King Fahd University of Petroleum & Minerals
Department of Electrical Engineering

Communications Engineering I
EE 370

Course Notes
Chapter 7

Prepared By:
Dr. Azzedine Zerguine

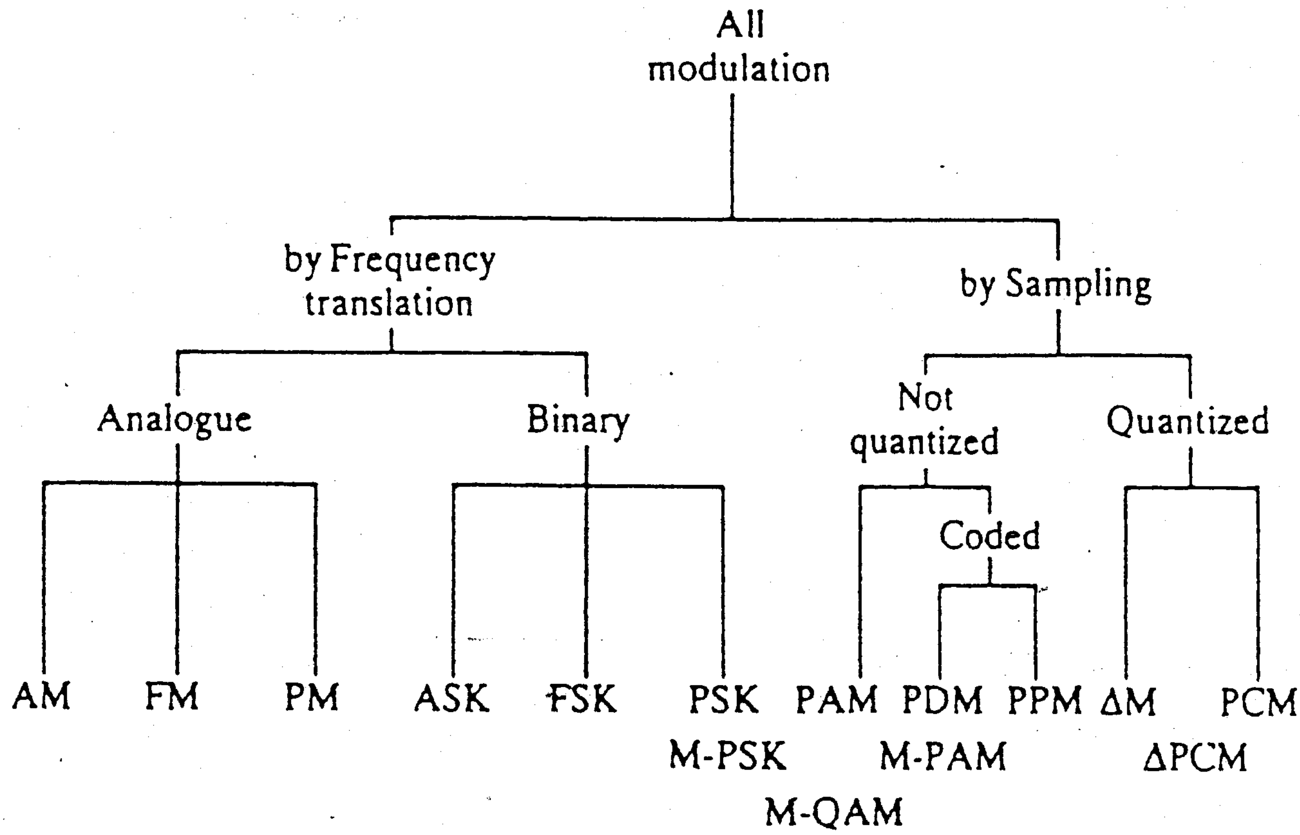


Figure Family tree of modulation methods

Principles of Digital Data Transmission

Digital Communication Systems:

Such system only deal with messages, or baseband signals, that are in digital format and not analog. Transmission of digital data can be classified into <u>two</u> categories:

(i) Baseband digital systems, where the signals are transmitted directly without any shift in the frequency of the signal.

(ii) Digital carrier systems, where the spectrum of the baseband digital signal is shifted to a higher frequency by modulating a high frequency sinusoidal carrier.

A baseband digital communication system is composed of several components.

1) Source:

This contains the information (message) in the form of a sequence of digits. Examples of such information can be a data set, a computer file, a digitized voice signal (PCM), a digital TV, a digital facsimile, --- etc.

2) Multiplexer:

The capacity of a practical channel exceeds the data rate of individual sources. Hence, in order to utilize the capacity of a channel efficiently, we combine different sources through a digital multiplexer using the process of interleaving (time-sharing).

Thus, a channel is time-shared by several messages simultaneously using TDM.

Line Coder:

The output of a multiplexer is coded into electrical pulses for transmission over the channel. This process is called line coding. There are many ways of assigning pulses to the digital data:

a- on-off (RZ)

b- polar (RZ)

c- bipolar (RZ), or alternate mark inversion (AMI), where 0 is encoded by no pulse and 1 is encoded by a pulse p(t) or -p(t), depending on whether the previous 1 is encoded by -p(t) or p(t).

d- on-off (NRZ)

e- Polar (NRZ)

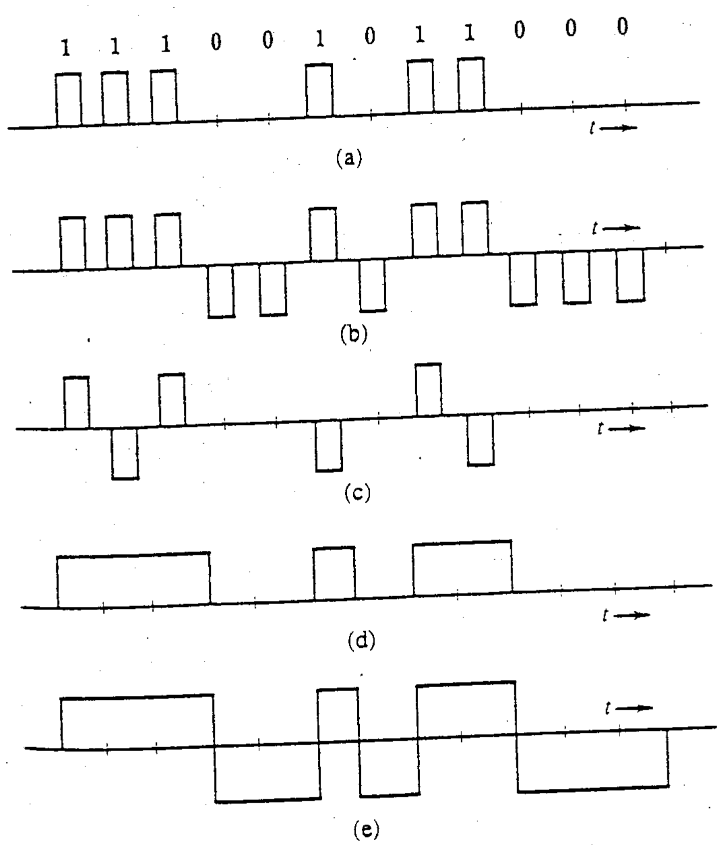


Figure 7.1 Some line codes. (a) On-off (RZ). (b) Polar (RZ). (c) Bipolar (RZ). (d) On-off (NRZ). (e) Polar (NRZ).

