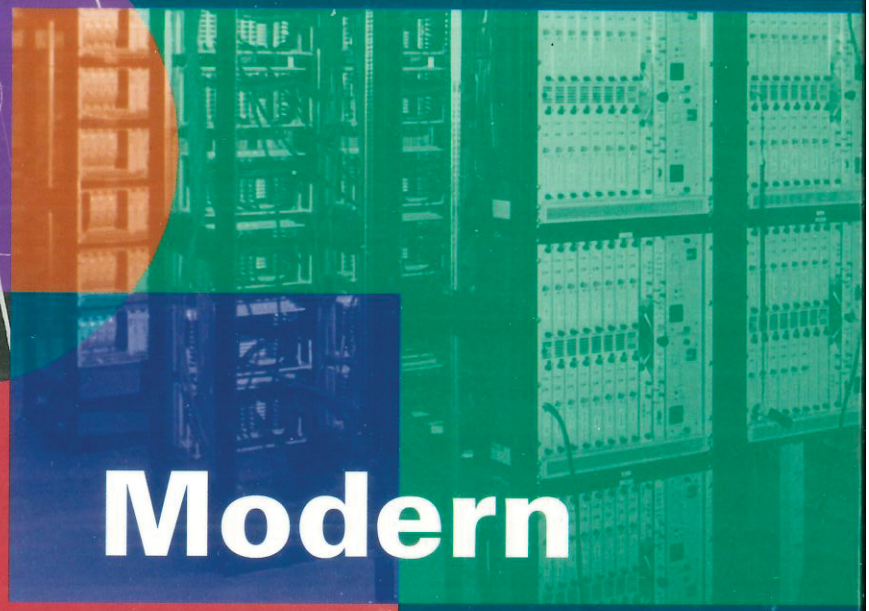
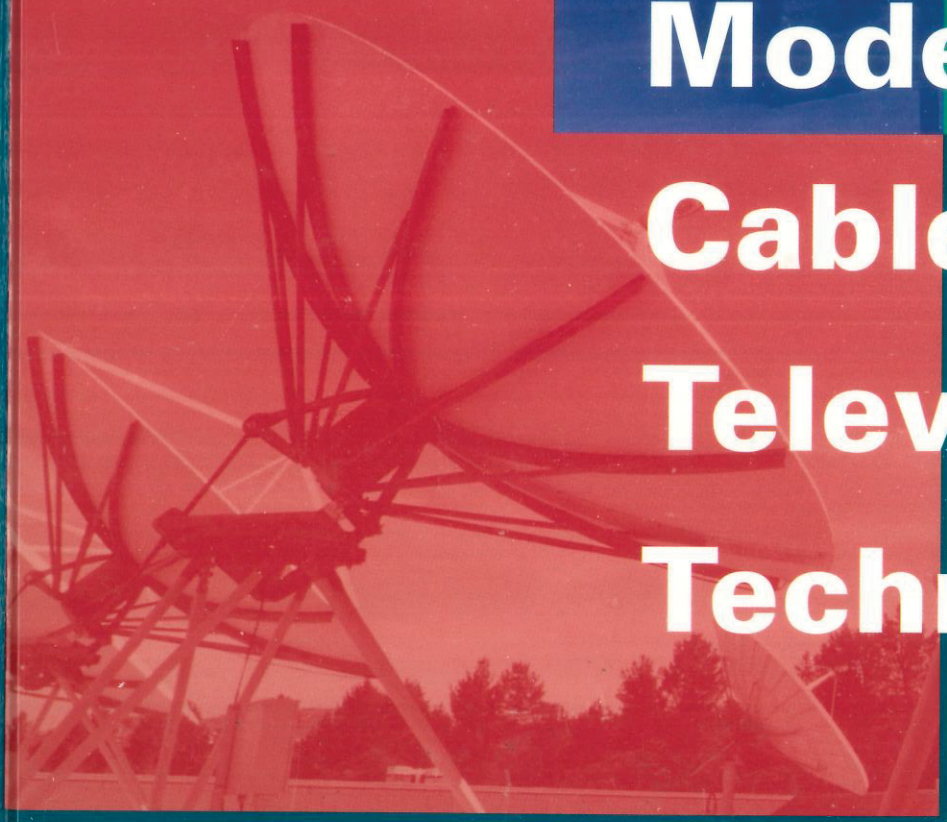
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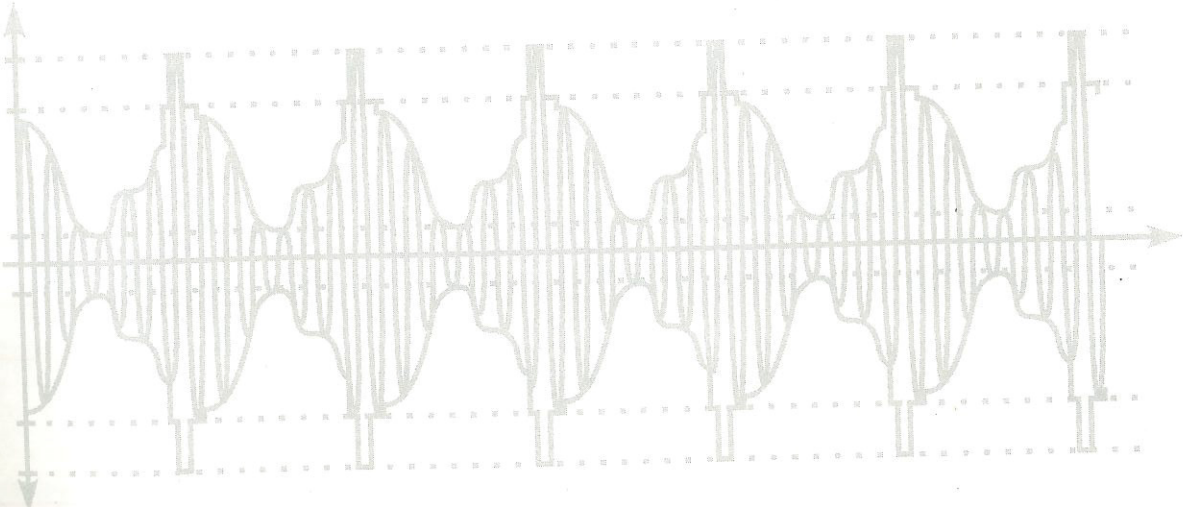
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4.1 Introduction

Perhaps the most discussed new service in cable television today is data transport. The primary emphases are cable telephony and Internet access services via cable, as well as digital video transmission. Factors making cable an attractive medium include economy, speed, and connectionless service.

This chapter does not purport to be a treatise in data communications. Rather, it offers concepts that are particularly germane to the practice of data transmission on HFC networks. It is intended primarily to be an introduction to data communications for cable practitioners who have not previously been exposed to the subject. The examples of modulation methods and communications protocols are restricted to those that appear to be most germane in cable data communications for the next few years.

4.2 Modulation Methods

The object of data transmission is to get a stream of bits from here to there. *Bit* stands for *bi*nary digi*t*, and is a simple voltage that can take on only one of two levels. We arbitrarily call these levels 1 and 0. Often, but not always, they are represented in electronic circuits by the voltages +5 and 0 volts (newer systems use 3.3 volts or so). A combination of many of these bits make up *digital data*.

It is possible to transmit bits of data over an RF carrier using the same variables as are used in analog transmission. One can vary, or "modulate" the amplitude, frequency, or phase of the carrier, with the information to be transmitted. Frequency modulation is often used for very simple data communications links, due to the ability to construct low-cost FM receivers capable of very good performance. However, frequency modulation does not provide good "spectral efficiency."

4.2.1 Spectral Efficiency

One figure of merit used to compare different modulation formats is *spectral efficiency*, expressed in bits per hertz: the number of bits per second of data, divided by the RF bandwidth required to transmit it. A modulation type with high spectral efficiency can transmit more bits per second of data in the same number of hertz of RF bandwidth. On the other hand, it can be expected to cost more and be less robust (more susceptible to noise and transmission distortion). For example, we will show that the theoretical spectral efficiency of BPSK modulation is one bit per hertz. That is, when using BPSK modulation, the theoretical bandwidth required to transmit the signal is equal to the bit rate: a 1-Mb/s (1 megabit per second) data signal will require 1 MHz of RF bandwidth. On the other hand, if the same signal were transmitted using 256-QAM, the theoretical bandwidth required is only 0.125 MHz because the spectral efficiency is eight bits per hertz (1/8 = 0.125). However, all else being equal, 256-QAM requires a carrier-to-noise

ratio (C/N) more than 13 dB better, and the cost of the hardware is considerably higher. In all cases, the actual occupied bandwidth is wider than the theoretical because of practical considerations.

4.2.2 Frequency Shift Keying (FSK)

The simplest commonly used modulation method is frequency modulation, called *frequency shift keying* (FSK). Information is transmitted by shifting an oscillator between two frequencies. The receiver consists of a conventional frequency demodulator. It is followed by a shaping section, which restores the pulses to their original shape.

FSK systems tend to be popular in cable applications for relatively low-speed data transmission where receiver cost is paramount. Applications include downstream data transmission to addressable set top terminals, upstream transmission from RF impulse-pay-per-view (IPPV) converters, and status monitoring. Spectral efficiency is very poor, but this is relatively unimportant in low-speed applications. The transmission is robust (not easily damaged by noise or distortion), and required hardware is quite low in cost. Single-chip data receivers have been available for many years. Because of its popularity, the technique is described in some detail.

Figure 4.1 illustrates a simple FSK communications system. Data, in the form of a series of 1s and 0s, is applied to a low-pass filter whose function is to minimize the number of sidebands produced by the modulation process. The filtered waveform is applied to a frequency modulated oscillator, whose frequency deviation is proportional to the voltage applied at the modulating input. A parameter of an FSK system is the peak-to-peak deviation, or the difference between the two frequencies of the oscillator, when 1 and 0 are applied.* The wider the deviation, the more immune the system is to noise, but the more spectrum the signal occupies.

After transmission to the receiving end, the signal is usually passed through a bandpass filter (BPF), then applied to a discriminator, which is very similar to one used for analog signal recovery. Low-pass filtering (LPF) is usually used after the discriminator to reduce noise picked up during transmission. Capacitive coupling is shown after the filter to make sure the signal can be "sliced" at its average value. From the capacitor, the signal is supplied to a data slicer, a comparator whose output is 1 when the input voltage is above ground and 0 when it is below ground (or vice versa). The slicing process restores the wave shape of the original signal.

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* In the analog world, deviation is normally measured on a peak basis—the difference between either extreme frequency and the center frequency. Peak-to-peak deviation is twice the peak deviation. In digital transmission, the carrier never rests at the center frequency, so peak deviation tends to have less meaning. We must be precise, however, in distinguishing between peak and peak-to-peak deviation.

