DIGITAL COMMUNICATIONS

Fundamentals and Applications

Second Edition

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4.1 WHY MODULATE?

Digital modulation is the process by which digital symbols are transformed into waveforms that are compatible with the characteristics of the channel. In the case of baseband modulation, these waveforms usually take the form of shaped pulses. But in the case of *bandpass modulation* the shaped pulses modulate a sinusoid called a *carrier wave,* or simply a *carrier;* for radio transmission the carrier is converted to an electromagnetic (EM) field for propagation to the desired destination. One might ask why it is necessary to use a carrier for the radio transmission of baseband signals. The answer is as follows. The transmission of EM fields through space is accomplished with the use of antennas. The size of the antenna depends on the wavelength λ and the application. For cellular telephones, antennas are typically $\lambda/4$ in size, where wavelength is equal to c/f, and c, the speed of light, is 3×10^8 m/s. Consider sending a baseband signal (say, $f = 3000$ Hz) by coupling it to an antenna directly without a carrier wave. How large would the antenna have to be? Let us size it by using the telephone industry benchmark of $\lambda/4$ as the antenna dimension. For the 3,000 Hz baseband signal, $\lambda/4 = 2.5 \times 10^4\ m \approx 15$ miles. To transmit a 3,000 Hz signal through space, *without carrier-wave modulation,* an antenna that spans 15 miles would be required. However, if the baseband information is first modulated on a higher frequency carrier, for example a 900 MHz carrier, the equivalent antenna diameter would be about 8 cm. For this reason, carrier-wave or bandpass modulation is an essential step for all systems involving radio transmission.

Bandpass modulation can provide other important benefits in signal transmission. If more than one signal utilizes a single channel, modulation may be used to separate the different signals. Such a technique, known as *frequency-division multiplexing,* is discussed in Chapter 11. Modulation can be used to minimize the effects of interference. A class of such modulation schemes, known as *spread-spectrum modulation,* requires a system bandwidth much larger than the minimum bandwidth that would be required by the message. The trade-off of bandwidth for interference rejection is considered in Chapter 12. Modulation can also be used to place a signal in a frequency band where design requirements, such as filtering and amplification, can be easily met. This is the case when radio-frequency (RF) signals are converted to an intermediate frequency (IF) in a receiver.

4.2 DIGITAL BANDPASS MODULATION TECHNIQUES

Bandpass modulation (either analog or digital) is the process by which an information signal is converted to a sinusoidal waveform; for digital modulation, such a sinusoid of duration T is referred to as a digital symbol. The sinusoid has just three features that can be used to distinguish it from other sinusoids: amplitude, frequency, and phase. Thus bandpass modulation can be defined as the process whereby the amplitude, frequency, or phase of an RF carrier, or a combination of them, is varied in accordance with the information to be transmitted. The general form of the carrier wave is

$$s(t) = A(t) \cos \theta(t) \tag{4.1}$$

where $A(t)$ is the time-varying amplitude and $\theta(t)$ is the time-varying angle. It is convenient to write

$$\theta(t) = \omega_0 t + \phi(t) \tag{4.2}$$

so that

$$s(t) = A(t) \cos \left[\omega_0 t + \phi(t)\right] \tag{4.3}$$

where ω_0 is the *radian frequency* of the carrier and $\phi(t)$ is the *phase.* The terms f and ω will each be used to denote frequency. When f is used, frequency in hertz is intended; when ω is used, frequency in radians per second is intended. The two frequency parameters are related by $\omega = 2\pi f$.

The basic *bandpass modulation/demodulation* types are listed in Figure 4.1. When the receiver exploits knowledge of the carrier's phase to detect the signals, the process is called *coherent detection;* when the receiver does not utilize such phase reference information, the process is called *noncoherent detection.* In digital communications, the terms *demodulation* and *detection* are often used interchangeably, although demodulation emphasizes waveform recovery, and detection entails the process of symbol decision. In ideal coherent detection, there is available at the receiver a prototype of each possible arriving signal. These prototype waveforms attempt to duplicate the transmitted signal set in every respect, even RF phase. The receiver is then said to be *phase locked* to the incoming signal. During demodulation, the receiver multiplies and integrates (correlates) the incoming signal with each of its prototype replicas. Under the heading of coherent modulation/demodulation in Figure 4.1 are listed phase shift keying (PSK), frequency shift keying (FSK), amplitude shift keying (ASK), continuous phase modulation (CPM), and hybrid combinations. The basic bandpass modulation formats are discussed in this chapter. Some specialized formats, such as offset quadrature PSK (OQPSK), minimum shift keying (MSK) belonging to the CPM class, and quadrature amplitude modulation (QAM), are treated in Chapter 9.

Noncoherent demodulation refers to systems employing demodulators that are designed to operate without knowledge of the absolute value of the incoming signal's phase; therefore, phase estimation is not required. Thus the advantage of noncoherent over coherent systems is reduced complexity, and the price paid is increased probability of error (P_E). In Figure 4.1 the modulation/demodulation types that are listed in the noncoherent column, DPSK, FSK, ASK, CPM, and hybrids, are similar to those listed in the coherent column. We had implied that phase information is not used for noncoherent reception; how do you account for the fact that there is a form of phase shift keying under the noncoherent heading? It turns out that an important form of PSK can be classified as noncoherent (or differentially coherent) since it does not require a reference in phase with the received carrier. This "pseudo-PSK," termed *differential PSK* (DPSK), utilizes phase information of the prior symbol as a phase reference for detecting the current symbol. This is described in Sections 4.5.1 and 4.5.2.