4) Regenerative Repeater :

An important feature of digital transmission is the ability to reconstruct the transmitted pulse train almost perfectly even though the channel (medium) may be noisy and dispersive. When this process of reconstruction is performed at intervals along the transmission path by **regenerative repeaters,** it becomes possible to transmit the digital signal over large distances without any impairment which will be noticeable or perceptible. Such regenerative repeaters perform (1) equalisation, (2) timing recovery and (3) regeneration.

Baseband Digital Transmission

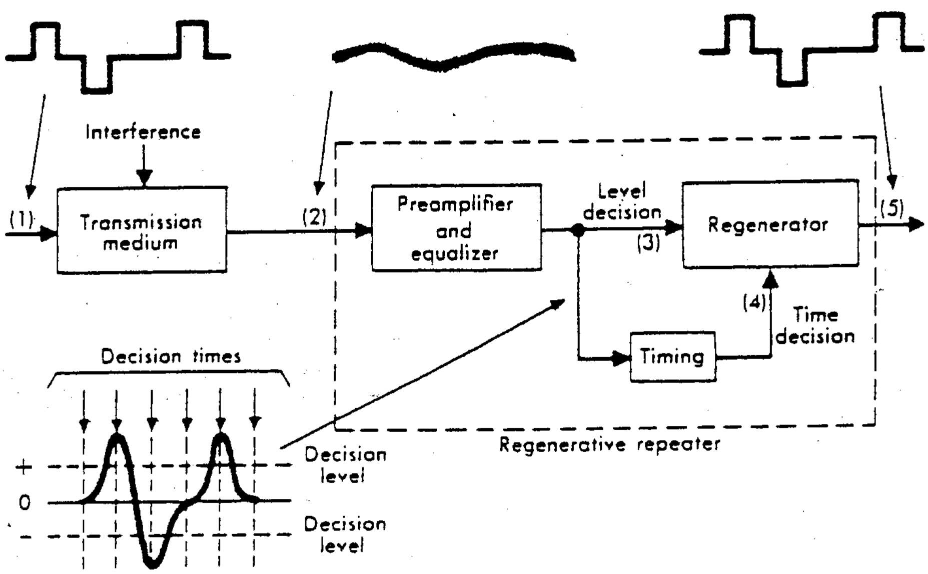


Fig. 14 Regenerative repeater section.

The signal at point 2 has experienced attenuation and dispersion by the transmission medium and it is also corrupted by additive noise.

The equaliser has the primary function of compensating for the distortion introduced by the channel (medium), and reshaping the pulses so that a correct decision can be made.

The regenerator will produce a clean pulse as shown in the previous figure. In practice, however, errors may still be present. If the interference at the decision time is sufficiently large, an incorrect decision may be taken by the regenerator and the pulse train at point 5 will have errors.

If the pulses are transmitted at a rate of R_b pulses per second, then we require the periodic timing information (the clock signal at R_b Hz) to sample the incoming pulses at a repeater. This timing information can be extracted from the received signal itself if the line code is chosen properly. For example, the polar (RZ) signal will give a periodic signal of clock frequency R_b Hz, when rectified.

Line Coding:

Digital data can be transmitted by using the different line codes such as on-off, polar, bipolar, -- etc. The desirable properties of a line code are:

1. The transmission bandwidth should be as small as possible.
2. For a given bandwidth and detection error probability, the transmitted power should be as small as possible.
3. Error detection & correction capability: it should be possible to detect, and possibly correct, detection errors. For example, in AMI codes, a single error will cause bipolar violation and can be easily detection.
4. Favourable Power Spectral Density (PSD): It is desirable to have zero PSD at dc ($f = 0$), because ac coupling and transformers are used at the repeaters.
5. Timing Information: It should be possible to extract timing or clock information from the line code signal.

PSD of Various Line Codes:

* Polar (RZ) line code with equal probabilities of binary 1 and binary 0.

  The pulse shape is rectangular of width $T_b/2$.

$$\therefore\ p(t) = \text{rect}\left(\frac{t}{T_b/2}\right) = \text{rect}\left(\frac{2t}{T_b}\right)$$

$$\therefore\ P(f) = \frac{T_b}{2}\,\text{sinc}\left(\frac{fT_b}{2}\right)$$

$$PSD = S(f) = \frac{T_b}{4}\,\text{sinc}^2\left(\frac{fT_b}{2}\right)$$

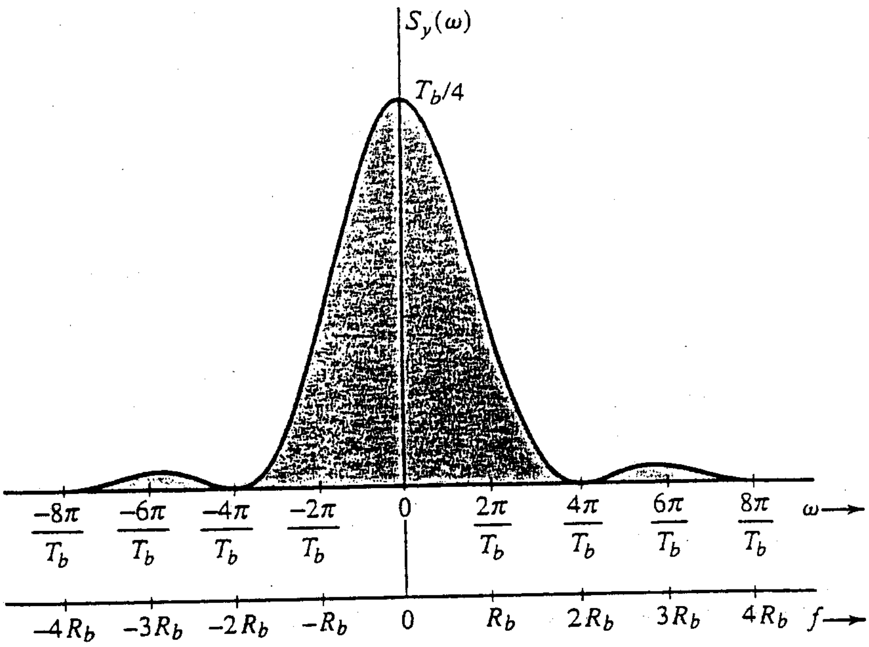


Figure 7.5 Power spectral density of a polar signal.

* On-off (RZ) line code with equal probabilities of binary 1 and binary 0.

  The pulse shape is rectangular of width $T_b/2$.

$$p(t) = \text{rect}\left(\frac{2t}{T_b}\right)$$

$$S(f) = \frac{T_b}{16}\,\text{sinc}^2\left(\frac{fT_b}{2}\right)\left[1 + \frac{1}{T_b}\sum_{n=-\infty}^{\infty}\delta\left(f - \frac{n}{T_b}\right)\right]$$

Note that $S(f)$ contains discrete & continuous parts.

