	+ + + - +
7	+ + + - - + -
11	+ + + - - - + - - + -
13	+ + + + + - - + - + - +

Note: Plus and minus signs may be interchanged
(+ + - changed to - - +); order of digits may be reversed
(+ + - changed to - + +). Codes in parentheses are complementary codes.

Figure 16-21. Barker codes come very close to the goal of producing no sidelobes. However, the largest Barker code contains only 13 digits.

forms have opposite phases (Figure 16-22). Therefore, if successive transmitted pulses are alternately modulated with the two forms of the code and filter each with their corresponding delay line, the returns from successive pulses can be added such that the sidelobes cancel.

Furthermore, by chaining the complementary forms together according to a certain pattern, codes of much greater length can be built. As illustrated in Figure 16-23, the two forms of the four-digit code are just such combinations of the two forms of the two-digit code, and these are just such combinations of the two fundamental binary digits, + and –.

Unlike the unchained Barker codes, the chained codes (also called *nested codes*) produce sidelobes having amplitudes greater than one. However, since the chains are complementary, these larger sidelobes—like the others—cancel when successive pulses are added (at least in the absence of Doppler).

Doppler Sensitivity. Compared with LFM chirp, coded modulations can be much more sensitive to Doppler frequency shift. If all segments of a phase-coded pulse are to add constructively when the pulse is centered in the delay line, while cancelling when it is not, very little additional shift in phase over the length of the pulse can be tolerated.

A Doppler shift of 10 kHz amounts to a phase shift of $10,000 \times 360^\circ/\text{s}$, or $3.6^\circ/\mu\text{s}$. If the uncompressed pulse width is as much as 50 μs (Figure 16-24), this shift will itself equal 180° over the length of the pulse, and performance will deteriorate. For the scheme to be effective, either the Doppler shifts must be comparatively small or the uncompressed pulses must be reasonably short.

One way to contend with phase-coding sensitivity to Doppler is through “Doppler tuning,” in which a bank of Doppler-shifted versions of the delay line matched filter outputs are applied. While this approach increases the overall hardware (analog filtering) or computational (digital filtering) requirements, it does have the benefit of avoiding the range–Doppler ambiguity problem of the LFM chirp.

The sidelobe cancellation property of complementary code sets is very sensitive to pulse-to-pulse Doppler shift. The sidelobes do not perfectly subtract when Doppler is present, so residual sidelobes emerge.

Polyphase Codes. Phase coding is not limited to just two increments (0° and 180°). Codes with phase constellations comprised of more than two possible values are collectively referred to as *polyphase codes*. Here a particular example is considered and is taken from a family called Frank codes.

The fundamental phase increment ϕ for a Frank code is established by dividing 360° by the number of different phases in the constellation, P . The coded pulse is then built by chaining together P groups of P segments each. The total number of segments in a pulse, therefore, equals P^2 .

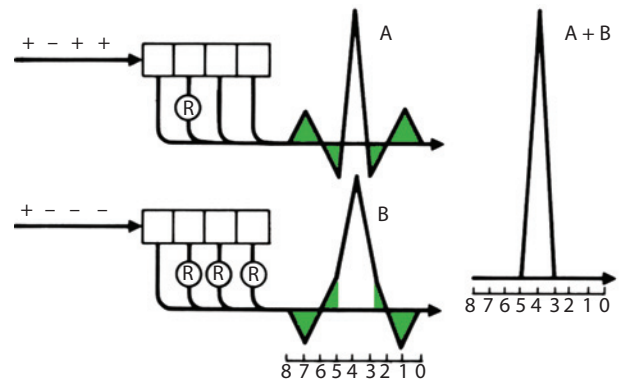


Figure 16-22. Echoes from complementary phase coding received from the same target in alternating pulses. When echoes are added, the time sidelobes cancel.

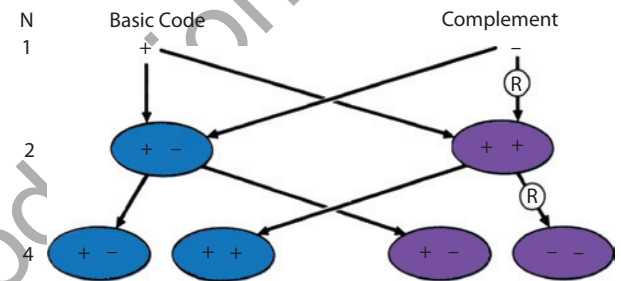


Figure 16-23. How complementary codes are formed. The basic two-digit code is formed by chaining a basic binary digit (+) to its complement (–). A complementary two-digit code is formed by chaining a basic binary digit (+) to its complement with the sign reversed (+). The basic four-digit code is formed by chaining a basic two-digit code (+–) to its complementary two-digit code (+). A complementary four-digit code is formed by chaining a basic two-digit code (+–) to its complementary two-digit code with the sign reversed (–), and so on.