4.2.1 Phasor Representation of a Sinusoid

Using a well-known trigonometric identity called Euler's theorem, we introduce the complex notation of a sinusoidal carrier wave as follows:

$$e^{j\omega_0 t} = \cos \omega_0 t + j \sin \omega_0 t \tag{4.4}$$

One might be more comfortable with the simpler, more straightforward notation $\cos \omega_0 t$ or $\sin \omega_0 t$. What possible benefit can there be with the complex notation? We will see (in Section 4.6) that this notation facilitates our description of how real-world modulators and demodulators are implemented. For now, let us point to the general benefits of viewing a carrier wave in the complex form of Equation (4.4).

First, within this compact from, $e^{j\omega_0 t}$, is contained the two important quadrature components of any sinusoidal carrier wave, namely the inphase (real) and the quadrature (imaginary) components that are orthogonal to each other. Second, the unmodulated carrier wave is conveniently represented in a polar coordinate system as a unit vector or phasor rotating counterclockwise at the constant rate of ω_0 radians/s, as depicted in Figure 4.2. As time is increasing (i.e., from t_0 to t_1) we can visualize the time-varying projections of the rotating phasor on the inphase (I) axis and the quadrature (Q) axis. These cartesian axes are usually referred to as the I channel and Q channel respectively, and the projections on them represent the

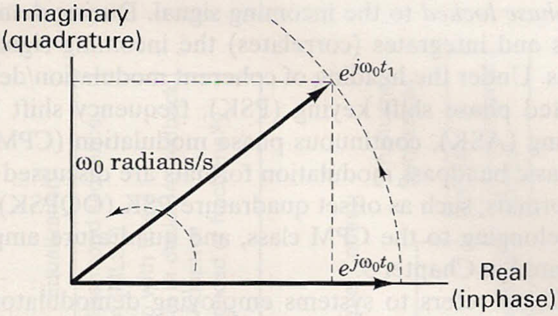


Figure 4.2 Phasor representation of a sinusoid.

signal components (orthogonal to each other) associated with those channels. Third, when it comes time to modulate the carrier wave with information, we can view this modulation as a methodical perturbation of the rotating phasor (and its projections).

For example, consider a carrier wave that is *amplitude modulated* (AM) with a sinusoid having an amplitude of unity and a frequency ω_m, where $\omega_m \ll \omega_0$. The analytical form of the transmitted waveform is

$$s(t) = \mathrm{Re}\left\{ e^{j\omega_0 t}\left(1 + \frac{e^{j\omega_m t}}{2} + \frac{e^{-j\omega_m t}}{2}\right)\right\} \tag{4.5}$$

where Re$\{x\}$ is the real part of the complex quantity $\{x\}$. Figure 4.3 illustrates that the rotating phasor $e^{j\omega_0 t}$ of Figure 4.2 is now perturbed by two sideband terms—$e^{j\omega_m t}/2$ rotating counterclockwise and $e^{-j\omega_m t}/2$ rotating clockwise. The sideband phasors are rotating at a much slower speed than the carrier-wave phasor. The net result of the composite signal is that the rotating carrier-wave phasor now appears to be growing longer and shorter pursuant to the dictates of the sidebands, but its frequency stays constant—hence, the term "amplitude modulation."

Another example to reinforce the usefulness of the phasor view is that of *frequency modulating* (FM) the carrier wave with a similar sinusoid having a frequency of ω_m radians/s. The analytical representation of *narrowband* FM (NFM) has an appearance similar to AM and is represented by

$$s(t) = \mathrm{Re}\left\{ e^{j\omega_0 t}\left(1 - \frac{\beta}{2} e^{-j\omega_m t} + \frac{\beta}{2} e^{j\omega_m t}\right)\right\} \tag{4.6}$$

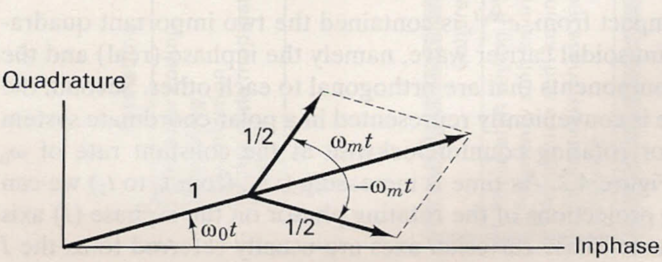


Figure 4.3 Amplitude modulation.

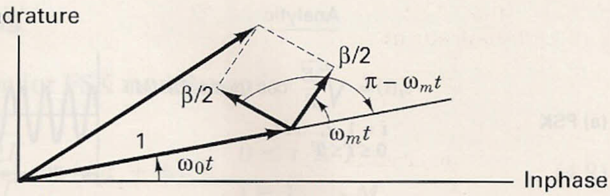


Figure 4.4 Narrowband frequency modulation.

where β is the modulation index [1]. Figure 4.4 illustrates that the rotating carrier-wave phasor is again perturbed by two sideband terms, but because one of the sideband terms carries a minus sign in Equation (4.6), the clockwise and counterclockwise rotating sideband phasors have a different symmetry than in the case of AM. In the case of AM the sideband symmetry results in the carrier-wave phasor growing longer and shorter with time. In NFM, the sideband symmetry (90° different than AM) results in the carrier-wave phasor speeding up and slowing down according to the dictates of the sidebands, but the amplitude says essentially constant—hence, the term "frequency modulation."

Figure 4.5 illustrates examples of the most common digital modulation formats: PSK, FSK, ASK, and a hybrid combination of ASK and PSK (ASK/PSK or APK). The first column lists the analytic expression, the second is a typical pictorial of the waveform versus time, and the third is a vector (or phasor) schematic, with the orthogonal axes labeled $\{\psi_j(t)\}$. In the general M-ary signaling case, the processor accepts k source bits (or channel bits if there is coding) at a time and instructs the modulator to produce one of an available set of $M = 2^k$ waveform types. Binary modulation, where $k = 1$, is just a special case of M-ary modulation.