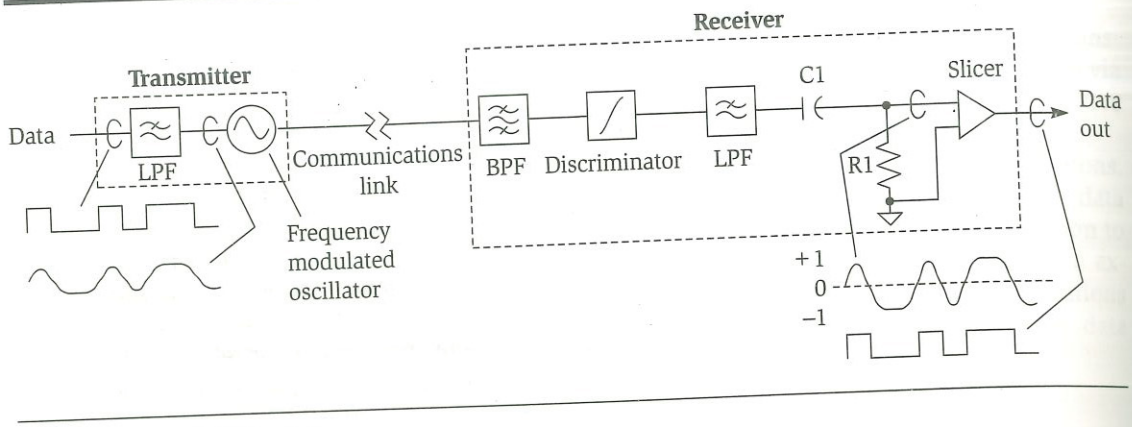


Figure 4.1 A simple FSK system.

Most transmission systems require that the dc value of the data waveform be removed for transmission. This means that, in some manner, the waveform averaged over some reasonable interval must have a value of 0 volts. This might be violated if the datastream contains a long string of 1s or 0s. If the signal doesn't have the same quantity of 1 and 0 states when averaged over some short time interval, C1 will take on a charge, and the slicing will not occur in the center of the waveform. If the signal has more 1s than 0s or vice versa, it is said to have a dc component, which is undesirable. Later we show several ways to ensure that a signal doesn't have a dc component.

Filtering in FSK Transmission Systems

At least three stages of filtering are used in the transmission system. A low-pass filter at the transmitter limits the spectrum of the baseband signal (before modulation). In many other modulation methods, the filter may be implemented either at the baseband or at RF, after modulation. However, in FSK the nonlinear nature of frequency modulation precludes doing so. The receiver contains a bandpass filter at RF (or IF), which is used to limit the energy applied to the discriminator. After demodulation, a baseband low-pass filter is used to remove as much noise as possible.

In low-cost FSK systems, the filtering philosophy often recognizes that either the transmitter or receiver is in the subscriber's home, and this end must be built at minimum cost. Placing most of the filtering at the headend helps keep the cost of the home equipment down. This somewhat compromises the noise performance and spectral efficiency in favor of low cost. For example, a data communications system that communicates from the headend to addressable set top terminals will employ significant filtering at the transmitter. At the receiver, the IF filter frequently employs a low-cost ceramic filter made for FM receiver use

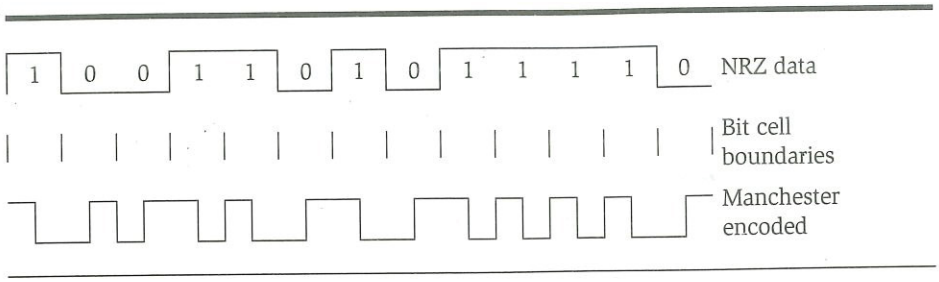


Figure 4.2 Manchester encoded data.

(180–200 kHz bandwidth), and the deviation is set to make the occupied spectrum match the width of the filter.

Manchester Encoding

We now describe one method for adding a clock to the data signal. It is independent of the modulation method but is introduced at this point because it is often used with FSK modulation. In data transmission, it is necessary not only to recover the transmitted data but also to recover a clock signal, which is used to define when a data state should be sampled. This is necessary in order to allow the receiver to "know" when a new bit arrives. Clock recovery, or making this determination, can be done in many ways, but Manchester encoding is often used in low-cost systems because of its simplicity. The data is said to be "self-clocking." The penalty is that the spectral efficiency of the transmission is halved compared with what it would be otherwise.

Figure 4.2 illustrates one form of Manchester encoding, which is applied to the data before it enters the modulator. At the top of the figure is a string of data represented in NRZ (nonreturn to zero) format, in which the data bits appear one after another, with nothing in between. A separate clock signal is typically used to transfer NRZ data from one place to another. However, in a communications link, a separate clock signal path does not exist. Therefore, in some manner, the clock must be embedded with the data. Manchester encoding provides a simple, though bandwidth inefficient, way to do this.

The center portion of Figure 4.2 shows the bit cell boundaries: the time when the signal transitions from one bit to the next. At the bottom of the figure is shown the Manchester encoded data. In the middle of each bit cell, Manchester encoding causes the signal to transition from one level to the other. This regularly spaced set of transitions forms a clock for the receiver. The first half of each cell contains the value of the bit, and the second half contains the complement (opposite state) of the bit. The opposite polarity is used in some systems. A variation is differential Manchester encoding: at the beginning of each cell containing a 0, the encoded signal makes another transition; at the beginning of each cell containing a 1, the transition is omitted.

Note that Manchester encoding is *not* a modulation method. Rather, it is a baseband coding method that is useful in low-cost systems to provide clock recovery and removal of a dc component of the data.

4.2.3 Baseband Spectrum of a Data Signal

The spectrum of a data signal may be understood with reference to the waveform of Figure 4.2 and the low-pass filtering of Figure 4.1. Figure 4.1 shows that it is possible (and usually desirable) to low-pass filter the data waveform so that when a string of alternating 1s and 0s (or a string of like bits in Manchester encoded data) is encountered, the waveform looks substantially like a sine wave. No useful data is lost in the filtering process, and the occupied bandwidth of the signal is thus minimized. Practical filters are somewhat wider than this, but the principle remains.

When a string of 1s and 0s is encountered in an NRZ signal and optimally filtered, the resulting waveform is substantially a sine wave at a frequency of one-half the bit rate. To understand this, recall that a complete cycle of a sine wave occupies the time from one positive going zero crossing to the next positive going zero crossing. In order to produce such a waveform, two bits, 1 followed by 0, are required.

In general, a data signal can contain alternating 1s and 0s. This pattern defines the highest baseband frequency, equal to one-half the bit rate, assuming perfect filtering. However, that data can also contain arbitrarily long strings of 1s or 0s, which represent a frequency approaching 0. Because of this, the spectrum of the baseband (unmodulated) data signal extends from one-half the bit rate, down to very close to 0. Some systems limit the number of consecutive 1s or 0s transmitted (another way to remove a dc component) because many clock recovery techniques cannot tolerate an infinitely long string with no transitions. Also, if the frequency gets too low, slicing thresholds can be thrown off.

In summary, the spectrum of a digital signal extends from very low to at least one-half the bit rate (really the symbol rate, a concept to be introduced later in the chapter). Practical systems have frequencies that extend above this due to real-world filtering constraints. The excess bandwidth may range from a few percent to much higher depending on the application.

4.2.4 Introduction to Bit Error Rate

An important figure of merit for any data communications system is the *bit error rate* (BER) of the system. This expresses the proportion of the bits transmitted, which are received incorrectly. If, on the average, two bits are in error for every million (10^6) bits transmitted, then the bit error rate is

$$\text{BER} = \frac{\text{number of bits received in error}}{\text{total number of bits transmitted}} = \frac{2}{10^6} = 2 \times 10^{-6}$$

The BER is a function of the type of modulation, the characteristics of the modulator and demodulator, and the quality of the transmission channel. A common way to compare different modulation systems is to plot BER versus signal-to-noise ratio (data communications practitioners tend to not differentiate between "signal" and "carrier" to noise). This is a convenient way to compare the capabilities of different modulation techniques and to predict performance in the presence of noise. This topic is covered in Section 4.2.13. Sometimes BER is plotted versus signal-to-noise ratio in the presence of certain other impairments, such as group delay distortion, interference, and phase noise.

Obtaining an equation relating carrier-to-noise ratio to BER for an FSK system is difficult because of the nonlinear nature of the frequency modulation process. The literature contains only a few attempts, and contradictions exist. Within the parameters commonly used in cable television work (very wide deviation compared with the bit rate), the BER can be quite good at very low C/N. The penalty is that very significant bandwidth is consumed, compared with that required for other modulation methods operating at the same bit rate.

4.2.5 ASK and PSK

Less commonly used, but sometimes encountered, forms of digital modulation include amplitude shift keying (ASK), in which the amplitude of the carrier is changed between two states to represent 1 and 0. Phase shift keying (PSK) varies the phase of the carrier between two states to represent 1 and 0. If the phase shift between the two states is 180°, the modulation is called BPSK.

4.2.6 Biphase Shift Keying (BPSK)

Biphase shift keying (BPSK) is the simplest of a class of modulation that is used extensively in cable television work. It is somewhat more complex to detect than is FSK though producing a BPSK signal is simple. In a BPSK system, the phase is shifted 180° between transmission of a 0 and a 1.

Figure 4.3 illustrates a simple BPSK transmission system. An NRZ datastream is low-pass filtered (LPF1) to remove higher-order harmonics. The filtered signal is passed to a balanced amplitude modulator (M1), whose RF input is supplied from a local oscillator.

The figure illustrates the output of the modulator, which consists of a carrier envelope that drops to zero amplitude in the transitions between 0 and 1. The modulated carrier has one phase relation for a 0 and the opposite phase when a 1 is being transmitted. Note the phase of the carrier with respect to the timing tick marks shown below the RF envelope. The reduction in amplitude to zero between states is a direct result of the filtering action of LPF1. Were it not for this filtering, the transition between 1 and 0 would not require the amplitude of the carrier to drop, but the occupied bandwidth would theoretically be infinite.

At the receiver, the signal is bandpass filtered in BPF1. From the filter, the signal is supplied to mixer M2, the demodulator. The second input to M2 is from

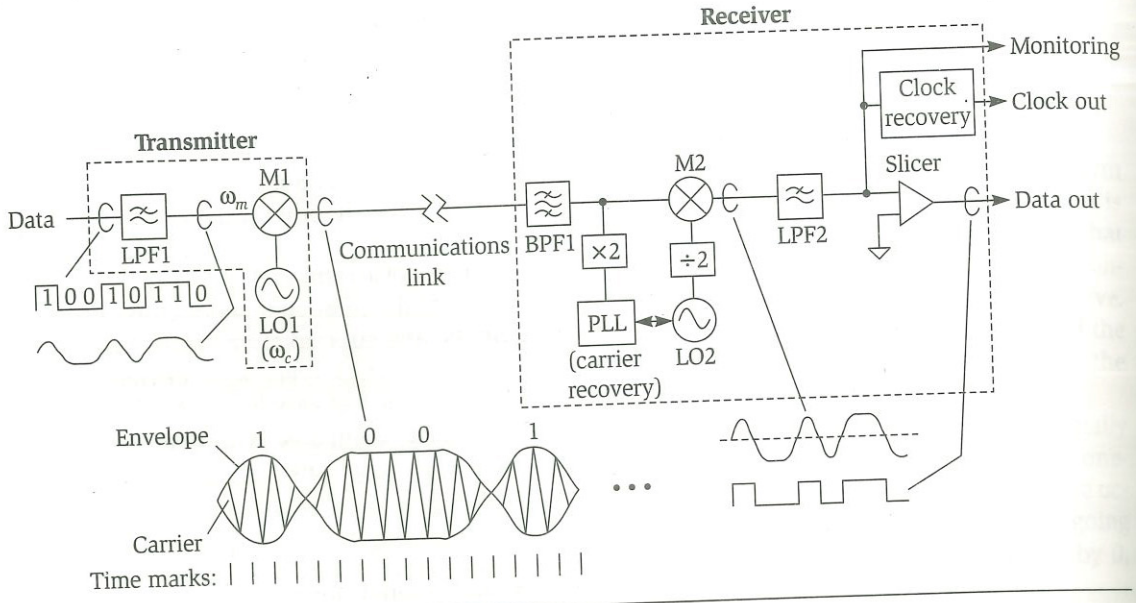


Figure 4.3 BPSK transmission system.

local oscillator LO2. This oscillator must be phase locked to the incoming RF carrier. A number of different carrier recovery circuits can be used. The circuit shown utilizes the fact that the second harmonic of the carrier contains no phase shift (when the frequency is doubled, a 180° phase shift becomes a 360° phase shift, or no phase shift at all). The incoming frequency is doubled by the times two multiplier, ×2, then used to lock LO2. The output of LO2 is divided by 2 and supplied as the second input to M2.