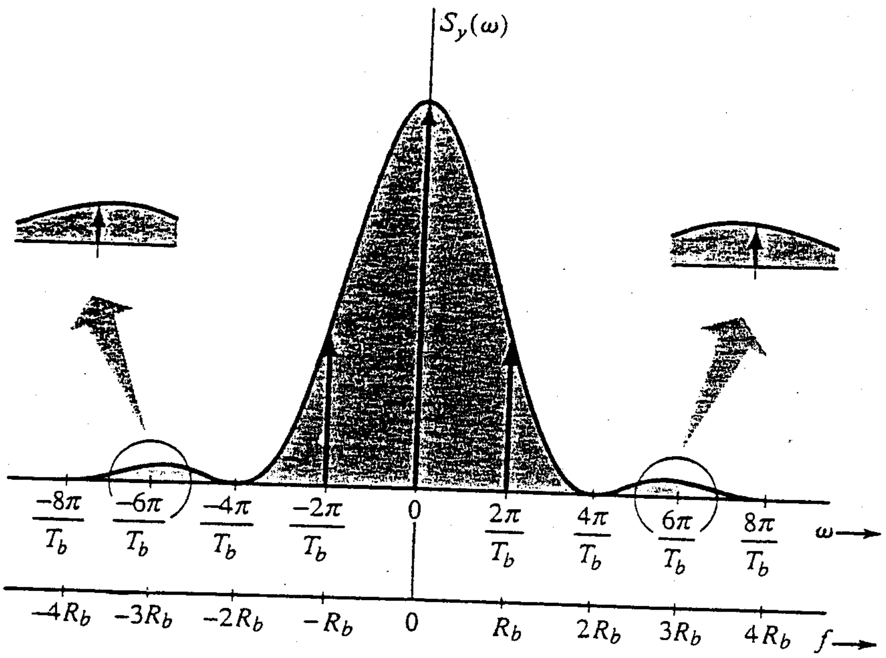


Figure 7.7 Power spectral density of an on-off signal.

* Bipolar or AMI (RZ) line codes.

$$S(f) = \frac{T_b}{4}\,\text{sinc}^2\left(\frac{f T_b}{2}\right) \cdot \sin^2(\pi f T_b)$$

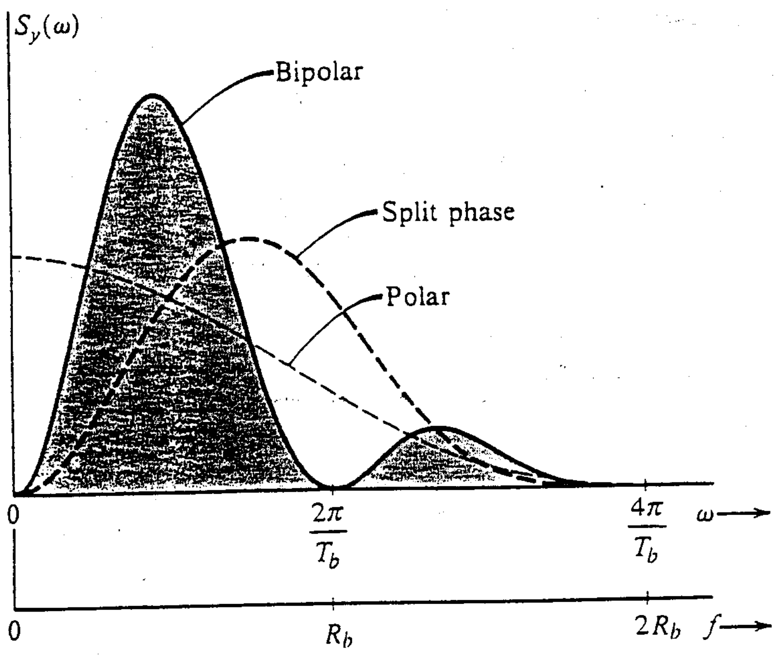


Figure 7.8 PSD of bipolar, polar, and split-phase signals normalized for equal powers. Half-width rectangular pulses are used.

7.3 PULSE SHAPING

The PSD $S_y(\omega)$ of a digital signal $y(t)$ can be controlled by a choice of line code or by the pulse shape $P(\omega)$. In the last section , we discussed how the PSD is controlled by a line code. In this section, we examine how $S_y(\omega)$ is influenced by the pulse shape $p(t)$, and how to shape a pulse $p(t)$ in order to achieve a desired $S_y(\omega)$. The PSD $S_y(\omega)$ is strongly and directly influenced by the pulse shape $p(t)$ because $S_y(\omega)$ contains the term $|P(\omega)|^2$. Thus, compared to the nature of the line code, the pulse shape is a much more potent factor in terms of shaping the PSD $S_y(\omega)$.

In the last section, we used a simple half-width rectangular pulse $p(t)$ for the sake of illustration. Strictly speaking, in this case the bandwidth of $S_y(\omega)$ is infinite since $P(\omega)$ has infinite bandwidth. But we found that the essential bandwidth of $S_y(\omega)$ was finite. For example, most of the power of a bipolar signal is contained within the essential band of 0 to R_b Hz. Note, however, that the PSD is small but is still nonzero in the range of $f > R_b$ Hz. Therefore, when such a signal is transmitted over a channel of bandwidth R_b Hz, a significant portion of its spectrum is transmitted, but a small portion of the spectrum is suppressed. In Sec. 3.6, we saw how such a spectral distortion tends to spread the pulse (dispersion). Spreading of a pulse beyond its interval T_b will cause it to interfere with neighboring pulses. This is known as **intersymbol interference (ISI)**, which can cause errors in the correct detection of pulses.

To resolve the difficulty of ISI, let us review briefly our problem. We need to transmit a pulse every T_b interval, the kth pulse being $a_k p(t-kT_b)$. The channel has a finite bandwidth, and we are required to detect the pulse amplitude a_k correctly (that is, without ISI). In our discussion so far, we are considering time-limited pulses. Since such pulses cannot be band-limited, part of their spectra is suppressed by a band-limited channel. This causes pulse distortion (spreading out) and, consequently, ISI. We can try to resolve this difficulty by using pulses which are band-limited to begin with so that they can be transmitted intact over a band-limited channel. But band-limited pulses cannot be time-limited. Obviously, various pulses will overlap and cause ISI. Thus, whether we begin with time-limited pulses or band-limited pulses, it appears that ISI cannot be avoided. It is inherent in the finite transmission bandwidth. Fortunately, there is an escape from this blind alley. Pulse amplitudes can be detected correctly despite pulse spreading (or overlapping) if there is no ISI at the decision-making instants. This can be accomplished by a properly shaped band-limited pulse. To eliminate ISI, Nyquist proposed three different criteria for pulse shaping.[5] We shall consider only the first two criteria. The third is inferior to the first two[6], and, hence, will not be considered here.