In Figure 4.2, we represented a carrier wave as a phasor rotating in a plane at the speed of the carrier-wave frequency ω_0 radians/s. In Figure 4.5, the phasor schematic for each digital-modulation example represents a constellation of information signals (vectors or points in the signaling space), where time is not represented. In other words, the constantly rotating aspect of the unmodulated carrier wave has been removed, and only the information-bearing phasor positions, relative to one another, are presented. Each example in Figure 4.5 uses a particular value of M, the set size.

4.2.2 Phase Shift Keying

Phase shift keying (PSK) was developed during the early days of the deep-space program; PSK is now widely used in both military and commercial communications systems. The general analytic expression for PSK is

$$s_i(t) = \sqrt{\frac{2E}{T}} \cos\left[\omega_0 t + \phi_i(t)\right] \qquad \begin{array}{l} 0 \le t \le T \\ i = 1, \ldots, M \end{array} \tag{4.7}$$

where the phase term, $\phi_i(t)$, will have M discrete values, typically given by

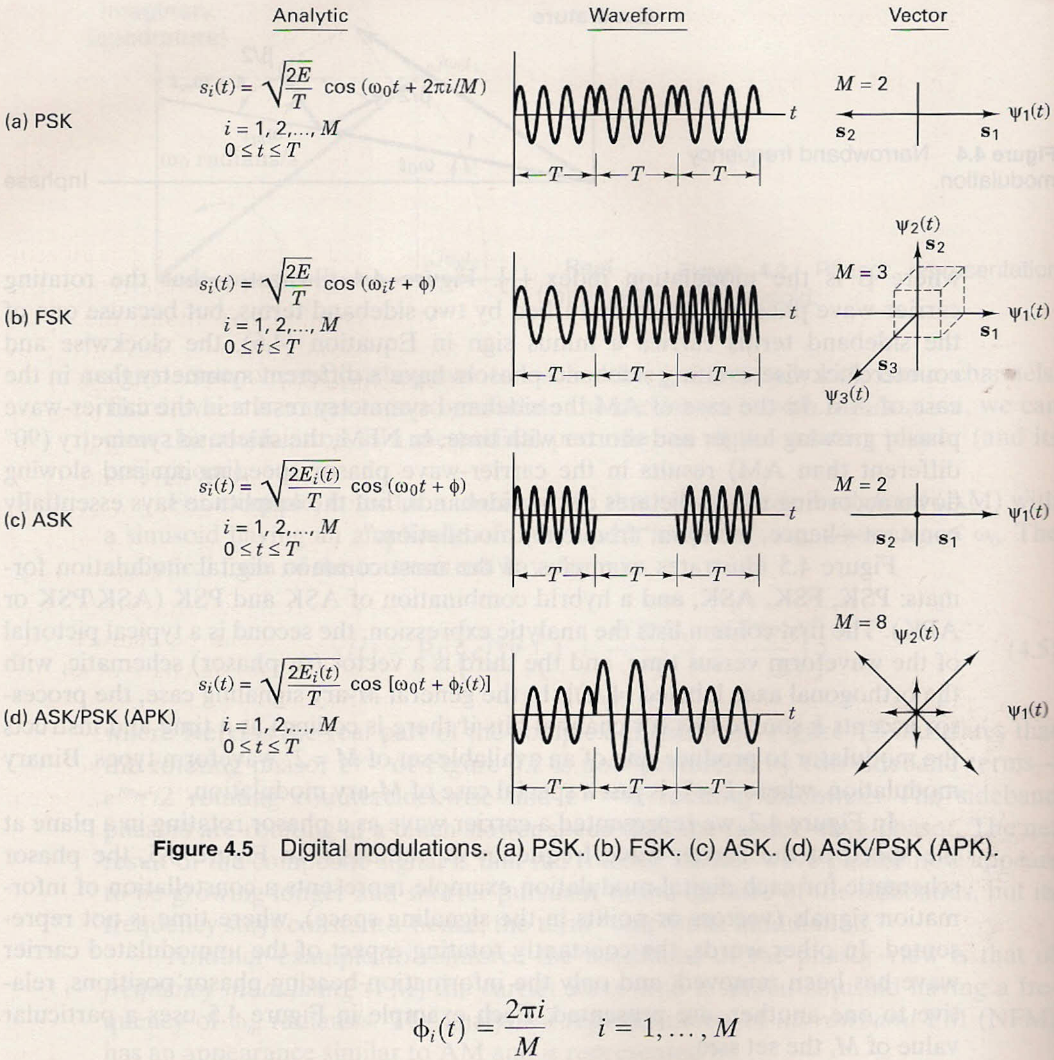


Figure 4.5 Digital modulations. (a) PSK. (b) FSK. (c) ASK. (d) ASK/PSK (APK).

$$\phi_i(t) = \frac{2\pi i}{M} \qquad i = 1, \ldots, M$$

For the binary PSK (BPSK) example in Figure 4.5a, M is 2. The parameter E is symbol energy, T is symbol time duration, and $0 \le t \le T$. In BPSK modulation, the modulating data signal shifts the phase of the waveform $s_i(t)$ to one of two states, either zero or π (180°). The waveform sketch in Figure 4.5a shows a typical BPSK waveform with its abrupt phase changes at the symbol transitions; if the modulating date stream were to consist of alternating ones and zeros, there would be such an abrupt change at each transition. The signal waveforms can be represented as vectors or phasors on a polar plot; the vector length corresponds to the signal amplitude, and the vector direction for the general M-ary case corresponds to the signal phase relative to the other $M - 1$ signals in the set. For the BPSK example, the vector picture illustrates the two 180° opposing vectors. Signal sets that can be depicted with such opposing vectors are called *antipodal signal sets.*