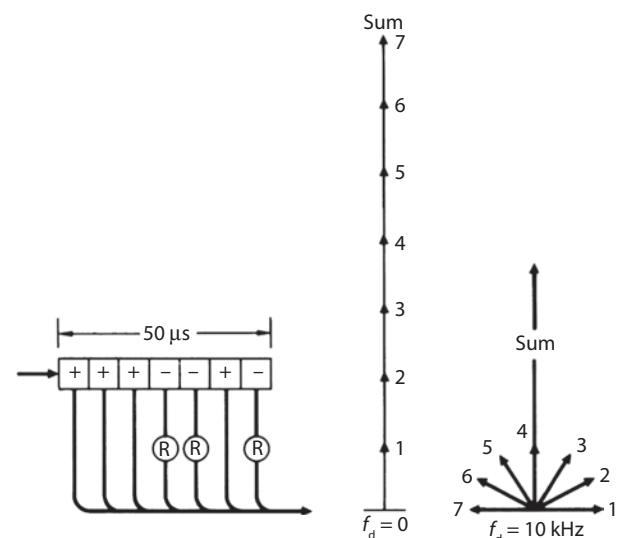


Figure 16-24. The reduction in peak output of a tapped delay line for a 50 μs , phase-coded pulse resulting from a Doppler shift of 10 kHz.

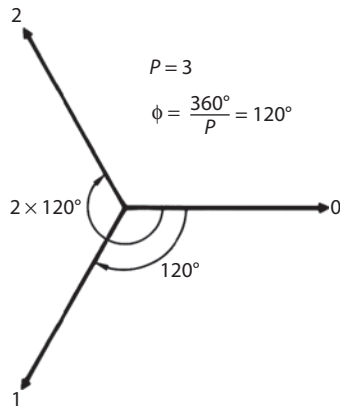


Figure 16-25. Phase increments for a Frank code in which the number of phases P is three.

In a three-phase code (Figure 16-25), for example, the fundamental phase increment is $360^\circ \div 3 = 120^\circ$, making the phases 0° , 120° , and 240° . The coded pulse consists of three groups of three segments—a total of nine segments.

Group 1			Group 2			Group 3		

Phases are assigned to the individual segments according to two simple rules: (1) the phase of the first segment of every group is 0° , that is, 0° ____, 0° ____, 0° ____, ____,; and (2) the phases of the remaining segments in each group increase in increments of

$$\Delta\Phi = (G - 1) \times (P - 1) \times \phi,$$

where

G = group number

P = number of phases

ϕ = basic phase increment.

For a three-phase code ($P = 3$, $\phi = 120^\circ$, $P - 1 = 2$), then $\Delta\Phi = (G - 1) \times 2\phi$. So the phase increment in Group 1 is 0° , the phase increment for Group 2 is 2ϕ , and the phase increment for Group 3 is 4ϕ .

Written in terms of ϕ , the nine digits of the code for $P = 3$ are

Group 1	Group 2	Group 3
0, 0, 0	0, 2ϕ , 4ϕ	0, 4ϕ , 8ϕ

Substituting 120° for ϕ and dropping multiples of 360° , the code becomes

Group 1	Group 2	Group 3
0° , 0° , 0°	0° , 240° , 120°	0° , 120° , 240°

Echoes are decoded by passing them through a tapped delay line (or the digital equivalent) in the same way as binary phase-coded echoes (Figure 16-26). The only difference is, the phase shifts in the taps have more than one value.

For a given number of segments, a Frank code provides the same pulse compression ratio as a binary phase code and the same ratio of peak amplitude to sidelobe amplitude as a Barker code. Yet, by using more phases (increasing P), the codes can be made of the greater length, P^2 . As P is increased, however, the size of the fundamental phase increment decreases, making performance more sensitive to externally introduced phase shifts (e.g., transmitter distortion) and imposing more severe restrictions on uncompressed pulse width and maximum Doppler shift.

Frank codes are an example of a class of codes in which the discrete phase sequence can be viewed as a sampled version of the LFM chirp. Other such codes are the Zadoff-Chu code, the “P” codes, and Golomb codes. Like the minimum peak