Basic Math of BPSK

You might be tempted to look at the waveform of Figure 4.3 and conclude that BPSK is a phase modulation technique because the carrier reverses phase between transmission of 1s and 0s. Indeed, the name of the modulation leads to that conclusion. The correct way to visualize BPSK is as double sideband suppressed carrier modulation. This type of modulation was described for analog signals in Chapter 2.

To understand the modulation process, assume that the data signal is an alternating series of 1s and 0s, and has been filtered to a sine wave, following the logic of earlier sections of this chapter. Then the modulating signal can be expressed as

$$e_m = M \sin \omega_m t \tag{4.1}$$

where

$_m$ = the modulating frequency

$\omega_m = 2\pi f_m$

As implied by Figure 4.3, the modulation process involves multiplying this modulating frequency by the carrier frequency, represented as

$$e_c = A \sin \omega_c t \tag{4.2}$$

The resulting modulated signal is expressed as follows, expanded using common trigonometric relations:

$$\begin{aligned} e_r &= e_c \cdot e_m \\ &= AM \sin \omega_c t \sin \omega_m t \\ &= \frac{AM}{2} \cos(\omega_c - \omega_m)t - \cos(\omega_c + \omega_m)t \end{aligned} \tag{4.3}$$

The center row shows the phase shift with modulation: when the sinusoid representing the data modulation (sin ω_mt) is positive, then the carrier has one phase. When it is negative, the carrier phase is reversed, as illustrated in Figure 4.3. Equation (2.9) expresses the same thing, with frequency expressed in hertz rather than radians ($\omega = 2\pi f$), and f_1 and f_2 used in place of ω_c and ω_m.

The bottom expression shows that the frequency components of the modulated signal exist at the carrier frequency ω_c minus the modulating frequency ω_m, and at the sum frequency. Thus, the bandwidth occupied by the signal is twice the modulating frequency, as is the case with conventional double sideband amplitude modulation. A component at the carrier frequency does not exist. The demodulation process follows the mathematics of synchronous detection in Chapter 2.

4.2.7 Some Fundamental Principles of Data Communications

Having introduced the concept of BPSK modulation, a fundamental digital modulation method, we use it to introduce a number of concepts that are useful in understanding many modulation formats. The principles presented in this section are applicable to many types of digital modulation.

Filtering Methodology

This section introduces the filtering methodology that is followed in many data communications applications and for many types of modulation (except FSK, which we covered earlier). The filters at the transmitter and receiver are effectively in cascade in the signal path, and together it is desirable that they form a filter that is just wide enough to convert the highest modulating frequency

into a sine wave. (As a practical matter, the filter will be somewhat wider than this.) A particularly useful class of filters is called *raised cosine* filters by the data communications industry and *sine squared* filters by the television engineering community.

The equivalence is seen in the trigonometric relationship

$$2\sin^2 \alpha = 1 - \cos 2\alpha \tag{4.4}$$

The left side of the equation shows the origin of the television usage of the term *sine squared,* whereas the right side shows the basis of the data communications term *raised cosine.* The 1 on the right "raises" the cosine term, which by itself, ranges from −1 to +1. In television work, a sine squared filter is used to produce the well-known sine squared pulse and bar, which are useful in analyzing video channel performance. In data communications, the cascaded performance of the low-pass filters at the transmitter and receiver, possibly in addition to the response of the IF filters, form a raised cosine filter, which ensures a fast transition halfway between the states, consistent with minimal spectrum occupancy.

It is desirable to heavily filter at the transmit end in order to reduce the spectrum usage, and heavy filtering is desirable at the receive end to minimize noise added during transmission. However, the total filtering applied must conform to the raised cosine shape. In many communications systems, the designer places one-half of the filter function at each end of the communication channel. When this is done, the filter response at each end is the square root of the complete filtering function, so the filter used on each end is often described as a *root raised cosine* filter. Note that the filters at the transmit and receive ends must be matched to each other and collectively are often referred to as *matched filters.*

Required Bandwidth of the Signal

The bandwidth requirements of BPSK can be understood by considering the worst case of an alternating string of 1s and 0s. This will produce a square wave, one cycle of which comprises an alternate 1 and 0, or two bits. Thus, the frequency of the square wave, f_m, is one-half the bit rate. Figure 4.4 illustrates the spectrum of the data signal at baseband. The square wave of alternating 1s and 0s illustrates the highest frequency of the data signal. A square wave consists of all odd harmonics of the fundamental frequency, as shown by the vertical lines in the spectrum plot of the figure. In real data systems, the data is not an alternating series of 1s and 0s. Rather, it is all possible combinations of strings of 1s and 0s. The resulting spectrum will fill in the spectrum from the maximum frequency (f_m) down to close to 0 frequency (and all harmonics thereof).

However, this total spectrum is not needed to convey the information contained in the datastream. If all frequencies above f_m are removed by filtering, then an alternating series of 1s and 0s will produce a sine wave of frequency f_m. Transitions that occurred at a slower rate (that is, from a series of like states) would follow a sinusoidal transition path, but would be flattened on top.

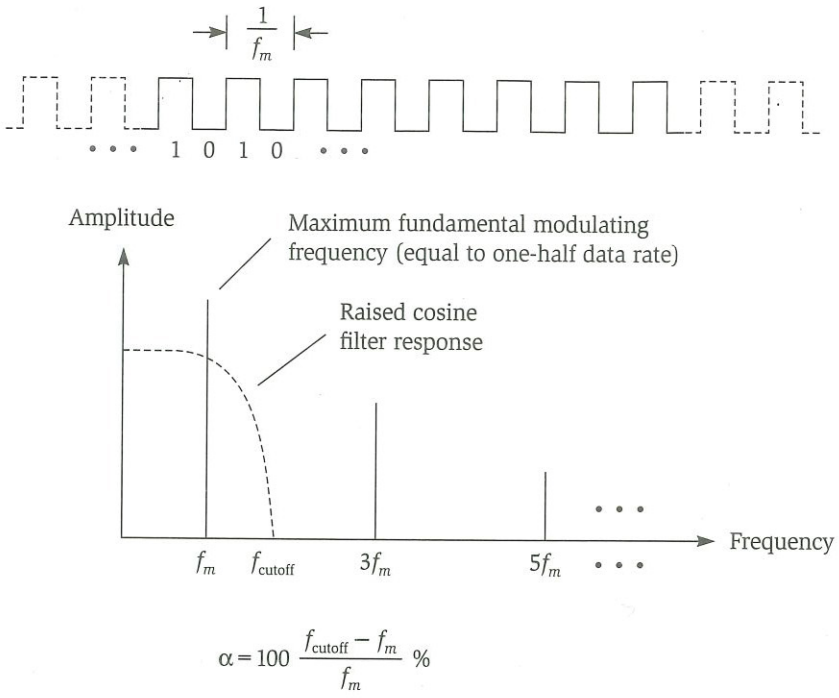


Figure 4.4 Baseband spectrum of BPSK signal.

The *theoretical* (not practical) spectral efficiency of a BPSK signal is developed as follows. The alternating 1–0 sequence defines the highest frequency of the data baseband signal. If it is filtered by removing all harmonics, leaving a sine wave, then the frequency of the sinusoid is one-half the data rate (one cycle comprises a set of 1 and the following 0). The bandwidth is twice the modulating frequency (the sum less the difference of the carrier and modulating frequency), which in turn is equal to one-half the bit rate. Thus, the spectral efficiency is two times one-half the data rate, or one times the bit rate, one bit per hertz.

A filter that completely cuts off just above f_m may be ideal, but is hardly practical. A practical filter has the raised cosine shape and is somewhat wider than f_m. The sharper the filter cutoff (that is, the less bandwidth required between the filter passband and the stopband), the less bandwidth is required. However, the sharper filter is more difficult to build, and more group delay is likely, depending on the realization. The filter width is described by the "alpha" of the filter, where alpha is defined as the excess of occupied bandwidth compared with the minimum required bandwidth:

$$\alpha = 100\frac{f_{\text{cutoff}} - f_m}{f_m}\% \tag{4.5}$$

Recall that the common filtering practice is to put the square root of the raised cosine filter at the transmitter, and the same at the receiver. For example, as shown later, the filter specification for the transmitter of a DOCSIS-compliant downstream modulator operating in the 64-QAM mode is an 18% root raised cosine filter, meaning that the filter is 18% wider than the minimum required bandwidth.[1] This may also be interpreted that the occupied bandwidth is 18% wider than the theoretical minimum.

Clock Recovery

Although self-clocking formats, such as Manchester encoding (see Section 4.2.2), could be used with BPSK, this is normally not done. The higher number of transitions associated with Manchester encoding would double the required bandwidth of the signal. Rather, an NRZ datastream is transmitted. Clock recovery consists of synchronizing a clock to the transitions in the received data. Where no transition occurs, the clock "freewheels" through the place where the transition should occur. If a long transmission of 0s or 1s occurs, it is possible to lose clock synchronization, so *scrambling* is often used to force a state change every so often.

Data Scrambling

As a method of dc removal, scrambling is often used. In order to effect proper operation of a number of circuits in the receiver, including clock recovery, dc removal is necessary. Scrambling used in the data transmission sense is not the same as scrambling used in the traditional cable television sense. In traditional cable television, the word *scrambling* refers to the process of rendering a video and/or audio signal unintelligible to a normal receiver. A special circuit must be used to restore the signal to its intelligible form.

In data transmission, scrambling is the process of combining a datastream with a pseudorandom bit sequence (PRBS), with the intent to remove any possible dc component of the datastream. An equivalent way of expressing the function of the scrambler is to say that it equalizes the power density over the occupied bandwidth. It can be shown that, if any data signal is exclusive ORed with a pseudorandom signal having no dc component, the resultant signal will not have a dc component. The signal can be recovered at the receiver by exclusive ORing the signal with the same pseudorandom signal.

The exclusive OR operation is a logical operation on two input bits of data. If the two bits are the same, the output of the exclusive OR operation is a logical 0. If the two bits are different, the output is a logical 1. One of the input bits is from the datastream and the other from the PRBS.