4.2.6 Waveform Amplitude Coefficient

The waveform amplitude coefficient appearing in Equations (4.7) to (4.10) has the same general form $\sqrt{2E/T}$ for all modulation formats. The derivation of this expression begins with

$$s(t) = A \cos \omega t \tag{4.11}$$

where A is the peak value of the waveform. Since the peak value of a sinusoidal waveform equals $\sqrt{2}$ times the root-mean-square (rms) value, we can write

$$\begin{aligned} s(t) &= \sqrt{2} A_{\text{rms}} \cos \omega t \\ &= \sqrt{2 A_{\text{rms}}^2} \cos \omega t \end{aligned}$$

Assuming the signal to be a voltage or a current waveform, A_{rms}^2 represents average power P (normalized to 1 Ω). Therefore, we can write

$$s(t) = \sqrt{2P} \cos \omega t \tag{4.12}$$

Replacing P watts by E joules/T seconds, we get

$$s(t) = \sqrt{\frac{2E}{T}} \cos \omega t \tag{4.13}$$

We shall use either the amplitude notation A in Equation (4.11) or the designation $\sqrt{2E/T}$ in Equation (4.13). Since the *energy* of a received signal is the key parameter in determining the error performance of the detection process, it is often more convenient to use the amplitude notation in Equation (4.13) because it facilitates solving directly for the probability of error P_E as a function of signal energy.

4.5 NONCOHERENT DETECTION

4.5.1 Detection of Differential PSK

The name *differential* PSK (DPSK) sometimes needs clarification because two separate aspects of the modulation/demodulation format are being referred to: the encoding procedure and the detection procedure. The term *differential encoding* refers to the procedure of encoding the data differentially; that is, the presence of a binary one or zero is manifested by the symbol's similarity or difference when compared with the preceding symbol. The term *differentially coherent detection* of differentially encoded PSK, the usual meaning of DPSK, refers to a detection scheme often classified as noncoherent because it does not require a reference in phase with the received carrier. Occasionally, differentially encoded PSK is *coherently* detected. This will be discussed in Section 4.7.2.

With noncoherent systems, no attempt is made to determine the actual value of the phase of the incoming signal. Therefore, if the transmitted waveform is

$$s_i(t) = \sqrt{\frac{2E}{T}} \cos\left[\omega_0 t + \theta_i(t)\right] \qquad \begin{array}{l} 0 \le t \le T \\ i = 1, \ldots, M \end{array}$$

the received signal can be characterized by

$$r(t) = \sqrt{\frac{2E}{T}} \cos\left[\omega_0 t + \theta_i(t) + \alpha\right] + n(t) \qquad \begin{array}{l} 0 \le t \le T \\ i = 1, \ldots, M \end{array} \tag{4.41}$$

where α is an arbitrary constant and is typically assumed to be a random variable uniformly distributed between zero and 2π, and $n(t)$ is an AWGN process.

For coherent detection, matched filters (or their equivalents) are used; for noncoherent detection, this is not possible because the matched filter output is a function of the unknown angle α. However, if we assume that α varies slowly relative to two period times $(2T)$, the phase difference between two successive waveforms $\theta_j(T_1)$ and $\theta_k(T_2)$ is independent of α; that is,

$$\left[\theta_k(T_2) + \alpha\right] - \left[\theta_j(T_1) + \alpha\right] = \theta_k(T_2) - \theta_j(T_1) = \phi_i(T_2) \tag{4.42}$$

The basis for *differentially coherent detection* of differentially encoded PSK (DPSK) is as follows. The carrier phase of the previous signaling interval can be used as a phase reference for demodulation. Its use requires *differential encoding* of the message sequence at the transmitter since the information is carried by the difference in phase between two successive waveforms. To send the ith message $(i = 1, 2, \ldots, M)$, the present signal waveform must have its phase advanced by $\phi_i = 2\pi i/M$ radians over the previous waveform. The detector, in general, calculates the coordinates of the incoming signal by correlating it with locally generated waveforms, such as $\sqrt{2/T}\cos\omega_0 t$ and $\sqrt{2/T}\sin\omega_0 t$. the detector then measures the angle between the currently received signal vector and the previously received signal vector, as illustrated in Figure 4.16.

In general, DPSK signaling performs less efficiently than PSK, because the errors in DPSK tend to propagate (to adjacent symbol times) due to the correlation between signaling waveforms. One way of viewing the difference between PSK and

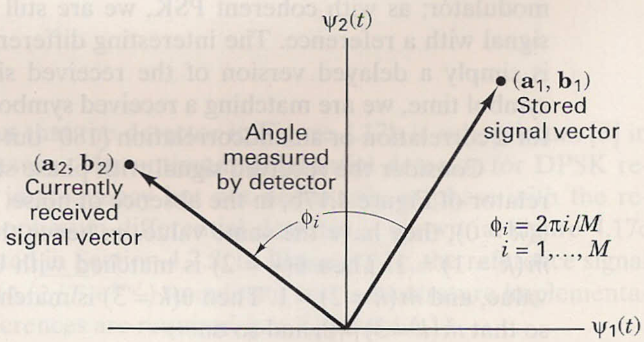


Figure 4.16 Signal space for DPSK.