7.3.1 Nyquist Criterion for Zero ISI

In the first method, Nyquist achieves zero ISI by choosing a pulse shape that has a nonzero amplitude at its center (say $t = 0$) and zero amplitudes at $t = \pm nT_b$ ($n = 1, 2, 3, \ldots$), where T_b is the separation between successive transmitted pulses (Fig. 7.10a),

$$p(t) = \begin{cases} 1 & t = 0 \\ 0 & t = \pm nT_b \end{cases} \quad \left(T_b = \frac{1}{R_b}\right) \tag{7.22}$$

A pulse satisfying this criterion causes zero ISI at all the remaining pulse centers, or signaling instants is shown in Fig. 7.10a, where we show several successive pulses (dotted) centered at

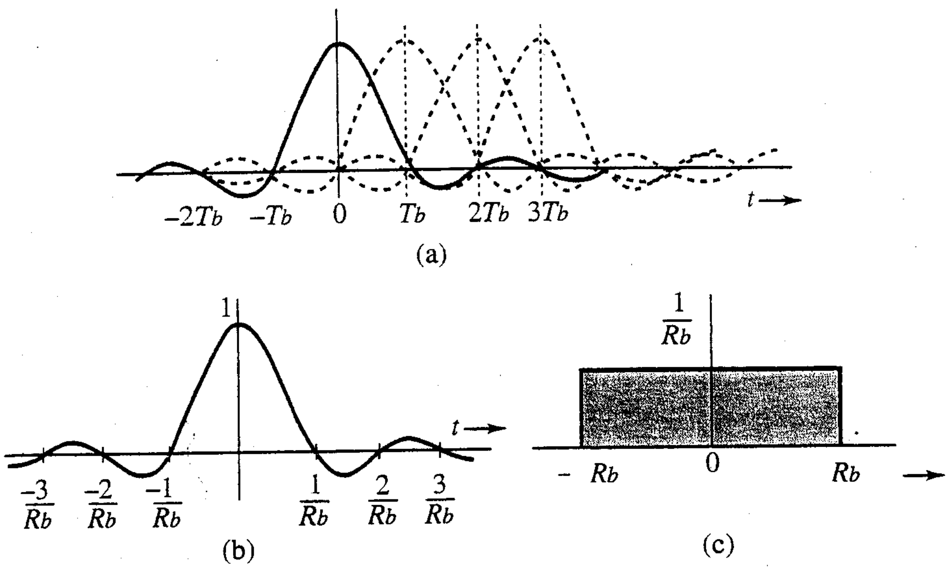


Figure 7.10 Minimum bandwidth pulse that satisfies the Nyquist criterion and its spectrum.

$t = 0,\ T_b,\ 2T_b,\ 3T_b,\ \ldots\ (T_b = 1/R_b)$. For the sake of convenience we have shown all pulses to be positive.* It is clear from this figure that the samples at $t = 0,\ T_b,\ 2T_b,\ 3T_b,\ \ldots$ consist of the amplitude of only one pulse (centered at the sampling instant) with no interference from the remaining pulses.

Now transmission of R_b bit/s requires a theoretical minimum bandwidth of $R_b/2$ Hz. It would be nice if a pulse satisfying Nyquist's criterion had this minimum bandwidth $R_b/2$ Hz. Can we find such a pulse $p(t)$? We have already solved this problem in Example 6.1 (with $B = R_b/2$), where we showed that there exists one (and only one) pulse that meets Nyquist's criterion (7.22) and has a bandwidth $R_b/2$ Hz. This pulse, $p(t) = \text{sinc}\,(\pi R_b t)$, (see Fig. 7.10b) has the property

$$\text{sinc}\,(\pi R_b t) = \begin{cases} 1 & t = 0 \\ 0 & t = \pm n T_b \end{cases} \quad \left(T_b = \frac{1}{R_b}\right) \tag{7.23a}$$

Moreover, the Fourier transform of this pulse is

$$P(\omega) = \frac{1}{R_b}\text{rect}\left(\frac{\omega}{2\pi R_b}\right) \tag{7.23b}$$

which has a bandwidth $R_b/2$ Hz, as seen from Fig. 7.10c. Using this pulse, we can transmit at a rate of R_b pulses per second without ISI, over a bandwidth of $R_b/2$.

This scheme shows that we can attain the theoretical limit of performance by using a sinc pulse. Unfortunately, this pulse is impractical because it starts at $-\infty$. We will have to wait an infinite time to generate it. Any attempt to truncate it would increase its bandwidth beyond $R_b/2$ Hz. But even if this pulse were realizable, it has the undesirable feature that it decays too slowly at a rate $1/t$. This causes some serious practical problems. For instance, if the nominal data rate of R_b bit/s required for this scheme deviates a little, the pulse amplitudes will not vanish at the other pulse centers. Because the pulses decay only as $1/t$, the cumulative interference at any pulse center from all the remaining pulses is of the form $\sum(1/n)$. It is well known that an infinite series of this form does not converge and can add up to a very

* Actually, a pulse corresponding to **0** would be negative. But considering all positive pulses does not affect our reasoning. Showing negative pulses would make the figure needlessly confusing.

large value. A similar result occurs if everything is perfect at the transmitter, but the sampling rate at the receiver deviates from the rate of R_b Hz. Again, the same thing happens if the sampling instants deviate a little because of pulse time jitter, which is inevitable even in the most sophisticated systems. This scheme therefore fails unless everything is perfect, which is a practical impossibility. And all this is because sinc $(\pi R_b t)$ decays too slowly (as $1/t$). The solution is to find a pulse $p(t)$ that satisfies Eq. (7.22) but decays faster than $1/t$. Nyquist has shown that such a pulse requires a bandwidth of $kR_b/2$, with $1 \le k \le 2$.

This can be proved as follows. Let $p(t) \Longleftrightarrow P(\omega)$, where the bandwidth of $P(\omega)$ is in the range $(R_b/2,\ R_b)$ (Fig. 7.11a). The desired pulse $p(t)$ satisfies Eq. (7.22). If we sample $p(t)$ every T_b seconds by multiplying $p(t)$ by $\delta_{T_b}(t)$ (an impulse train), then because of the property (7.22), all the samples, except the one at the origin, are zero. Thus, the sampled signal $\overline{p(t)}$ is

$$\overline{p}(t) = p(t)\delta_{T_b}(t) = \delta(t) \tag{7.24}$$

From Eq. (6.4) we know that the spectrum of a sampled signal $\overline{p}(t)$ is ($1/T_b$ times) the spectrum of $p(t)$ repeating periodically at intervals of the sampling frequency ω_b. Therefore, the Fourier transform of both sides of Eq. (7.24) yields

$$\frac{1}{T_b}\sum_{n=-\infty}^{\infty} P(\omega - n\omega_b) = 1 \qquad \omega_b = \frac{2\pi}{T_b} = 2\pi R_b \tag{7.25}$$

or

$$\sum_{n=-\infty}^{\infty} P(\omega - n\omega_b) = T_b \tag{7.26}$$

Thus, the sum of the spectra formed by repeating $P(\omega)$ every ω_b is a constant T_b, as shown in Fig. 7.11b.*