Differential Encoding

The phase locked loop used for carrier recovery in Figure 4.3 creates a problem, in that a phase error ambiguity exists: if LO2 divided by 2 is in phase with the car-

rier applied to M1, the recovered data will have one polarity. If it is out of phase, the recovered data will have the opposite polarity. Since the loop is closed at twice the frequency of the carrier, there does not exist a simple way to determine the correct phase. One way would be to detect a code word buried in the data and correct for a phase error.

A method frequently used is *differential encoding*. In differential BPSK, a 0 is represented by no change in phase from the previous state. A 1 is represented as a 180° phase change. The detector is arranged to compare the phase of the previous bit with the phase of the present bit. If the phase doesn't change, a 0 is delivered as the received bit. If the phase is changed, a 1 is delivered (or vice versa). Differential encoding is used with a number of different modulation methods.

Evaluation of the Quality of a Digitally Modulated Signal

The recovered waveform at the "monitoring" point shown in Figure 4.3 should be the same as the waveform applied to modulator M1. However, distortion in the transmission path will change the wave shape, as suggested in Figure 4.5 (the dotted line is the original wave shape, and the solid line is the received wave shape). Noise will put "ripple" on the waveform as seen on an oscilloscope. If enough noise is added, then the state of the signal may be misinterpreted, producing a bit error. In addition, group delay can distort the wave shape, as can echoes and amplitude tilt in the channel.

A useful oscilloscope display is called an *eye diagram*, as shown on the oscilloscope. It is produced by supplying the received waveform (before the slicer) to the vertical input of the oscilloscope. The horizontal time base is set to display more than one bit cell and is triggered by the recovered clock. (For simplicity, the diagram shows only one bit cell width, though typically some additional time would be displayed.) The eye diagram is so named for its resemblance to an eye. When a bit is undergoing a transition from 0 to 1 at the trigger time, the waveform will be positive going on the left. When it is in transition from 1 to 0, it is negative going. It may also be the same as the last bit, in which case it forms a flat line at the top or bottom of the display. Similarly, the display toward the right is formed by transitions to the next bit after the one displayed.

The slicer interprets the signal to be either a 1 or a 0. The sample time should be in the center of the bit cell, when the eye is as "open" as possible. A figure of merit is the "eye opening," or the percentage of the peak (fully opened) eye, that is seen as open in the eye diagram.

One of the practical error sources, which is often not considered in computing BER versus noise, is errors in the timing of the sample: if the timing is in error owing to problems in the clock recovery circuit (perhaps because of noise, interference, or waveform distortion), then it is more probable that the bit will be misinterpreted.

Eye closure can be caused by a number of transmission problems. The effect of noise is illustrated in Figure 4.5. Besides noise, errors in the filter shape can

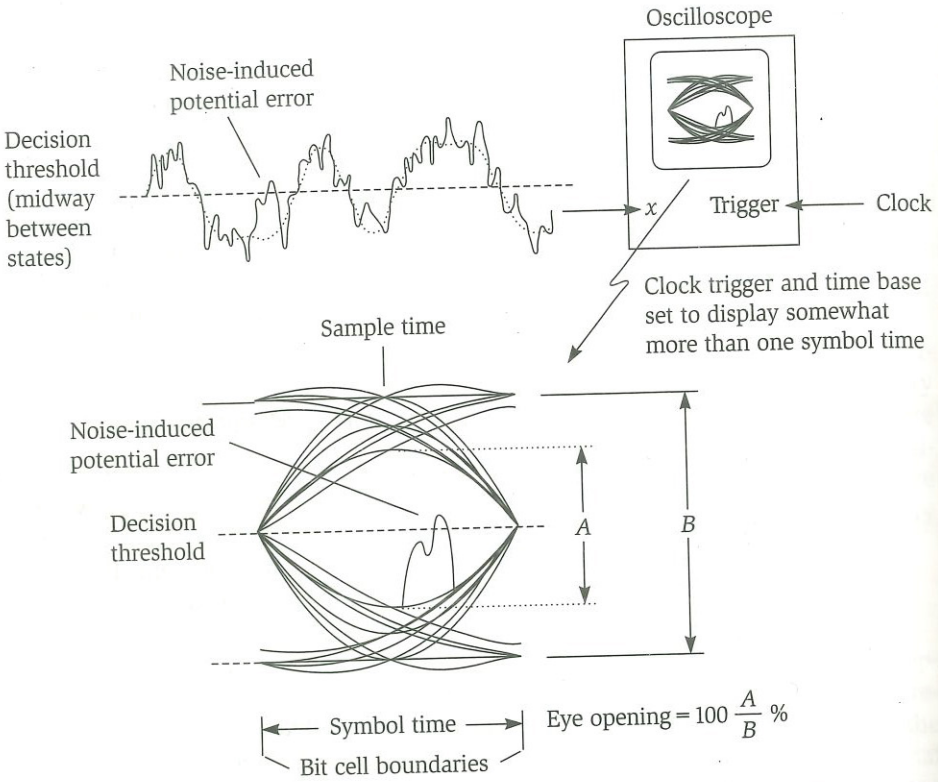


Figure 4.5 BPSK received waveform and eye diagram.

cause pulse distortion, which will tend to close the eye. If the transmission path exhibits excessive group delay, or if an echo exists, then one bit may not completely "go away" before the next bit starts. This condition is called *intersymbol interference* (ISI). The word *symbol* is used rather than *bit* for reasons to be introduced later.

Illustrated in Figure 4.5 is a potential bit error, in which a noise spike carries the waveform so far from its correct position that it actually crosses the decision threshold of the slicer. If the clock causes the slicer output to be read at the time of the noise spike, then a bit error would occur.

Peak-to-Average Ratio

Though we have argued that the BPSK signal is properly viewed as an amplitude modulated signal, the two signal states have the same amplitude. Were it not for the filtering of LPF1 (Figure 4.3), the signal would not have any amplitude char-

acteristics. However, since the signal is filtered, the transition between states takes some time, and during that time, the carrier amplitude passes through zero, again as shown in Figure 4.3. Because of this, the peak signal level is slightly higher than the average level. Digital signals in cable television work are usually measured based on the average level, yet the peak level is sometimes important.

4.2.8 Quadrature Phase Shift Keying (QPSK)

BPSK modulation provides very good immunity to noise, is simple, and is low in cost. The problem is that it is inefficient in its use of the spectrum. In order to make better use of the spectrum, it is necessary to take some small steps back toward (but not reaching) the analog domain, in that possible states are represented by different analog levels and/or phases. This is done by transmitting more than one bit at a time. The next modulation formats to be discussed are expansions of BPSK, which improve on the spectral efficiency by transmitting more than one bit at a time. The penalties are an increase in complexity and reduced immunity to noise and other transmission impairments.

Figure 4.6 illustrates a basic QPSK modulator and demodulator. Data into the transmitter is split into two channels. For reasons that will soon be apparent, the two are called the I (in-phase) and Q (quadrature) channels. Two bits are transmitted simultaneously, one in each channel. Spectral efficiency is improved because the spectrum required to transmit the two bits is no wider than the spectrum required to transmit one bit using BPSK.

After the data is split into the I and Q channels, each channel is modulated onto a carrier, just as in the case of BPSK. The same carrier frequency is used for each channel, but the phase of the two are 90° apart, or *in quadrature*. It can be shown that the two can be transmitted through a common communications channel, ideally without interfering with each other. (In the real world, several mechanisms can cause the two channels to interfere, or crosstalk, but they can be controlled.)

After the two channels are modulated, the modulated carriers are combined for transmission to the receiver. At the receiver, a phase locked loop (PLL) locks a local oscillator, LO2, to the incoming carrier. The signal from LO2 is split into in-phase and quadrature components. The in-phase component is sent to demodulator M3, and the quadrature component to the quadrature demodulator, M4. An encoder combines the two bits from the I and Q channels into the resulting datastream. A clock recovery circuit reconstructs the clock signal from the incoming data.

Bit Rate and Baud Rate

Perhaps the most common mistake made in communications terminology today is to confuse the rate at which *data* is transmitted with the rate at which *symbols*

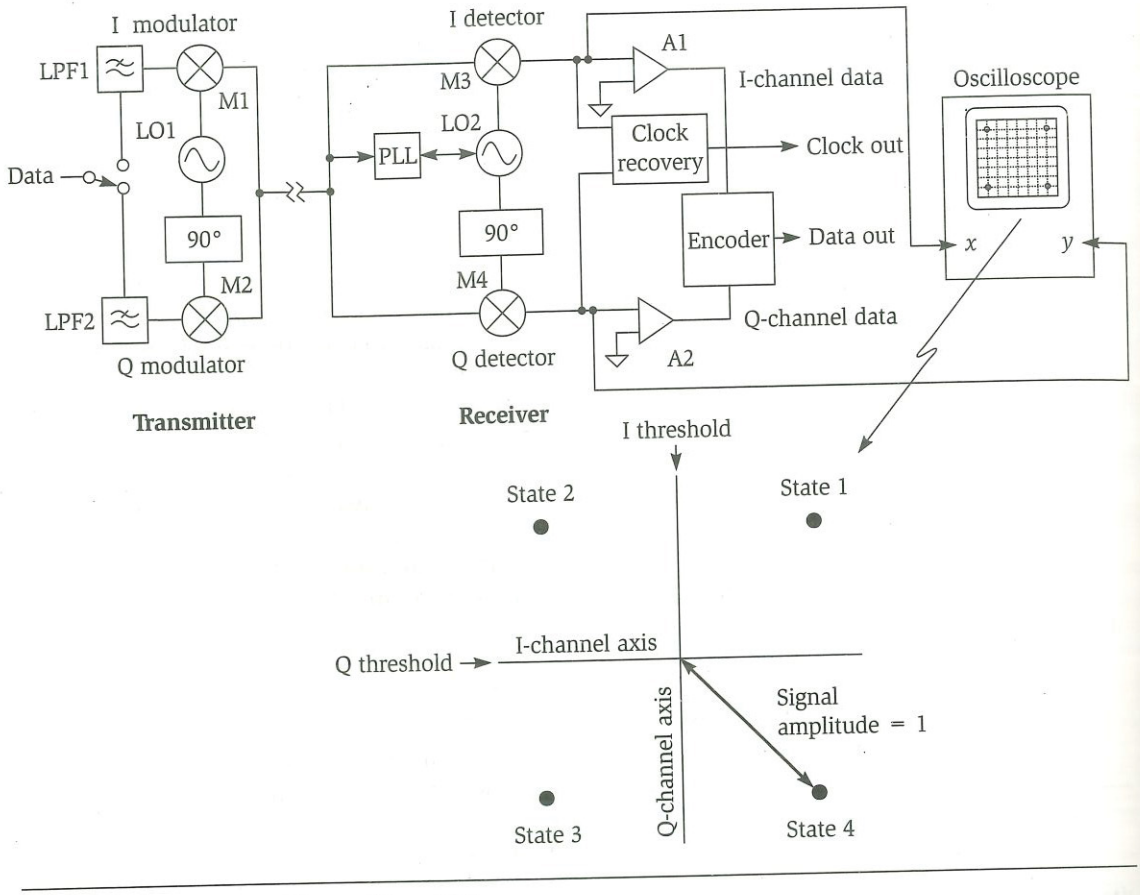


Figure 4.6 Basic QPSK modulator and demodulator.

are transmitted. The rate at which bits are transmitted (the *bit rate*) is measured in *bits per second.* However, in QPSK and most other modulation methods, more than one bit is transmitted at a time. The bits that are transmitted simultaneously are grouped into a *symbol.* The rate at which symbols are transmitted is the *baud rate.*

In an older telephone modem that employed QPSK, 2,400 bits were transmitted each second. However, since two bits were transmitted simultaneously in a symbol, the *symbol rate* was only 1,200 baud. (Note that the plural of *baud* is *baud,* without an *s.*) As we will show, 256-QAM modulation groups eight bits into a symbol, so the baud rate is only 1/8 of the bit rate. The occupied bandwidth of the signal is a function of the baud, or symbol, rate.