DPSK is that the former compares the received signal with a clean reference; in the latter, however, two noisy signals are compared with each other. We might say that there is twice as much noise associated with DPSK signaling compared to PSK signaling. Consequently, as a first guess, we might estimate that the error probability for DPSK is approximately two times (3 dB) worse than PSK; this degradation decreases rapidly with increasing signal-to-noise ratio. The trade-off for this performance loss is reduced system complexity. The error performance for the detection of DPSK is treated in Section 4.7.5.

4.5.2 Binary Differential PSK Example

The essence of differentially coherent detection in DPSK is that the identity of the data is inferred from the changes in phase from symbol to symbol. Therefore, because the data are detected by differentially examining the waveform, the transmitted waveform must first be encoded in a differential fashion. Figure 4.17a illustrates a differential encoding of a binary message data stream $m(k)$, where k is the sample time index. The differential encoding starts (third row in the figure) with the first bit of the code-bit sequence $c(k = 0)$, chosen arbitrarily (here taken to be a one). Then the sequence of encoded bits $c(k)$ can, in general, be encoded in one of two ways:

$$c(k) = c(k-1) \oplus m(k) \tag{4.43}$$

or

$$c(k) = \overline{c(k-1) \oplus m(k)} \tag{4.44}$$

where the symbol $\oplus$ represents modulo-2 addition (defined in Section 2.9.3) and the overbar denotes complement. In Figure 4.17a the differentially encoded message was obtained by using Equation (4.44). In other words, the present code bit $c(k)$ is a one if the message bit $m(k)$ and the prior coded bit $c(k-1)$ are the same, otherwise, $c(k)$ is a zero. The fourth row translates the coded bit sequence $c(k)$ into the phase shift sequence $\theta(k)$, where a one is characterized by a 180° phase shift, and a zero is characterized by a 0° phase shift.

Figure 4.17b illustrates the binary DPSK detection scheme in block diagram form. Notice that the basic product integrator of Figure 4.7 is the essence of the demodulator; as with coherent PSK, we are still attempting to correlate a received signal with a reference. The interesting difference here is that the reference signal is simply a delayed version of the received signal. In other words, during each symbol time, we are matching a received symbol with the prior symbol and looking for a correlation or an anticorrelation (180° out of phase).

Consider the received signal with phase shift sequence $\theta(k)$ entering the correlator of Figure 4.17b, in the absence of noise. The phase $\theta(k=1)$ is matched with $\theta(k=0)$; they have the same value, π; hence the first bit of the detected output is $\hat{m}(k=1) = 1$. Then $\theta(k=2)$ is matched with $\theta(k=1)$; again they have the same value, and $\hat{m}(k=2) = 1$. Then $\theta(k=3)$ is matched with $\theta(k=2)$; they are different, so that $\hat{m}(k=3) = 0$, and so on.

Sample index, k	0	1	2	3	4	5	6	7	8	9	10
Information message, $m(k)$		1	1	0	1	0	1	1	0	0	1
Differentially encoded message (first bit arbitrary), $c(k)$	1	1	1	0	0	1	1	1	0	1	1
Corresponding phase shift, $\theta(k)$	π	π	π	0	0	π	π	π	0	π	π

(a)

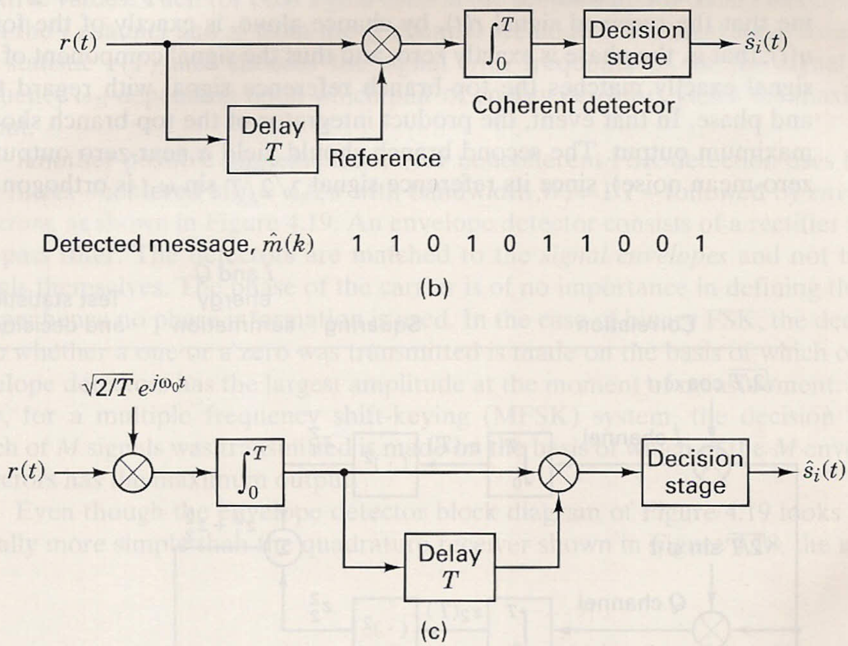


Figure 4.17 Differential PSK (DPSK). (a) Differential encoding. (b) Differentially coherent detection. (c) Optimum differentially coherent detection.