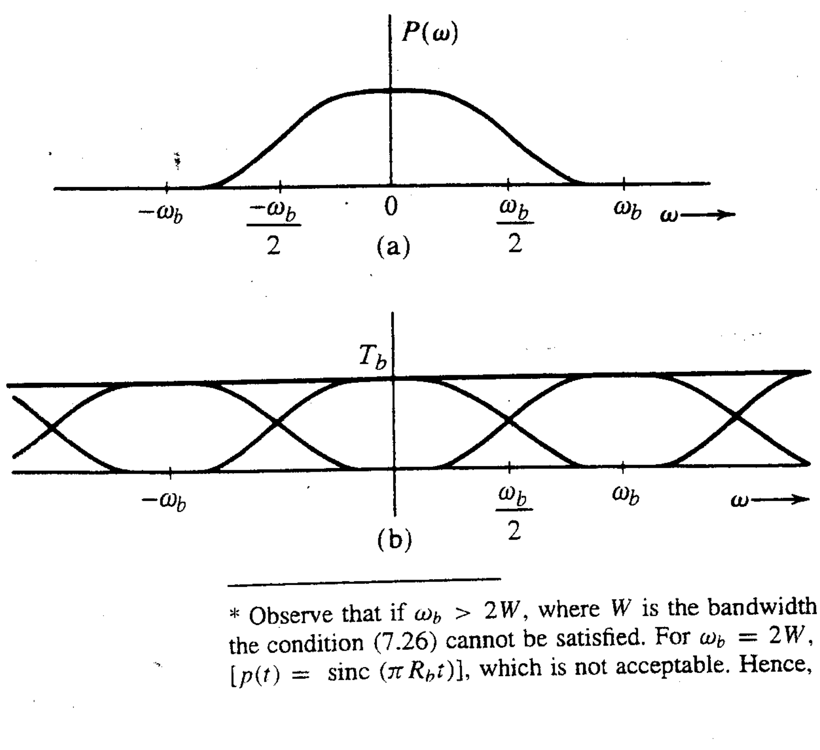


Figure 7.11 Derivation of the Nyquist criterion pulse.

* Observe that if $\omega_b > 2W$, where W is the bandwidth of $P(\omega)$, the repetitions of $P(\omega)$ are nonoverlapping, and the condition (7.26) cannot be satisfied. For $\omega_b = 2W$, the condition is satisfied only for the ideal low-pass $P(\omega)$ $[p(t) =$ sinc $(\pi R_b t)]$, which is not acceptable. Hence, we must have $W > \omega_b/2$.

Consider the spectrum in Fig. 7.11b over the range $0 < \omega < \omega_b$. Over this range only the two terms $P(\omega)$ and $P(\omega - \omega_b)$ in the summation in Eq. (7.26) are involved. Hence,

$$P(\omega) + P(\omega - \omega_b) = T_b \qquad 0 < \omega < \omega_b$$

Letting $\omega = x + \omega_b/2$,

$$P\left(x + \frac{\omega_b}{2}\right) + P\left(x - \frac{\omega_b}{2}\right) = T_b \qquad |x| < \frac{\omega_b}{2} \tag{7.27}$$

Use of the result in Eq. (3.9) in Eq. (7.27) yields

$$P\left(\frac{\omega_b}{2} + x\right) + P^*\left(\frac{\omega_b}{2} - x\right) = T_b \qquad |x| < \frac{\omega_b}{2} \tag{7.28}$$

If we assume $P(\omega)$ of the form

$$P(\omega) = |P(\omega)|e^{-j\omega t_d}$$

then the term $e^{-j\omega t_d}$ represents pure time delay, and only $|P(\omega)|$ needs satisfy Eq. (7.28). Because $|P(\omega)|$ is real, Eq. (7.28) implies

$$\left|P\left(\frac{\omega_b}{2} + x\right)\right| + \left|P\left(\frac{\omega_b}{2} - x\right)\right| = T_b \qquad |x| < \frac{\omega_b}{2} \tag{7.29}$$

Hence, $|P(\omega)|$ should be of the form shown in Fig. 7.12. This curve has an odd symmetry about the set of axes intersecting at point x [the point on the $|P(\omega)|$ curve at $\omega = \omega_b/2$]. Note that this requires that $|P(\omega_b/2)| = 0.5|P(0)|$.

The bandwidth of $P(\omega)$ is $(\omega_b/2) + \omega_x$, where ω_x is the bandwidth in excess of the theoretical minimum bandwidth. Let r be the ratio of the excess bandwidth ω_x to the theoretical minimum bandwidth $\omega_b/2$;

$$r = \frac{\text{excess bandwidth}}{\text{theoretical minimum bandwidth}} = \frac{\omega_x}{\omega_b/2} = \frac{2\omega_x}{\omega_b} \tag{7.30}$$

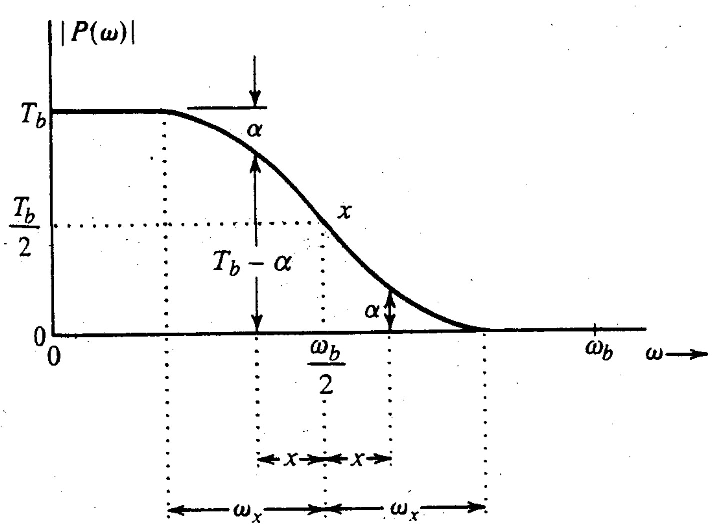


Figure 7.12 Vestigial spectrum.

Observe that because ω_x is at most equal to $\omega_b/2$,

$$0 \leq r \leq 1 \tag{7.31}$$

The theoretical minimum bandwidth is $R_b/2$ Hz, and the excess bandwidth is $f_x = rR_b/2$ Hz. Therefore, the bandwidth of $P(\omega)$ is

$$B_T = \frac{R_b}{2} + \frac{rR_b}{2} = \frac{(1+r)R_b}{2} \tag{7.32}$$

The constant r is called the **roll-off factor** and is also expressed in percent. For example, if $P(\omega)$ is a Nyquist criterion spectrum with a bandwidth that is 50% higher than the theoretical minimum, its roll-off factor $r = 0.5$, or 50%.

Although the phase of $P(\omega)$ must be linear up to the frequency where $|P(\omega)|$ goes to 0, for most practical applications it is sufficient to equalize the phase characteristics up to the 10- to 15-dB attenuation point [i.e., the point where $P(\omega)$ is 10 to 15 dB below its peak]. A filter having an amplitude response with the same characteristics is required in the vestigial sideband modulation discussed in Sec. 4.6. For this reason, we shall refer to the spectrum $P(\omega)$ in Eqs. (7.28) and (7.29) as a **vestigial spectrum**.

The pulse $p(t)$ in Eq. (7.22) has zero ISI at the centers of all other pulses transmitted at the rate of R_b pulses per second. A pulse $p(t)$ that causes zero ISI at the centers of all the remaining pulses (or signaling instants) is the Nyquist criterion pulse. We have shown that a pulse with a vestigial spectrum [Eq. (7.28) or Eq. (7.29)] satisfies the Nyquist criterion for zero ISI.