The Constellation Diagram

It is useful to display the eye diagram of a QPSK signal independently for the I and Q channels. Another, particularly useful, representation of QPSK and higher-order signals is the constellation diagram, shown at the bottom of Figure 4.6. It is formed by routing the demodulated I signal, before slicing, to the x axis of an oscilloscope, and the Q signal to the y axis. Each of the four possible positions of the symbol is formed from the I- and Q-channel data: the position along the x axis being determined by the I data, and that along the y axis, by the Q data. The peak amplitude of the signal is proportional to the length of the vector from the origin to any one of the four states, which are equidistant from the origin. (Note that, in the absence of any transmission impairments, the four points in the constellation are small points. They are made bigger in the illustration to make them easier to see.)

The y axis forms the decision threshold for the I channel since it is halfway between the I-channel states. Similarly, the decision threshold for the Q channel is defined by the x axis. Computation of bit error rate (BER) is based on determining the proportion of time that any one "spot" in the constellation could be forced to cross a threshold into another state's place in the constellation. When this happens, at least one, and sometimes more, bits are received in error. Obviously, the closer together the states in the constellation (that is, the "denser" the modulation), the less noise or other disturbance can be tolerated before a state can be misinterpreted as that of a neighbor.

The peak amplitude of the signal is equal to the length of a vector from the origin to the farthest point in the constellation. In the case of QPSK, all four points are equidistant from the origin, so any of them can be used as the peak amplitude. Note that the reading of peak amplitude is a voltage, not power, reading on the constellation diagram.

Though the QPSK signal has four states of the same amplitude, between states the amplitude of the carrier drops because of filtering at the modulator. This filtering is represented by LPF1 and LPF2 of Figure 4.6. Also see Figure 4.3 for an illustration of the resultant modulated waveform. Since QPSK is nothing more than two BPSK signals combined 90° out of phase, the same waveform of Figure 4.3 would generally apply to a QPSK signal: between symbols, the signal amplitude passes through 0 if the state of both bits in the symbol changes. This usually doesn't cause much concern in cable television work. However, in the case of systems having high AM/PM (pronounced "AM to PM") conversion, where a change in amplitude of the signal can cause a change in its phase, the amplitude of the signal dropping to 0 can be a problem. A satellite communications link is such a system, where the power amplifier on the satellite is driven slightly into saturation. To reduce the effect, satellite communications often use offset QPSK, in which the transition times of the data in the I and Q channels are offset by one-half of a bit cell. Since the two channels don't change at the same time, the amplitude tends to be more constant.

AM/PM conversion is a phenomenon in which changes in the amplitude of a signal result in a change to the phase of the signal. In data communications, it can distort the phase reference of a receiver, making accurate recovery of the data less likely. It can be caused by certain nonflatness of the channel and by capacitance modulation in an amplifier that is operating out of its linear range.

The Effects of Interference on the Constellation Diagram

The constellation diagram is useful for understanding the effect of interference on the signal. In Figure 4.7, quadrant I illustrates the effect of a CW interfering signal in the passband of the receiver. An interfering carrier creates a beat note, which forces each point in the constellation to spread out in a circle around the desired point. The radius of the circle is proportional to the relative amplitude of the interfering carrier. Unlike the situation that exists with analog video signals, the impact of the interfering carrier tends to be independent of its location within the passband, except for some effect caused by nonflatness of the receive filter.

Quadrant II illustrates the effect of phase noise on the signal. Phase noise is introduced when a local oscillator has significant phase noise. It can also be shown to occur when the transmission channel does not have flat response. The noise is transferred to the modulated signal. It causes the constellation to rotate about the origin.

Random noise (quadrant III) causes each point in the constellation to "blur," somewhat as with the CW interference, except that the spreading of the point tends to be more nearly uniform within the circle and the circle does not have a well-defined radius. Again, recall that the spreading of points in the constellation becomes a problem when the spread points cross a threshold to another state. With QPSK, that threshold is defined by the axes of the constellation diagram. With more dense modulation, the thresholds are closer together, so less interference will cause the threshold to be crossed. As a practical matter, an error may occur when a point is forced close to a threshold boundary: real demodulators don't work quite as well as the theory predicts.

Finally, reflections, which cause ghosting in analog transmission, cause a replication of the constellation around each point in the constellation, as illustrated in quadrant IV. The constellation is rotated with respect to the "direct path" constellation, due to phase shift between the direct and reflected RF carriers. Usually, systems that employ denser modulation formats include adaptive equalizers, which can compensate for echoes, if the echo doesn't change too quickly. QPSK systems don't usually include such equalizers because the high noise immunity of QPSK doesn't justify the cost.

If the peak amplitude of the signal is 1, as shown in quadrant IV, the distance from the nominal location of any of the four states to the nearest state boundary is 0.707. This result is easily obtained from simple geometry, where the amplitude vector, of length 1, is the hypotenuse of a right triangle, and the distance to the nearest decision threshold is one of the two other sides of the triangle. For consis-

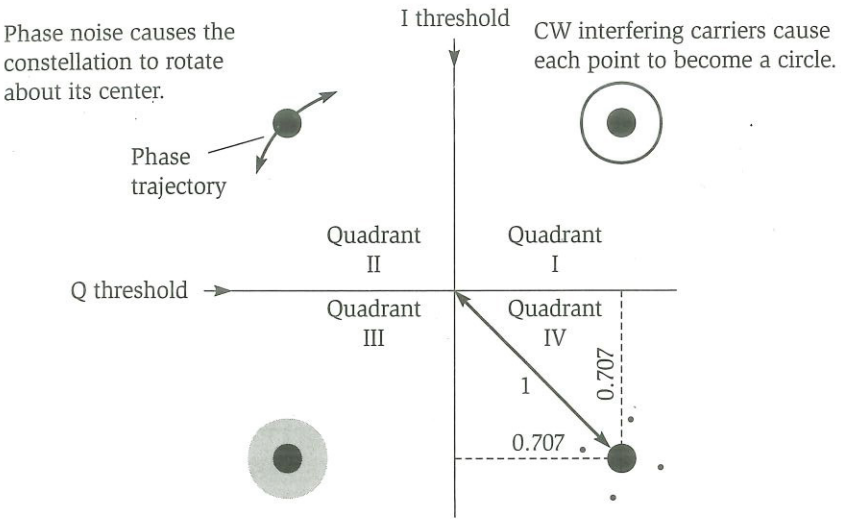


Figure 4.7 The effect of interference on the constellation diagram.

tency, we compare different constellations by normalizing them to the same peak amplitude of 1. However, bear in mind that it is more common to measure the amplitude of a digital signal by using the average amplitude, which is a concept we will explore.

4.2.9 Higher-Order QAM Modulation

16-QAM Modulation

The next denser modulation format commonly found in cable television work is 16-QAM (16-state quadrature amplitude modulation). It has been proposed for return services that require higher bandwidth efficiency than that offered by QPSK. It is similar to QPSK, except that each axis is divided into four levels so that each channel (I and Q) can carry the information of two bits. Each of two phases carries two bits at a time, so there are four bits per symbol: the baud rate is 1/4 of the bit rate.

Figure 4.8 illustrates a basic 16-QAM modulator and demodulator. Note the similarities between this and the QPSK modulator of Figure 4.6. (Clock recovery

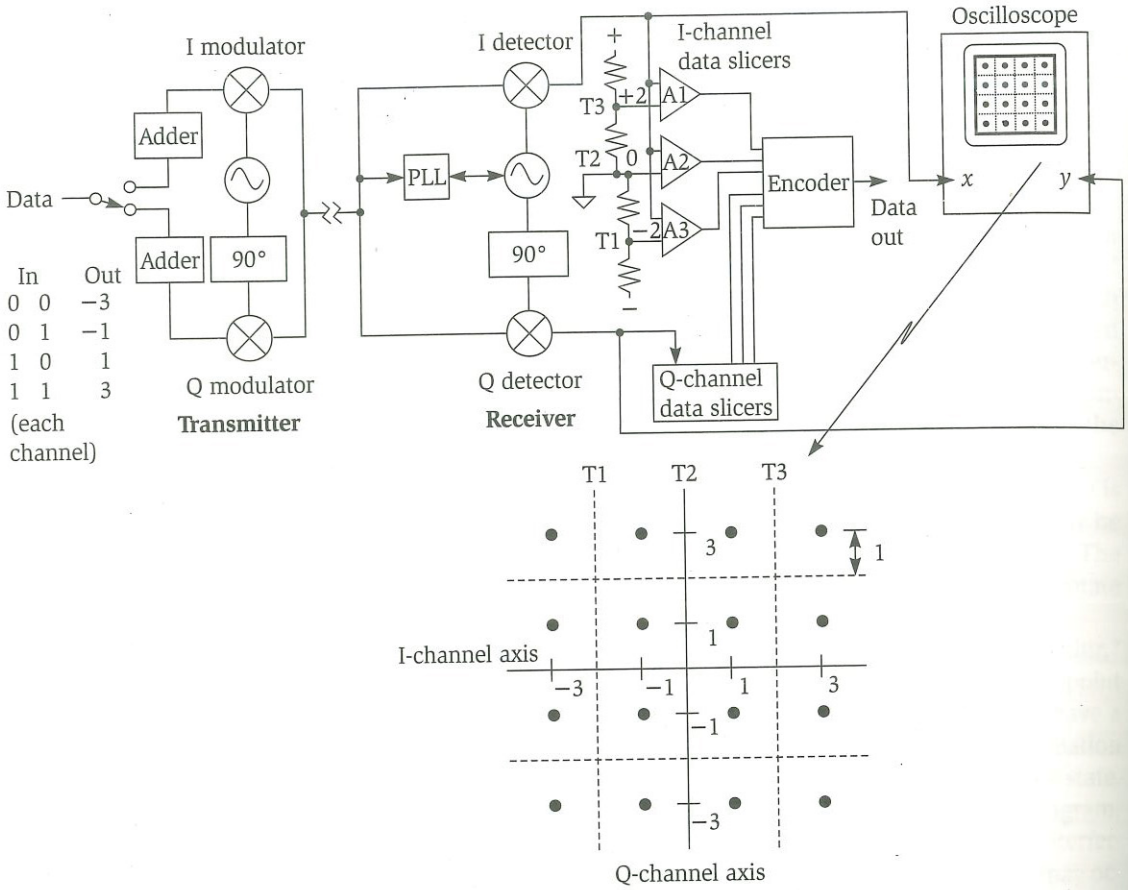


Figure 4.8 16-QAM modulator and demodulator.

is done as illustrated in Figure 4.6.) Each channel of the 16-QAM modulator can take on not only the two phases of QPSK but also the intermediate amplitude values. Data is split into two channels, the I and Q channels. Instead of only one bit being routed to each channel at a time, two bits are routed to each. They are added as shown in the table beside the Q-channel modulator: for example, if the two bits are 11 (pronounced "one-one," since we are working in binary numbers), then the output of the adder is +3, and the Q-channel modulator puts out an amplitude of 3, with a positive phase (calling this the "positive" phase is arbitrary). If the data set is 10 (one-oh), then the adder supplies an analog signal of amplitude +1 to the modulator. Similarly, an 01 data set yields an analog level of −1, which causes the modulator to supply a 1 level signal with the opposite phase. An 00 data set will cause the adder to supply −3 to the modulator, producing a higher-amplitude negative phase signal.