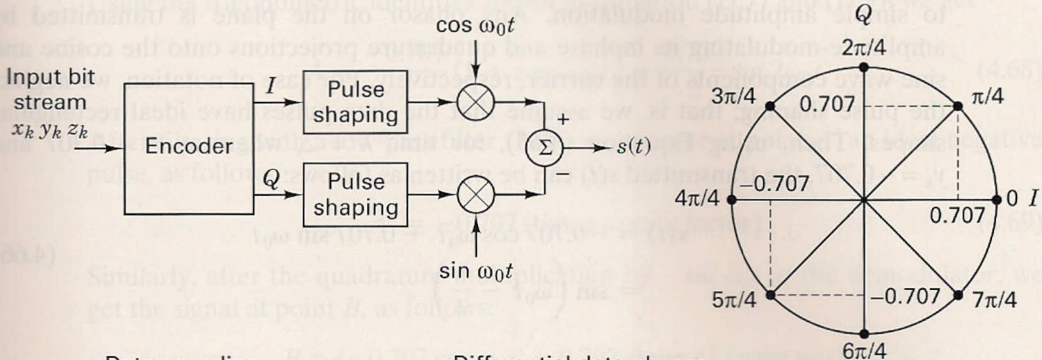
It must be pointed out that the detector in Figure 4.17b is suboptimum [3] in the sense of error performance. The optimum differential detector for DPSK requires a reference carrier in frequency but not necessarily in phase with the received carrier. *Hence the optimum differential* detector is shown in Figure 4.17c [4]. Its performance is treated in Section 4.7.5. In Figure 4.17c, the reference signal is shown in complex form ($\sqrt{2/T}\; e^{j\omega_0 t}$) to indicate that a quadrature implementation using both I and Q references are required (see Section 4.6.1).

4.6.2 D8PSK Modulator Example

Figure 4.23 depicts a quadrature implementation of a differential 8-PSK (D8PSK) modulator. Because the modulation is 8-ary, we assign a 3-bit message (x_k, y_k, z_k) to each phase $\Delta\phi_k$. Because the modulation is differential, at each kth transmission time we send a data phasor ϕ_k, which can be expressed as

$$\phi_k = \Delta\phi_k + \phi_{k-1} \tag{4.64}$$

The process of adding the current message-to-phase assignment $\Delta\phi_k$ to the prior data phase ϕ_{k-1} provides for the differential encoding of the message. A sequence of phasors created by following Equation (4.64) yields a similar differential encoding as the procedures described in Section 4.5.2. You might notice in Figure 4.23 that the assignment of 3-bit message sequences to $\Delta\phi_k$ does not proceed along the natural binary progression from 000 to 111. There is a special code being used here called a *Gray code.* (The benefits that such a binary to M-ary assignment provides is explained in Section 4.9.4.)



Data encoding

x_k	y_k	z_k	$\Delta\phi_k$
0	0	0	0
0	0	1	$\pi/4$
0	1	1	$2\pi/4$
0	1	0	$3\pi/4$
1	1	0	$4\pi/4$
1	1	1	$5\pi/4$
1	0	1	$6\pi/4$
1	0	0	$7\pi/4$

Differential data phasor

$\phi_k = \phi_{k-1} + \Delta\phi_k$

Let $\phi_0 = 0$	$k = 1$	$k = 2$	$k = 3$	$k = 4$
$x_k\, y_k\, z_k$	110	001	110	010
$\Delta\phi_k$:	π	$\pi/4$	π	$3\pi/4$
ϕ_k:	π	$5\pi/4$	$\pi/4$	π
I:	-1	-0.707	0.707	-1
Q:	0	-0.707	0.707	0

Figure 4.23 Quadrature implementation of a D8PSK modulator.

For the modulator in Figure 4.23, let the input data sequence at times $k = 1, 2, 3, 4$, be equal to 110, 001, 110, 010, respectively. Next, we use the date-encoding table shown in Figure 4.23 and Equation (4.64), with the starting phase at time $k = 0$ to be $\phi_0 = 0$. At time $k = 1$, the differential data phase corresponding to $x_1\, y_1\, z_1 = 110$ is $\phi_1 = 4\pi/4 = \pi$. Taking the magnitude of the rotating phasor to be unity, the inphase (I) and the quadrature (Q) baseband pulses are -1 and 0, respectively. As indicated in Figure 4.23, these pulses are generally shaped with a filter (such as a root-raised-cosine).

For time $k = 2$, the table in Figure 4.23 shows that the message 001 is assigned $\Delta\phi_2 = \pi/4$. Therefore, following Equation (4.64), the second differential data phase is $\phi_2 = \pi + \pi/4 = 5\pi/4$, and for time $k = 2$, the I and Q baseband pulses are $x_k = -0.707$ and $y_k = -0.707$, respectively. The transmitted waveform follows the form of Equation (4.61), rewritten as

$$\begin{aligned} s(t) &= \mathrm{Re}\{(x_k + j y_k)(\cos \omega_0 t + j \sin \omega_0 t)\} \\ &= x_k \cos \omega_0 t - y_k \sin \omega_0 t \end{aligned} \tag{4.65}$$

For a signaling set that can be represented on a phase-amplitude plane, such as MPSK or MQAM, Equation (4.65) provides an interesting observation. That is, quadrature implementation of the transmitter transforms all such signaling types to simple amplitude modulation. Any phasor on the plane is transmitted by amplitude-modulating its inphase and quadrature projections onto the cosine and sine wave components of the carrier, respectively. For ease of notation, we neglect the pulse shaping; that is, we assume that the data pulses have ideal rectangular shapes. Then, using Equation (4.65), for time $k = 2$, where $x_k = -0.707$ and $y_k = -0.707$, the transmitted $s(t)$ can be written as follows:

$$\begin{aligned} s(t) &= -0.707 \cos \omega_0 t + 0.707 \sin \omega_0 t \\ &= \sin\left(\omega_0 t - \frac{\pi}{4}\right) \end{aligned} \tag{4.66}$$

4.6.3 D8PSK Demodulator Example

In the previous section, the quadrature implementation of a modulator began with multiplying the complex envelope (baseband message) by $e^{j\omega_0 t}$, and transmitting the real part of the product $s(t)$, as described in Equation (4.63). Using a similar quadrature implementation, demodulation consists of reversing the process—that is, multiplying the received bandpass waveform by $e^{-j\omega_0 t}$ in order to recover the baseband waveform. The left side of Figure 4.24 shows the modulator of Figure 4.23 in simplified form, and the waveform $s(t) = \sin(\omega_0 t - \pi/4)$ that was launched at time $k = 2$ for that example. We continue this same example in this section, and also show a quadrature implementation of the demodulator on the right side of Figure 4.24.