Because $0 \leq r < 1$, the bandwidth of $P(\omega)$ is restricted to the range of $R_b/2$ to R_b Hz. The pulse $p(t)$ can be generated as a unit impulse response of a filter with transfer function $P(\omega)$. But because $P(\omega) = 0$ over a band, it violates the Paley-Wiener criterion and is therefore unrealizable. However, the vestigial roll-off characteristic is gradual, and it can be more closely approximated by a practical filter. One family of spectra that satisfies the Nyquist criterion is

$$P(\omega) = \begin{cases} \frac{1}{2}\left\{1 - \sin\left(\frac{\pi[\omega - (\omega_b/2)]}{2\omega_x}\right)\right\} & \left|\omega - \frac{\omega_b}{2}\right| < \omega_x \\ 0 & |\omega| > \frac{\omega_b}{2} + \omega_x \\ 1 & |\omega| < \frac{\omega_b}{2} - \omega_x \end{cases} \tag{7.33}$$

Figure 7.13a shows three curves, corresponding to $\omega_x = 0$ $(r = 0)$, $\omega_x = \omega_b/4$ $(r = 0.5)$, and $\omega_x = \omega_b/2$ $(r = 1)$. The respective impulse responses are shown in Fig. 7.13b. It can be seen that increasing ω_x (or r) improves $p(t)$; that is, more gradual cutoff reduces the oscillatory nature of $p(t)$ and causes it to decay more rapidly. For the case of the maximum value of $\omega_x = \omega_b/2$ $(r = 1)$, Eq. (7.33) reduces to

$$P(\omega) = \frac{1}{2}\left(1 + \cos\frac{\omega}{2R_b}\right)\text{rect}\left(\frac{\omega}{4\pi R_b}\right) \tag{7.34a}$$

$$= \cos^2\left(\frac{\omega}{4R_b}\right)\text{rect}\left(\frac{\omega}{4\pi R_b}\right) \tag{7.34b}$$

This characteristic is known in the literature as the **raised-cosine** characteristic, because it represents a cosine raised by its peak amplitude. It is also known as the **full-cosine roll-**

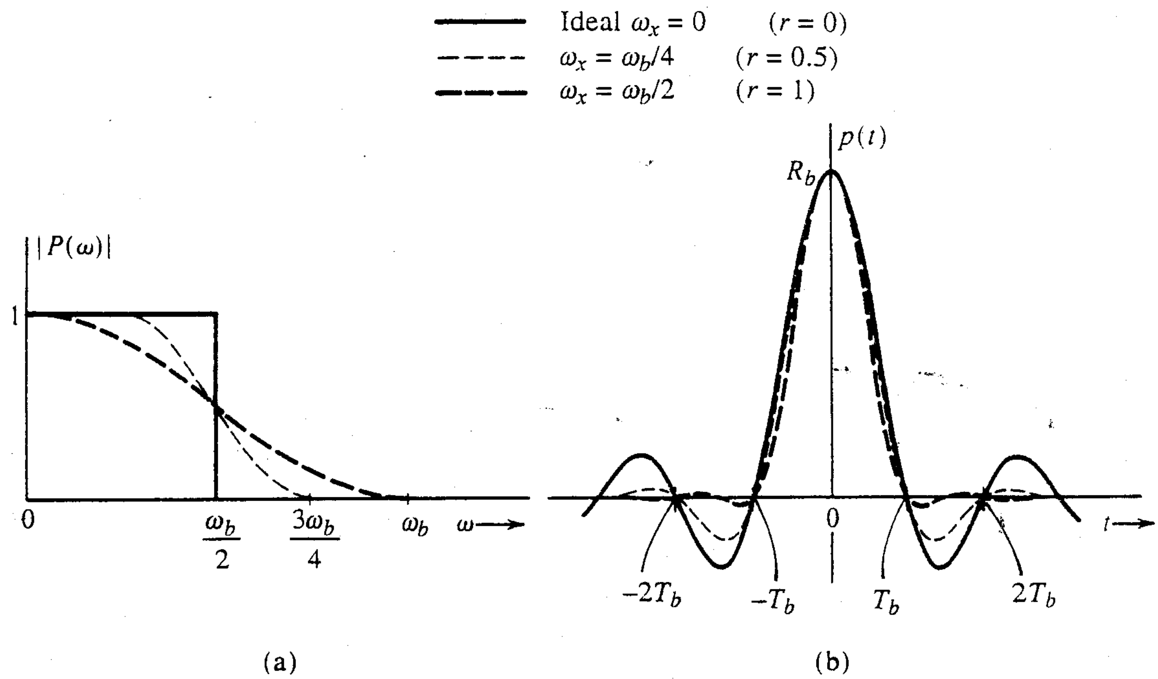


Figure 7.13 Pulses satisfying the Nyquist criterion.

off characteristic. The inverse Fourier transform of this spectrum is readily found as (see Prob. 7.3-7)

$$p(t) = R_b \frac{\cos \pi R_b t}{1 - 4R_b^2 t^2} \operatorname{sinc}(\pi R_b t) \tag{7.35}$$

This pulse is shown in Fig. 7.13b ($r = 1$). We can make several important observations about the raised-cosine pulse. First, the bandwidth of this pulse is R_b Hz and has the value R_b at $t = 0$ and zero not only at all the remaining signaling instants but also at points midway between all the signaling instants. Secondly, it decays rapidly, as $1/t^3$. As a result, the raised-cosine pulse is relatively insensitive to deviations of R_b, sampling rate, timing jitter, and so on. Furthermore, the pulse-generating filter with transfer function $P(\omega)$ [Eq. (7.34b)] is closely realizable. The phase characteristic that goes along with this filter is very nearly linear, so that no additional phase equalization is needed. Lastly, we shall see that the raised-cosine pulse can also be used as a duobinary pulse, which uses the principle of controlled ISI for correct detection.

It should be remembered that it is the pulses received at the detector input that should have the Nyquist form for zero ISI. In practice, because the channel is not ideal (distortionless), the transmitted pulses should be shaped so that after passing through the channel with transfer function $H_c(\omega)$, they will be received in the proper shape (such as raised-cosine pulses) at the receiver. Hence, the transmitted pulse $p_i(t)$ should satisfy

$$P_i(\omega)H_c(\omega) = P(\omega)$$

where $P(\omega)$ is the vestigial spectrum in Eq. (7.29).

MPLE 7.1 Determine the pulse transmission rate in terms of the transmission bandwidth B_T and the roll-off factor r. Assume a scheme using the Nyquist criterion.

In practical systems, the principles outlined in this chapter are used to ensure that random channel noise from thermal effects and intersystem cross-talk will cause errors in a negligible percentage of the received pulses. Switching transients, lightning strikes, power line load switching, and other singular events cause very high-level noise pulses of short duration to contaminate the cable pairs that carry digital signals. These pulses, collectively called **impulse noise**, cannot conveniently be engineered away, and they constitute the most prevalent source of errors from the environment outside the digital systems. Errors are virtually never, therefore, found in isolation, but occur in bursts of up to several hundred at a time. To correct error burst, we use special **burst-error-correcting codes**, described in Chapter 16.