The same thing takes place in the I channel. Each channel can thus take on any of four states. With the I channel in any of four states, and the Q channel independently in any of four states, a total of 16 states is possible. This is the number of states required to represent 4 bits ($2^4 = 16$). The modulation technique, 16-QAM, derives its name from the 16 possible states of the modulated signal.

The receiver contains in-phase and quadrature demodulators, as does the QPSK receiver. However, the output of each demodulator must be further processed to recover two bits. This is done for the I channel, using the three data slicers shown. The data slicers have thresholds set by the four voltage divider resistors, at levels of T1, T2, and T3. These levels are set halfway between the received states in the I channel. The thresholds are shown in the constellation diagram at the bottom of Figure 4.8. By reading the outputs of the three comparators, A1 through A3, the encoder can deduce the transmitted state. The constellation diagram is formed on the oscilloscope exactly as it was for QPSK, by supplying the I-detector output to the x input, and the Q-detector output to the y input.

Binary Numbering

We showed groups of two bits whose value is designated as 00, 01, 10, or 11. The correct way to speak this is oh-oh, oh-one, and so on. These are not to be confused with the base 10 numbers that look the same. If confusion between base 2 (binary) and base 10 (our familiar numbering system) is possible, then the number will be followed by a subscript 2 or 10. The following are equivalent numbers in base 2 and base 10: $010_2 = 2_{10}$ and $1101_2 = 13_{10}$.

For downstream transmission, the industry employs modulation formats that are similar to 16-QAM except that they have more states. Put another way, they transmit more bits per symbol, so the constellation is denser. The distance between states is smaller, so the susceptibility to noise and transmission path distortion is greater. On the other hand, the spectral efficiency is greater, meaning that more bits can be transmitted in the same RF bandwidth.

64- and 256-QAM

Figure 4.9 illustrates the constellation diagrams for 4-QAM (that is, QPSK), 16-QAM, 64-QAM, and 256-QAM. The diagrams are drawn to the same peak scale, so it is easy to see the higher density of the higher-order modulation formats. Also shown for comparison are several characteristics of the different modulation formats. Most of the characteristics are shown measured in decibels with respect to a QPSK signal.

For each modulation type, the theoretical spectral efficiency in bits per hertz is shown first. The spectral efficiency is a figure of merit, defining how many bits per second can be transmitted per hertz of RF bandwidth. The actual spectral efficiency will be less than this, due to the practical need for excess filter bandwidth. Next is shown the approximate power–peak-to-average ratio. This is computed by averaging the power of each state, ignoring the effects of filtering.

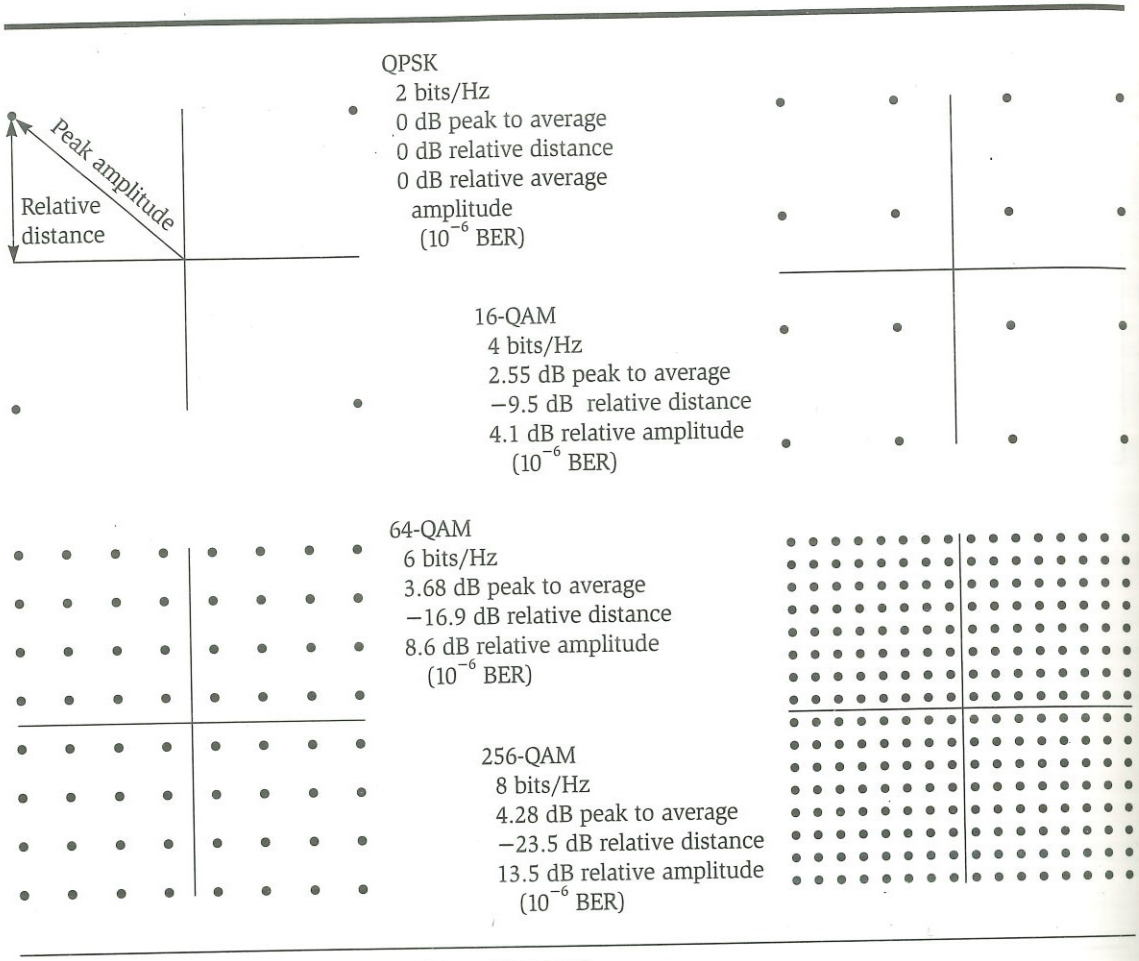


Figure 4.9 Constellation diagrams, 4-, 16-, 64-, and 256-QAM.

Section 4.2.11 shows how this is done. A more precise peak-to-average ratio must include the effects of filtering, and so cannot be stated in general.

Below the peak-to-average ratio is shown the relative distance between a state and the closest state boundary, expressed in decibels. The relative distance between a state and the boundary is defined in the QPSK constellation, as is the peak amplitude. It is the decibel measurement of how much closer to each other the states are in the denser modulation formats. The reference is a QPSK constellation, and the distance is normalized to the peak, not average, amplitude.

The relative amplitude entry is derived from the BER versus signal-to-noise ratio of the equation graphed in Figure 4.13. It is an indication of the increase in signal-to-noise ratio required to maintain a 10^{-6} bit error rate. Recall that the signal level to which we refer in this case is the average, not the peak level.

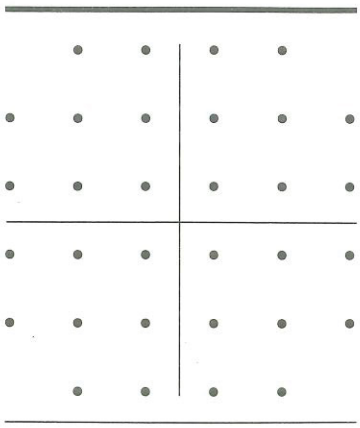


Figure 4.10 32-QAM constellation.

32-QAM Modulation

A similar modulation format to those shown earlier is 32-QAM, which has 32 states and a theoretical spectral efficiency of 5 bits per hertz. The constellation is shown in Figure 4.10. It is distinguished from the QAM formats shown above in that the four corner states are not used (this would yield 36 states). A total of 32 states represents 5 bits ($2^5 = 32$). With 16-, 64-, and 256-QAM, each axis, I and Q, can represent a certain number of bits, since the number of states along each axis is a power of 2. On the other hand, 32-QAM does not have this correspondence between bits and states. The five bits making up a symbol can, however, be encoded into one of the 32 possible states.

4.2.10 8- and 16-VSB Modulation

The FCC has designated 8-VSB modulation as the standard for off-air digital television in the United States. It is very similar in characteristics to 64-QAM, and 16-VSB is similar to 256-QAM. In VSB modulation formats, only one phase of the carrier is used: there is no quadrature component. All bits making up the symbol modulate the same phase of the carrier. That carrier (which is almost completely suppressed) is placed very near the band edge rather than in the center, and most of one sideband is removed by filtering.

Both 8-VSB and 64-QAM have essentially the same spectral efficiency. 64-QAM carries data on two quadrature phases of the carrier but must place the carrier in the center of the passband. Where carriers in quadrature are employed, it is necessary to transmit both sidebands to allow for separation of the two channels. 8-VSB uses only one phase of the carrier, so separation of components is not an issue. Compensating for the lack of the quadrature phase, 8-VSB is able to transmit twice the symbol rate in the same bandwidth since the second sideband is not transmitted.

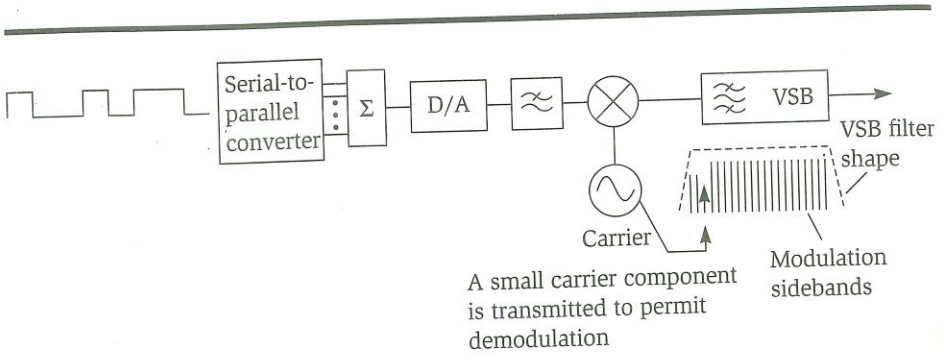


Figure 4.11 VSB modulator.