Notice the subtle difference between the $-\sin \omega_0 t$ term at the modulator and the $-\sin \omega_0 t$ term at the demodulator. At the modulator, the minus sign stems from taking the real part of the complex waveform (product of the complex envelope and complex carrier wave). At the demodulator, $-\sin \omega_0 t$ stems from multiplying the bandpass waveform by the conjugate $e^{-j\omega_0 t}$ of the modulator carrier wave; demodulation is coherent if phase is recovered. To simplify writing the basic relationships of the process, the noise is neglected. After the inphase multiplication by $\cos \omega_0 t$ in the demodulator, we get the signal at point A:

$$\begin{aligned} A &= (-0.707 \cos \omega_0 t + 0.707 \sin \omega_0 t) \cos \omega_0 t \\ &= -0.707 \cos^2 \omega_0 t + 0.707 \sin \omega_0 t \cos \omega_0 t \end{aligned} \tag{4.67}$$

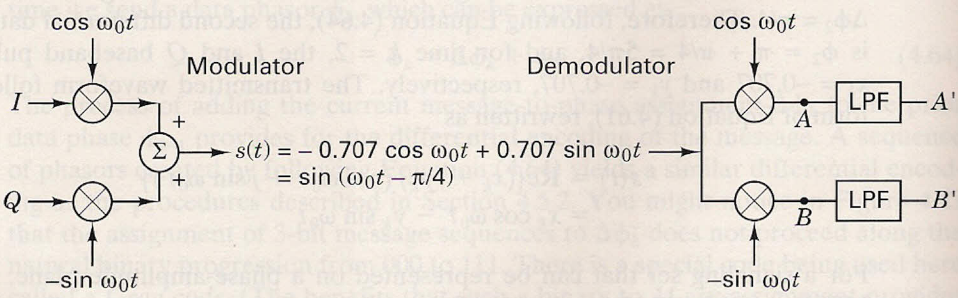


Figure 4.24 Modulator/demodulator example.

Using the trigonometric identities shown in Equation (D.7) and (D.9), we get

$$A = \frac{-0.707}{2}(1 + \cos 2\omega_0 t) + \frac{0.707}{2} \sin 2\omega_0 t \tag{4.68}$$

After filtering with a low-pass filter (LPF) we recover at point A' an ideal negative pulse, as follows:

$$A' = -0.707 \text{ (times a scale factor)} \tag{4.69}$$

Similarly, after the quadrature multiplication by $-\sin \omega_0 t$ in the demodulator, we get the signal at point B, as follows:

$$\begin{aligned} B &= (-0.707 \cos \omega_0 t + 0.707 \sin \omega_0 t)(-\sin \omega_0 t) \\ &= \frac{0.707}{2} \sin 2\omega_0 t - \frac{0.707}{2}(1 - \cos 2\omega_0 t) \end{aligned} \tag{4.70}$$

After the LPF we recover at point B' an ideal negative pulse, as follows:

$$B' = -0.707 \text{ (times a scale factor)} \tag{4.71}$$

Thus, we see from the demodulated and filtered points A' and B' that the differential (ideal) data pulses for the I and Q channels are each equal to -0.707. Since the modulator/demodulator is differential, then for this $k = 2$ example,

$$\Delta\phi_{k=2} = \phi_{k=2} - \phi_{k=1} \tag{4.72}$$

We presume that the demodulator at the earlier time $k = 1$ had properly recovered the signal phase to be π. Then, from Equation (4.72), we can write

$$\Delta\phi_{k=2} = \frac{5\pi}{4} - \pi = \frac{\pi}{4} \tag{4.73}$$

Referring back to the data-encoding table in Figure 4.23, we see that the detected data sequence is $x_2\, y_2\, z_2 = 001$, which corresponds to the data that was sent at time $k = 2$.

4.9 SYMBOL ERROR PERFORMANCE FOR *M*-ARY SYSTEMS ($M > 2$)

4.9.1 Probability of Symbol Error for MPSK

For large energy-to-noise ratios, the symbol error performance $P_E(M)$, for equally likely, coherently detected M-ary PSK signaling, can be expressed [7] as

$$P_E(M) \approx 2Q\left(\sqrt{\frac{2E_s}{N_0}}\sin\frac{\pi}{M}\right) \tag{4.105}$$

where $P_E(M)$ is the probability of symbol error, $E_s = E_b(\log_2 M)$ is the energy per symbol, and $M = 2^k$ is the size of the symbol set. The $P_E(M)$ performance curves for coherently detected MPSK signaling are plotted versus E_b/N_0 in Figure 4.35.