7.7 M-ARY COMMUNICATION

Digital communication uses only a finite number of symbols for communication, the minimum number being two (the binary case). Thus far we have restricted ourselves to only the binary case. We shall now briefly discuss some aspects of M-ary communication (communication using M symbols). This subject will be discussed in depth in Chapters 13 and 14.

We can readily show that the information transmitted by each symbol increases with M. For example, when $M = 4$ (4-ary or quaternary case), we have four basic symbols, or pulses, available for communication (Fig. 7.25a). A sequence of two binary digits can be transmitted by just one 4-ary symbol. This is because a sequence of two binary digits can form only four

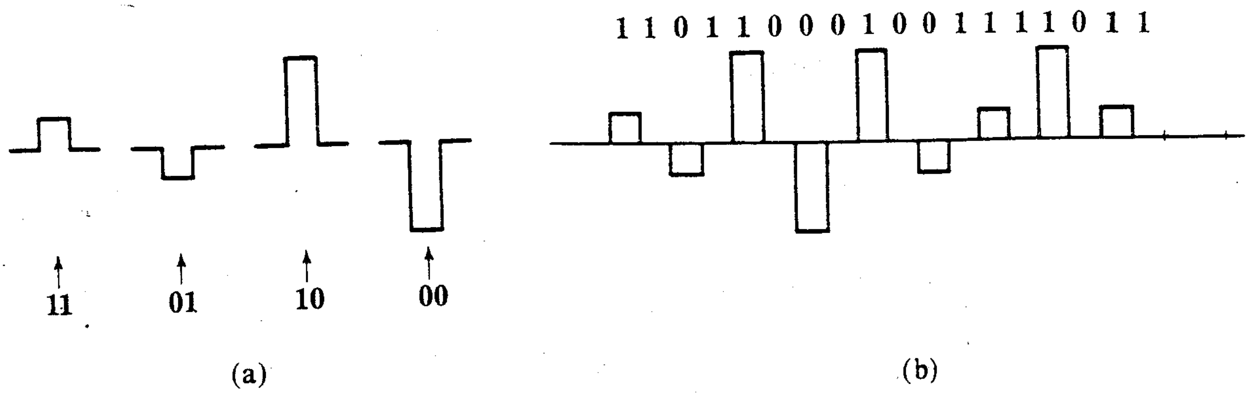


Figure 7.25 4-ary multiamplitude signal.

possible sequences (viz., **11**, **10**, **01**, and **00**). Because we have four distinct symbols available, we can assign one of the four symbols to each of these combinations (Fig. 7.25a). This signaling (**multiamplitude signaling**) allows us to transmit each pair of binary digits by one 4-ary pulse (Fig. 7.25b). Hence, to transmit n binary digits, we need only $(n/2)$ 4-ary pulses. This means one 4-ary symbol can transmit the information of two binary digits. Also, because three binary digits can form $2 \times 2 \times 2 = 8$ combinations, a group of 3 bits can be transmitted by one 8-ary symbol. Similarly, a group of 4 bits can be transmitted by one 16-ary symbol. In general, the information I_M transmitted by an M-ary symbol is

$$I_M = \log_2 M \text{ binary digits, or bits} \tag{7.55}$$

This means we can increase the rate of information transmission by increasing M. But the transmitted power increases as M, because to have the same noise immunity, the minimum separation between pulse amplitudes should be comparable to that of binary pulses. Therefore, pulse amplitudes increase with M (see Fig. 7.25). It will be shown in Chapter 13 that the transmitted power increases as M^2 (see Prob. 7.7.4). Thus, to increase the rate of communication by a factor of $\log_2 M$, the power required increases as M^2. Because the transmission bandwidth depends only on the pulse rate and not on pulse amplitudes, the bandwidth is independent of M. But if we wish to maintain the same rate of data transmission as in the binary case, we can reduce the transmission bandwidth by a factor of $\log_2 M$ at the cost of increased power.

Although most of the terrestrial digital telephone network uses binary encoding, the subscriber loop portion of the integrated services digital network (ISDN) uses the quaternary code 2B1Q, shown in Fig. 7.25a. It uses NRZ pulses to transmit 160 kbit/s of data using a **baud** rate (pulse rate) of 80 kbit/s. Of the various line codes examined by the ANSI standards committee, 2B1Q provided the greatest baud rate reduction in the noisy and cross-talk-prone local cable plant environment.

Pulse Shaping in the Multiamplitude Case: In this case, we can use the Nyquist criterion pulses because these pulses have zero ISI at the pulse centers, and, therefore, their amplitudes can be correctly detected by sampling at the pulse centers. We can also use the controlled ISI (partial response signaling) for the M-ary case.[9]

Figure 7.25 shows just one possible M-ary scheme (multiamplitude signaling). There are infinite possible ways of structuring M waveforms. For example, we may use M orthogonal pulses $\varphi_1(t)$, $\varphi_2(t)$, ..., $\varphi_M(t)$ with the property

Digital Modulation Techniques

Digital modulation is the process that transforms digital symbols into waveforms that are compatible with the characteristics of the channel.

In baseband modulation, these waveforms are pulses, while in bandpass modulation the information signal modulates a sinusoidal carrier wave.

Digital Bandpass Modulation

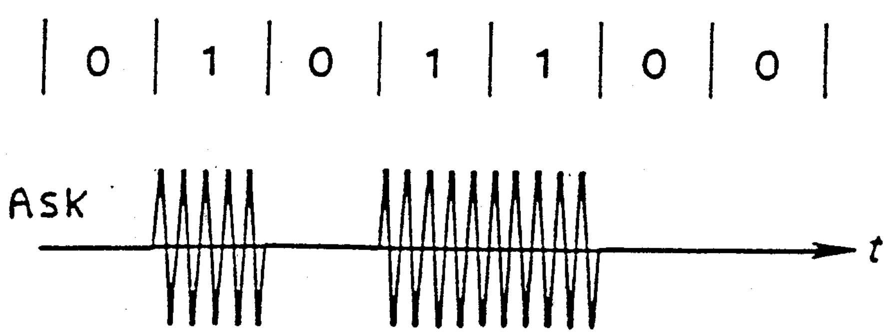
This can be achieved by varying the amplitude, frequency or phase of an RF carrier, or a combination of them in accordance with the information to be transmitted.

$$s(t) = A(t)\cos[\omega_c t + \phi(t)]$$

ASK, FSK and PSK are the most common digital modulation formats. In M-ary signalling schemes, the processor accepts k source bits at a time and instructs the modulator to produce one of an available set of $M = 2^k$ waveform types. For binary modulation, $k = 1$ and hence $M = 2$.

Amplitude Shift Keying (ASK)

This was one of the earliest digital modulation schemes used in wireless telegraphy. It is no longer widely used, but it is useful to study ASK as a model due to its simplicity.



Frequency Shift Keying (FSK)

Very widely used in modems designed for bit rates up to 1200 bps.