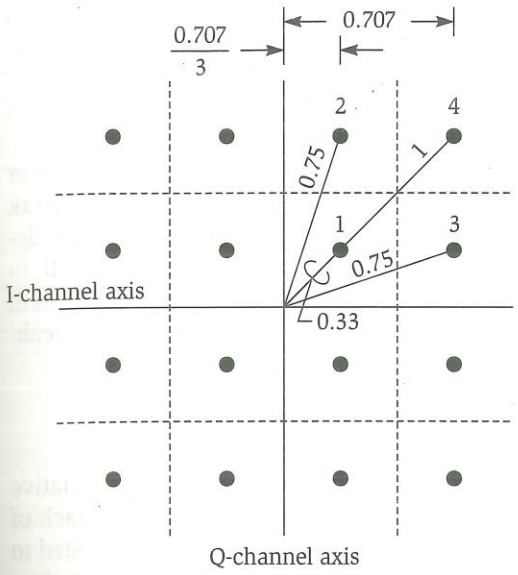
Figure 4.11 illustrates the principle of VSB digital modulation. A set of bits is converted to a parallel word. The number of bits taken at a time is three if 8-VSB ($2^3 = 8$) is being transmitted (proposed for over-the-air use), or four if 16-VSB is being transmitted. These bits are added, and the result is converted to analog and supplied to a balanced modulator. The modulator output is filtered in a vestigial sideband (VSB) filter. Ideally, the VSB filter wouldn't pass any of the lower sideband, but it would pass the carrier. As a practical matter, some of the lower sideband must be transmitted, as in the case of analog transmission. Advances in filter technology since the NTSC television system was developed have made a much smaller vestigial sideband possible.

A small amount of carrier is inserted in the VSB signal to facilitate carrier recovery. With QAM modulation formats, the carrier may be recovered by taking advantage of the symmetrical sideband nature of the signal. It may be shown that where symmetrical sidebands exist, a suppressed carrier can be recovered by doubling the frequency of the received signal, which yields a second harmonic of the carrier. This is not the case with VSB, so a small carrier component must be transmitted to allow recovery.

The susceptibility to transmission impairments of a VSB signal and the corresponding QAM signal are similar. VSB has a claimed advantage in the presence of phase noise. As shown in Figure 4.7, phase noise causes the constellation to rotate around the origin. With VSB transmission, the distance to the state boundary on the phase trajectory is longer than it is with QAM.

4.2.11 Peak-to-Average Ratio

Higher-order modulation techniques exhibit a higher peak level than the average level read on a power meter. Figure 4.12 illustrates one reason for this, based on a 16-QAM signal. To the left is the same constellation diagram depicted in Figure 4.8. The constellation is scaled such that the amplitude of any one of the four corner states is 1. This represents the peak level of the signal. The chart to the right computes the average signal level based on equal occupancy of any of the 16 pos-



Element	Length (voltage)	Length^2 (power)
1	0.33	0.11
2	0.75	0.56
3	0.75	0.56
4	1.00	1.00
Average:	0.71	0.56
Peak to average:	$20 \log \frac{1}{.71} = 3$ dB	$10 \log \frac{1}{.56} = 2.55$ dB

Figure 4.12 Peak and average level of a 16-QAM signal.

sible states. Since all four quadrants behave the same way, it is necessary to average over only one quadrant.

The voltage amplitude of each state is determined by simple geometry, assuming that the highest amplitude state has an amplitude of 1. It follows then that the coordinates of that state are (0.707, 0.707), from the geometry of a right triangle. The state closest to the center has a distance on each axis of one-third of this, so its total distance is one-third that of the farthest state. States 2 and 3 each have one coordinate of 0.707, and one coordinate of one-third of that. Application of the Pythagorean theorem shows that the magnitude is

$$\sqrt{0.707^2 + (0.707/3)^2} = 0.75.$$

When looked at on a voltage basis, the voltage average is the average length of the four vectors that define the states in the quadrant. The power average is obtained by squaring the voltage vector for each state because power is equal to the square of the voltage, divided by the resistance:

$$P = \frac{E^2}{R}$$

The resistance would cancel out in the computation, so it is omitted. The right column in Figure 4.12 computes the power–peak-to-average ratio, based on squaring the voltage vectors.

The peak-to-average ratio is not completely characterized by the state distribution as implied, however. It is also a function of the filtering at the

transmitter. This was illustrated in Figure 4.3, which shows an amplitude dip in a BPSK signal when the transmitted bit changes states. The same thing happens with higher-order modulation techniques, and that somewhat changes the peak-to-average ratio. Peak-to-average ratio is of concern in cable systems, particularly in the operation of return plant, where a limited number of digital signals exist on return spectrum handled by the return laser.

Not only can the average amplitude be less than predicted, it could be greater if certain filter shapes are used at the transmitter. The filter can ring, causing peak signal excursions exceeding the highest steady-state level of the signal. The details are a function of the filtering employed, so the peak-to-average ratio will, to an extent, be a function of the system being measured. Unless otherwise noted, the effect of filtering is not taken into account in this chapter when we state peak-to-average ratio.

Effect of Filtering on Peak-to-Average Ratio

It was shown that the peak-to-average ratio is partially a function of the relative amplitude of each state, and the percentage of time the signal spends in each of those states. Another factor, not yet taken into account, is the filtering applied to the signal. As the filtering becomes more aggressive (lower alpha corresponding to smaller excess RF bandwidth), the filter tends to ring more, meaning that the signal will be of even greater amplitude than predicted, for a small percentage of the time.

Ito[2] studied this theoretically for a filter having an alpha of 11.52%. He computed the percentage of time the power of a 64-QAM signal exceeded its average by various amounts. The curve follows the percentage computed based on state occupancy, until the percentage of time is less than the percentage of time the signal spends in the highest state. He found that the signal exceeded the average by 6 dB or more for about 0.1% of the time, and was 9 dB above the average for about 0.00001% of the time (6.25% of the time the 64-QAM signal is in the highest amplitude state, which is 3.68 dB higher than the average power).

Multiple carrier modulation such as OFDM (described in Section 4.3.3) has a much higher peak-to-average ratio. It exceeds the average by 12 dB for 0.00001% of the time. This is because multiple carrier modulation signals tend to have the amplitude distribution of random noise, which has a finite probability of being very high for a very short time. The subject of peak-to-average ratio is revisited in Chapter 14.

4.2.12 Measuring the Level of a Digitally Modulated Signal

Analog television signals are measured at the peak amplitude of the signal, which corresponds to the sync tips. This level is constant with modulation. The formal definition of the signal level is "the RMS value of the carrier amplitude during sync tips."

Digital signals are sometimes measured based on the peak level, but it is standard in cable practice to measure the signal based on average level. Measurement

of the average level is likely more accurate since it can be measured by supplying the signal to a thermocouple type RF power meter, and making a reading, usually in decibels with respect to 1 milliwatt (dBm). Such meters are simple and quite accurate. A thermocouple meter forms the basis for the signal level definition commonly used in cable television work.

The formal interpretation of "signal level" as applied to digital signals is as follows:

> The definition of the level of a digital signal shall be the power level as measured by a power meter which uses a thermocouple as a transducer. That is, the measurement shall be the average power in the signal, integrated over the actual occupied bandwidth of that signal. The signal level should be presented in decibels with respect to one millivolt RMS in a 75 ohm system. Thus the measurement reported is the RMS value of the sinusoidal voltage that would produce the same heating in a 75-ohm resistor, as does the actual signal.[3]

This definition simply means that the reported signal level is the level of a CW signal that would produce the same heating in a resistor as does the actual signal. The signal level is converted from dBm (commonly reported by power meters) to the equivalent value in decibels with respect to 1 millivolt in a 75-ohm system (dBmV), the common unit of measure of signal strength in cable television practice. The relationship between signal level measured in dBm and measured in dBmV is the same as in the analog domain:

$$\text{signal level (dBmV)} = \text{signal level (dBm)} + 48.75$$

4.2.13 Bit Error Rate Versus Carrier-to-Noise Ratio

It is appropriate to discuss the bit error rate versus the carrier-to-noise ratio of the signal. Usually in data communications, though, the independent variable is not considered to be carrier-to-noise ratio, but is expressed in terms of the energy per bit divided by the noise per hertz of bandwidth. We explain this concept and then discuss the requirements for various modulation formats.

The Concept of Eb/No

For a given modulation format, as you increase the data rate, the required transmission bandwidth will increase proportionally. If a 1-Mb/s datastream requires 0.5 MHz of transmission bandwidth, then a 2-Mb/s datastream will require 1 MHz of bandwidth. The bandwidth required has doubled, doubling the noise that of necessity must be admitted by the filtering, but the data rate has also doubled.

In order to compare the performance of different modulation formats, which may be used to transmit different bit rates, we use the concept of energy per bit divided by noise power per hertz, Eb/No (pronounced "E B over N zero"). The parameter removes data rate and noise bandwidth from the expression. It serves the same purpose as a carrier-to-noise ratio in analog video transmission, but it is not the same as carrier-to-noise ratio. You can think of this parameter as the

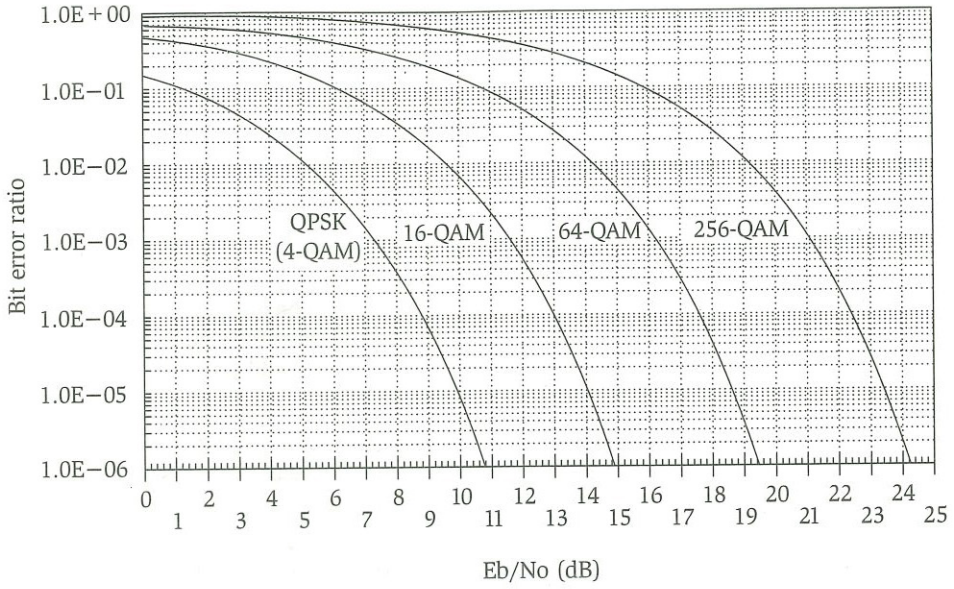


Figure 4.13 Bit error ratio versus Eb/No for several levels of QAM.

average carrier-to-noise ratio per bit. In a given case, you can convert to carrier-to-noise ratio by multiplying Eb/No by the bit rate, and then dividing by the noise bandwidth. In logarithms, this is done as follows:

$$\frac{C}{N} = \frac{Eb}{No} + 10\log(\text{bit rate in b/s}) - 10\log(\text{noise bandwidth in Hz})$$

The use of Eb/No permits you to compare the performance of different modulation systems without having to correct for different bit rates between the systems.

BER Versus Eb/No

The bit error rate of various levels of QAM modulation can be approximated by the following equation:[4]

$$P_M = 2\left(1 - \frac{1}{\sqrt{M}}\right) erfc\left(\sqrt{\frac{3k\gamma_b}{2(M-1)}}\right)\left[1 - \frac{1}{2}\left(1 - \frac{1}{\sqrt{M}}\right) erfc\left(\sqrt{\frac{3k\gamma_b}{2(M-1)}}\right)\right]$$

where

M = number of states per symbol (for example, 64 for 64-QAM)

γ_b = average C/N ratio per bit (Eb/No)

The error function, *erfc,* is a mathematical expression of the probability that a normal random variable (such as random noise) will exceed a certain level.