The symbol error performance for differentially coherent detection of M-ary DPSK (for large E_s/N_0) is similarly expressed [7] as

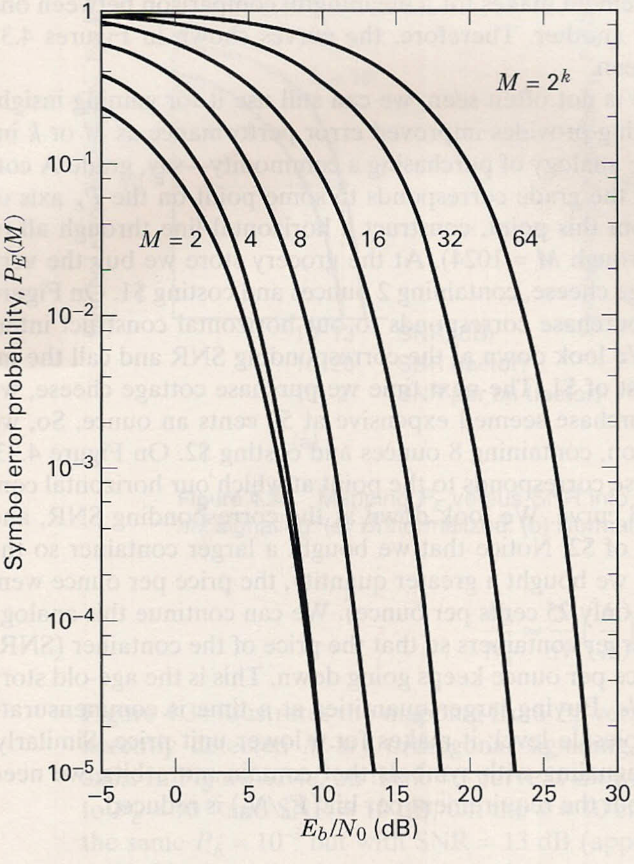


Figure 4.35 Symbol error probability for coherently detected multiple phase signaling. (Reprinted from W. C. Lindsey and M. K. Simon, *Telecommunication Systems Engineering,* Prentice-Hall, Inc. Englewood Cliffs, N.J., 1973, courtesy of W. C. Lindsey and Marvin K. Simon.)

$$P_E(M) \approx 2Q\left(\sqrt{\frac{2E_s}{N_0}} \sin \frac{\pi}{\sqrt{2}M}\right) \tag{4.106}$$

4.9.4 Bit Error Probability Versus Symbol Error Probability for Multiple Phase Signaling

For the case of MPSK signaling, P_B is less than or equal to P_E, just as in the case of MFSK signaling. However, there is an important difference. For orthogonal signaling, selecting any one of the $(M-1)$ erroneous symbols is equally likely. In the case of MPSK signaling, each signal vector is not equidistant from all of the others. Figure 4.39a illustrates an 8-ary decision space with the pie-shaped decision regions denoted by the 8-ary symbols in binary notation. If symbol (0 1 1) is transmitted, it is clear that should an error occur, the transmitted signal will most likely be mistaken for one of its closest neighbors, (0 1 0) or (1 0 0). The likelihood that (0 1 1) would get mistaken for (1 1 1) is relatively remote. If the assignment of bits to symbols follows the binary sequence shown in the symbol decision regions of Figure 4.39a, some symbol errors will usually result in two or more bit errors, even with a large signal-to-noise ratio.

For nonorthogonal schemes, such as MPSK signaling, one often uses a binary-to-M-ary code such that binary sequences corresponding to adjacent sym-

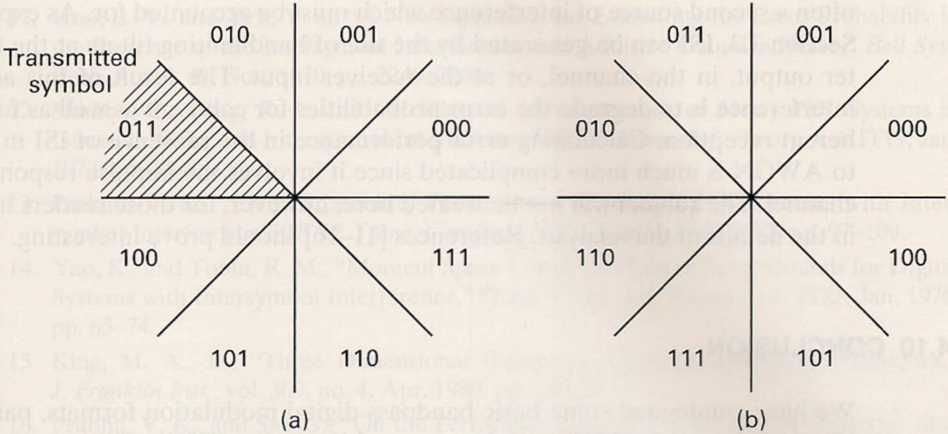


Figure 4.39 Binary-coded versus Gray-coded decision regions in an MPSK signal space. (a) Binary coded. (b) Gray coded.

bols (phase shifts) differ in only one bit position; thus when an M-ary symbol error occurs, it is more likely that only one of the k input bits will be in error. A code that provides this desirable feature is the Gray code [7]; Figure 4.39b illustrates the bit-to-symbol assignment using a Gray code for 8-ary PSK. Here it can be seen that neighboring symbols differ from one another in only one bit position. Therefore, the occurrence of a multibit error, for a given symbol error, is much reduced compared to the uncoded binary assignment seen in Figure 4.39a. Implementing such a Gray code, represents one of the few cases in digital communications where a benefit can be achieved without incurring any cost. The Gray code is simply an assignment that requires no special or additional circuitry. Utilizing the Gray code assignment, it can be shown [5] that

$$P_B \approx \frac{P_E}{\log_2 M} \quad (\text{for } P_E \ll 1) \tag{4.113}$$

Recall from Section 4.8.4 that BPSK and QPSK signaling have the same bit error probability. Here, in Equation (4.113), we verify that they do not have the same symbol error probability. For BPSK, $P_E = P_B$. However, for QPSK, $P_E \approx 2P_B$.

An exact closed-form expression for the bit-error probability P_B of 8-ary PSK, together with tight upper and lower bounds on P_B for M-ary PSK with larger M, may be found in Lee [10].