It is relatively easy to generate and detect FSK signals.

$$y(t) = \begin{cases} A\cos(\omega_1 t), & \text{for binary 1} \\ A\cos(\omega_2 t), & \text{for binary 0} \end{cases}$$

or

$$y(t) = \begin{cases} A\cos(\omega_c t + \omega_d t), & \text{for binary 1} \\ A\cos(\omega_c t - \omega_d t), & \text{for binary 0} \end{cases}$$

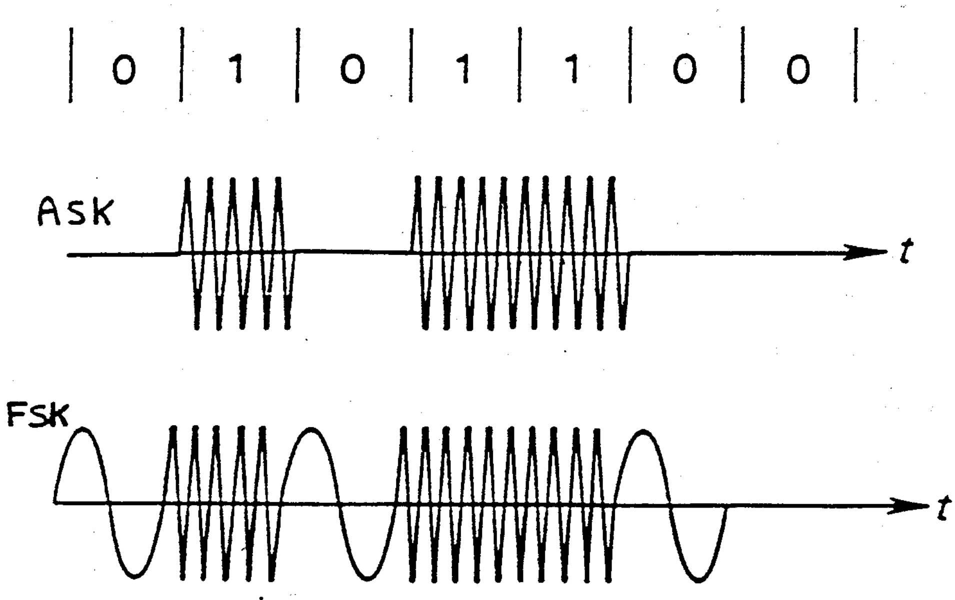


Fig. 28 Frequency shift keying

Example

Full duplex operation of FSK over a two-wire telephone line is achieved by dividing the bandwidth into two halves; one half is used for transmission in one direction and the other half for the reverse direction.

The CCITT frequency assignment for the 300 bits/s V21 recommendation using a two-wire switched circuit is shown below:

binary	forward direction frequency (Hz))	reverse direction frequency (Hz)
0	1180	1850
1	980	1650

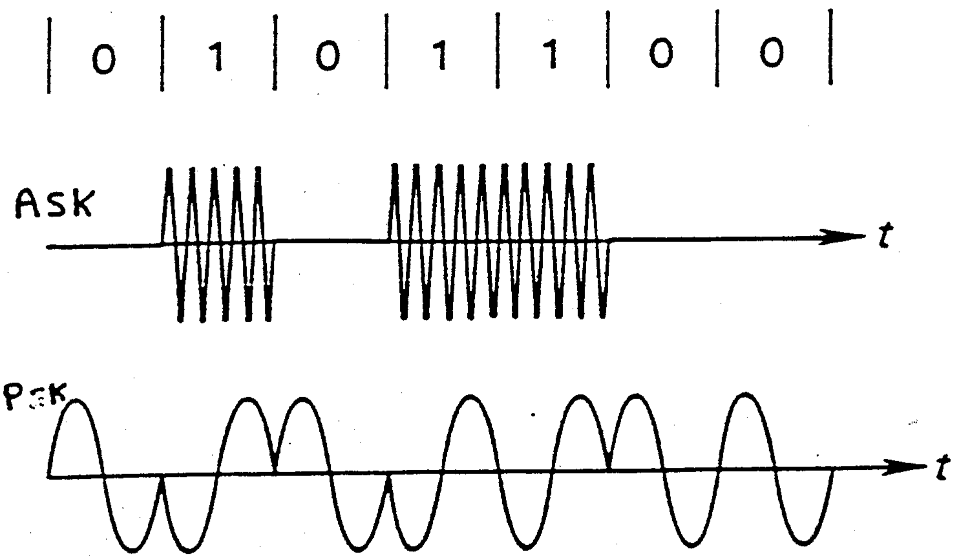
In this application, $f_d = 100$ Hz and the carrier frequency is $f_c = 1080$ Hz in the forward direction and $f_c = 1750$ Hz in the reverse direction.

Phase Shift Keying (PSK)

This has proved to be a very successful form of digital modulation.

A PSK signal can be expressed as

$$y(t) = \begin{cases} A\cos(\omega_c t + \pi) & , \text{ for binary } 1 \\ A\cos(\omega_c t) & , \text{ for binary } 0 \end{cases}$$



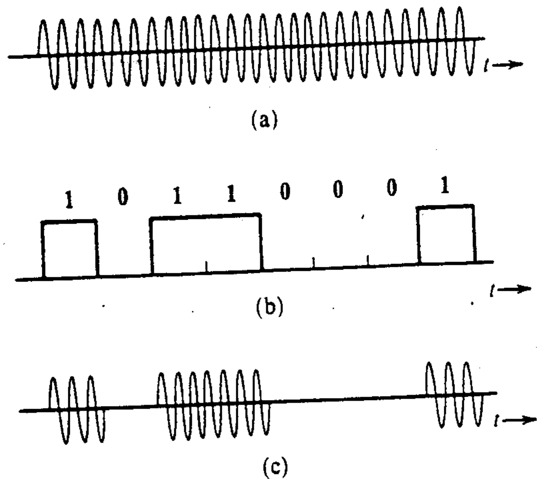


Figure 7.27 (a) Carrier $\cos \omega_c t$. (b) Modulating signal $m(t)$. (c) ASK: modulated signal $m(t) \cos \omega_c t$.

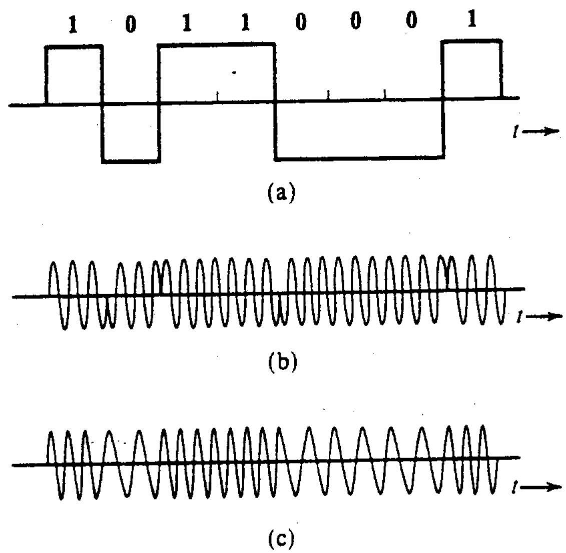


Figure 7.28 (a) Modulating signal $m(t)$. (b) PSK: modulated signal $m(t) \cos \omega_c t$. (c) FSK: modulated signal.

Figure 7.29 PSD of: (a) ASK. (b) PSK. (c) FSK.