Graphing this expression for a number of levels of QAM yields the chart of Figure 4.13. This graph is normalized for a carrier-to-noise ratio expressed as Eb/No, the energy per bit divided by the noise per hertz of bandwidth. This is a common expression for carrier-to-noise ratio, which is independent of bit rate and bandwidth, as described in the last section.

Note that digital communications engineers often refer to this as *signal*-to-noise ratio rather than *carrier*-to-noise ratio. Cable television engineers often reserve the term *signal*-to-noise ratio for the baseband measurement, and use *carrier*-to-noise ratio for the RF measurement. Digital communications engineers don't always make this distinction.

4.3 Forms of Spectrum Sharing

This section discusses several forms of spectrum sharing, in which multiple users share the same spectrum.[5] The basic reason for spectrum sharing is better efficiency of a scarce resource (spectrum). Some of the forms shown also claim advantages in the efficacy of data transmission. Note that these forms do *not* represent new modulation techniques—they generally use some level of QAM modulation—but they do divide the spectrum differently. Nonetheless, some writers have used the term *modulation* to describe them. You will note similarities between some of the techniques; differences may sometimes be subtle.

4.3.1 Time Division Multiple Access (TDMA)

One mature method of sharing spectrum is time division multiple access (TDMA). It is being used extensively in early two-way data systems on cable because of its relative simplicity and maturity. It is quite useful for return path transmission, in which a number of return transmitters need to communicate to the headend.

Figure 4.14 illustrates time division multiple access. The reference system is a number of transmitters in one node, each of which must communicate back to the headend. In TDMA systems, each transmitter transmits on the same frequency, but at different times. After each transmitter has its turn, the sequence begins again. One set of transmissions from each transmitter is called a *frame.* Ideally, the level of each is adjusted such that a consistent receive level is measured back at the headend. TDMA systems are covered in more detail in Chapter 5. TDMA is used in several early entry cable telephony systems and is specified as the return sharing method in the DOCSIS purchase standard, covered in this chapter.

Some guard time must be allowed between individual transmissions to allow for errors in the transmission time of each individual transmitter. The receiver must also be able to lock to each transmission when it starts. Of course, since each transmitter is not transmitting full time, the transmission rate must be fast

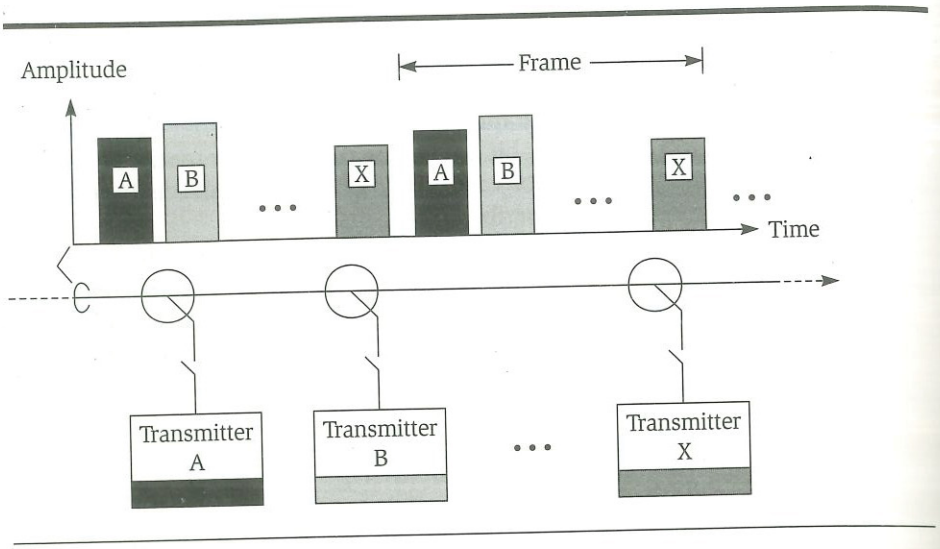


Figure 4.14 Time division multiple access.

enough to allow for each transmitter to transmit its complete set of bits during its turn. The bandwidth required is increased accordingly. A synchronizing signal must be sent from the headend, which defines the time slots.

It is not necessary that each time slot (A, B, and X) be of equal length. Several protocols, including DOCSIS, allow the length to be varied depending on the amount of data to be transmitted. This is explained in Section 4.5.4.

For downstream signal transmission, where only one transmitter is involved, a similar system can be used. The only difference is that it is not necessary to turn off one transmitter and turn on another. This transmission method is called *time division multiplexing* (TDM). It is covered in more detail in Chapter 5.

4.3.2 Frequency Division Multiplexing (FDM)

Another established method of sharing spectrum is frequency division multiplexing (FDM), as illustrated in Figure 4.15. FDM systems use a separate narrow subchannel for each transmission, dedicating the channel to that transmitter. Compared with TDMA systems, FDM requires multiple receivers at the headend though arguably digital signal-processing (DSP) techniques significantly ease that burden. (DSP uses a specialized microprocessor to implement signal-handling functions digitally. Modulation and demodulation functions, for example, can be described mathematically. Traditional circuitry approximates the appropriate mathematical functions. DSP techniques directly implement the appropriate mathematics.)

A guard band is required between channels to allow for practical filtering limitations. This causes some loss of efficiency though arguably TDMA systems also exhibit some loss of efficiency due to the need for guard times between transmissions. Any suitable modulation method may be used by each transmitter. It is

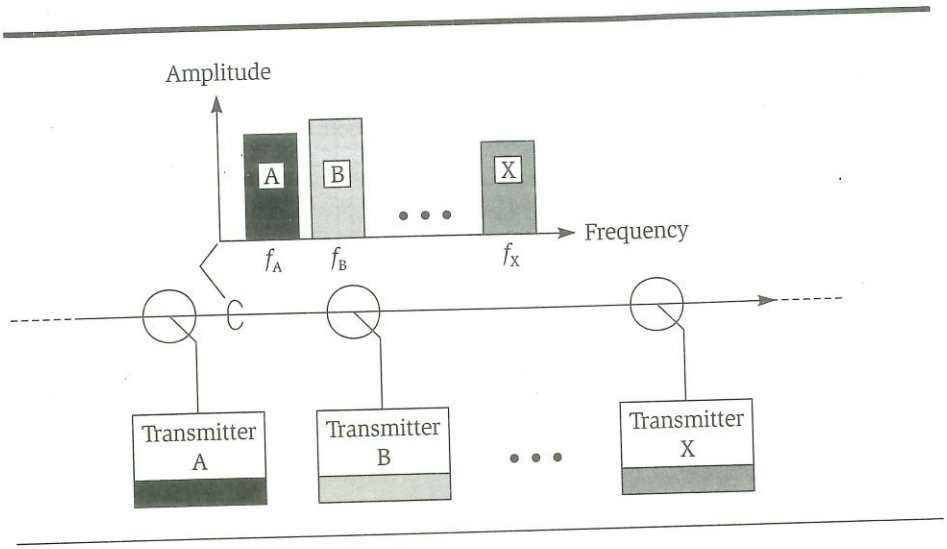


Figure 4.15 Frequency division multiplexing.

possible to monitor the error rate on each and to vary the modulation format to suit the quality of the channel.

4.3.3 Orthogonal Frequency Division Multiplexing (OFDM)

Figure 4.16 illustrates the idea of OFDM transmission. The incoming data is broken into a number of individual datastreams, each of which is then QAM modulated onto its own carrier. The carriers are spaced just far enough apart to prevent undue adjacent channel interference. Each carrier has modulation sidebands around it, as illustrated by the dotted lines in the figure (separated for clarity). The carriers may be moved closer together than you would normally expect based on the data rate if the timing of bits on adjacent carriers is carefully controlled. The adjacent signals are then said to be "orthogonal."

Any convenient modulation may be used on each of the carriers. QPSK is implied in the figure, but other techniques may be used. To generalize the type of modulation, communications engineers often use the term *m-ary* to refer to a modulation of any type. Note that OFDM is not itself a modulation technique though you will often see it discussed as if it were. However, it is a method of frequency sharing, analogous to TDM and FDM.

By dividing the spectrum as shown, you replaces a single band of high-speed data with a large number of bands of lower-speed data. Since demodulation is done over a narrower band, the effects of impulse noise and group delay are reduced, and the effective carrier-to-noise ratio is improved in each channel by the narrowing of the passband.

It is possible to monitor the error rate on each carrier individually. If one channel develops an unacceptable error rate, its density of modulation can be reduced (for example, go from 16-QAM to QPSK modulation). This will necessitate

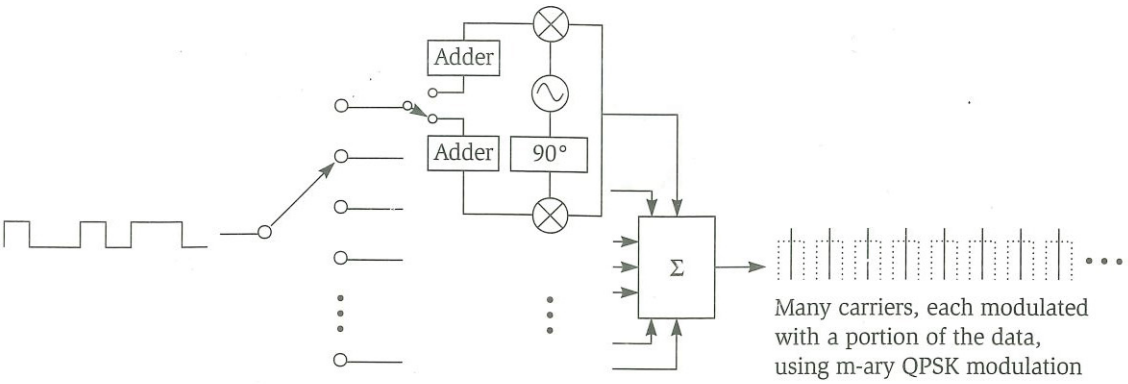


Figure 4.16 Principle of OFDM transmission.

reducing the bit rate on the carrier, meaning that either some bits will have to be reassigned to another channel or the overall bit rate will have to be reduced.

For return path services, it is possible to allow bits on different subchannels to originate from different transmitters. In this case, the technique has been called orthogonal frequency division multiple access (OFDMA). It is also possible to utilize each subchannel in a time division multiple access form, where each subchannel is shared by many transmitters, one at a time. This has been called variable constellation multitone (VCMT).

A closely allied sharing method is synchronous discrete multitone (S-DMT). Adjacent subchannels are synchronized such that spectrum overlap is minimized, allowing closer spacing of carriers. This effectively negates the disadvantage of having to allow guard bands between channels. It does require careful synchronization between transmitters, however. Adjacent channels so synchronized are said to be *orthogonal.* Yet another similar method is known as discrete wavelet multitone (DWMT). The method again assigns small subchannels to carry data in long symbols (a low data rate in each subchannel). You might observe that some of these methods are different names for the same basic technique.

4.3.4 Code Division Multiple Access (CDMA)

Code division multiple access (CDMA) is a form of spread spectrum communications that is being used for, among other applications, personal communications systems (PCSs), a newer form of cellular telephone service using particularly small cells. (A second PCS system uses TDMA.) Some vendors use it in cable telephony return path transmission. The method is also known as SCDMA, or synchronous CDMA.

Figure 4.17 illustrates the principle of CDMA transmission. The basic modulation is often QAM of some level. The data to be modulated on each axis is exclu-

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