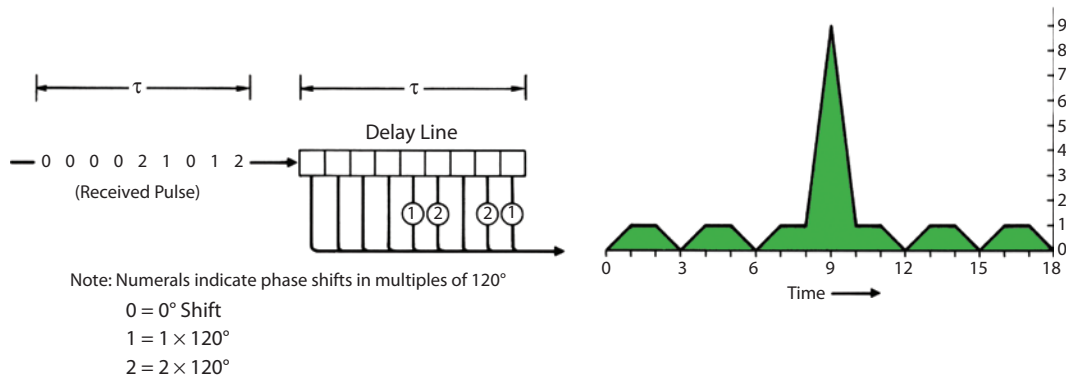


Figure 16-26. Processing of Frank codes is similar to that of binary codes. Phase shifts introduced in taps complement shifts in corresponding segments of the coded pulse. If the phase of a segment is shifted by $1 \times 120^\circ$, the corresponding tap adds a shift of $2 \times 120^\circ$, making the total shift when the pulse fills the line equal $3 \times 120^\circ = 360^\circ$. This phase relationship is identical to the matched filter.

sidelobe codes for the binary phase constellation, it is also possible to perform an exhaustive computer search for polyphase codes of arbitrary length and phase constellation.

While binary codes are in widespread use, the implementation of polyphase codes is more limited. The reason is that binary codes can be implemented in a phase-continuous manner in the transmitter while, until very recently, polyphase codes could not. These phase discontinuities at the chip transitions produce spectral spreading and can also limit the fidelity with which a polyphase-coded waveform can be generated by a practical transmitter. However, the design freedom provided by polyphase codes serves as the basis for new emerging radar capabilities. This topic is discussed further in Chapter 45.

16.4 Summary

Since radar transmitters are peak power limited, pulse compression provides the means to achieve sufficient energy on target for detection while enabling the requisite range resolution. Pulse compression comprises transmission of a modulated waveform and receiver filtering to compress the resulting echoes in range.

The most commonly used pulse compression techniques are the LFM chirp and binary phase coding.

With LFM, the frequency of each transmitted pulse is continuously increased or decreased. Applying the receive filter that is matched to the waveform results in a compressed pulse width of $1/\Delta F$, where ΔF is the total change in frequency (i.e., the bandwidth) of the waveform. The LFM range sidelobes may be reduced by amplitude weighting the receiver matched filter at a cost of reduced range resolution and mismatch loss.

When only a narrow range swath is of interest, the LFM chirp can be decoded using stretch processing, whereby range is converted to frequency in the receiver. Differences in frequency are resolved by a bank of tuned filters implemented with the efficient fast Fourier transform. With the LFM waveform and stretch processing, very large compression ratios and fine range resolution can be achieved. The LFM is rather

insensitive to Doppler frequency shift, though such a shift produces an ambiguity in range.

In binary phase modulation, each pulse is marked off into segments, with the phase of certain segments reversed. Received echoes are passed through a tapped delay line having phase reversals in taps corresponding to those in the code. Binary codes are more sensitive to Doppler frequency shift than the LFM chirp.

Barker codes represent a form of binary phase modulation in which the mainlobe-to-sidelobe ratio equals the pulse-compression ratio, though the longest Barker code is only 13 digits.

Sidelobes may be eliminated by alternately transmitting complementary codes that are obtained from chained Barker codes. However, this property requires little to no Doppler shift.

Polyphase (e.g., Frank) codes can also be used, but these are more sensitive to Doppler shift than binary codes, due to smaller phase increments. Polyphase codes tend to produce phase discontinuities, which results in spectral spreading and sensitivity to transmitter distortion.

Further Reading

1. N. Levanon and E. Mozeson, *Radar Signals*, John Wiley & Sons, 2004.
2. M. A. Richards, J. A. Scheer, and W. A. Holm, eds., *Principles of Modern Radar: Basic Principles*, SciTech, 2010, chap. 20.

Test your understanding

1. To achieve a range resolution of 0.5 m with a 20 μ s pulse, determine the required chirp rate and the associated pulse compression ratio.
2. Using the process described in Figure 16-18, determine the tapped delay line output for the following binary codes:
 - a. Length 11 Barker code as defined in Figure 16-21
 - b. $[+ + + - - - + - - + +]$ (Barker 11, but with the last digit flipped)
3. Using the tapped delay line outputs from problem 2, determine the largest sidelobe value relative to the mainlobe. This ratio is known as the PSL.
4. For $P = 4$ phase values, determine the length 16 Frank code and compute its output from the tapped delay line.



STIMSON'S Introduction to Airborne Radar Third Edition

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