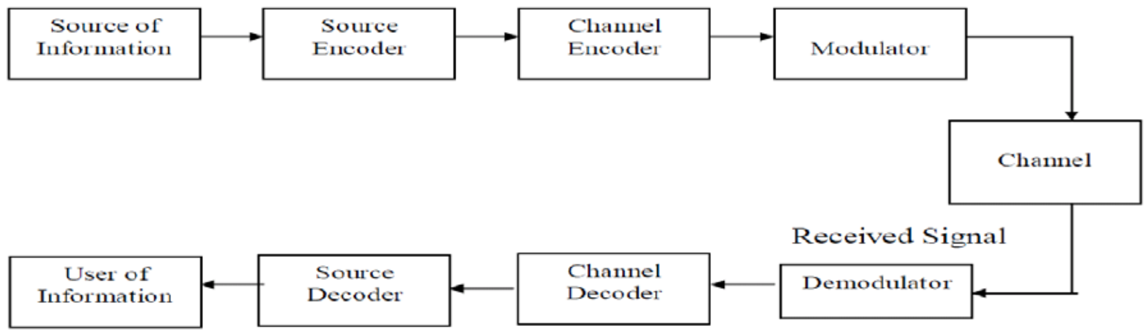
Q.2 a. Draw the block diagram of digital communication system and explain the function of each block. Differentiate the system with analog communication. (8)

Answer:



A block diagram with all the functional units – Discrete information source – source encoder – channel encoder – modulator- channel (+ noise) – demodulator - channel decoder – source decoder – detailed explanation about the functions of each blocks – If source is analog the sampling and quantization. Source encoder for assigning code words – assigning higher probable vents lower code word length for efficient transmission – channel encoder - error detection and error correction. The Modulator converts the input bit stream into an electrical waveform suitable for transmission over the communication channel. The extraction of the message from the information bearing waveform produced by the modulation is accomplished by the demodulator. The output of the demodulator is bit stream. The important parameter is the method of demodulation.

Block diagram of analog communication – Block diagram of digital communication – major differences in functional blocks –

Advantages

1. The effect of distortion, noise and interference is less in a digital communication system. This is because the disturbance must be large enough to change the pulse from one state to the other
2. Regenerative repeaters can be used at fixed distance along the link, to identify and regenerate a pulse before it is degraded to an ambiguous state.

3. Digital circuits are more reliable and cheaper compared to analog circuits.

4. The Hardware implementation is more flexible than analog hardware because of the use of microprocessors, VLSI chips etc.

5. Signal processing functions like encryption, compression can be employed to maintain the secrecy of the information.

6. Error detecting and Error correcting codes improve the system performance by reducing the probability of error.

7. Combining digital signals using TDM is simpler than combining analog signals using FDM. The different types of signals such as data, telephone, TV can be treated as identical signals in transmission and switching in a digital communication system.

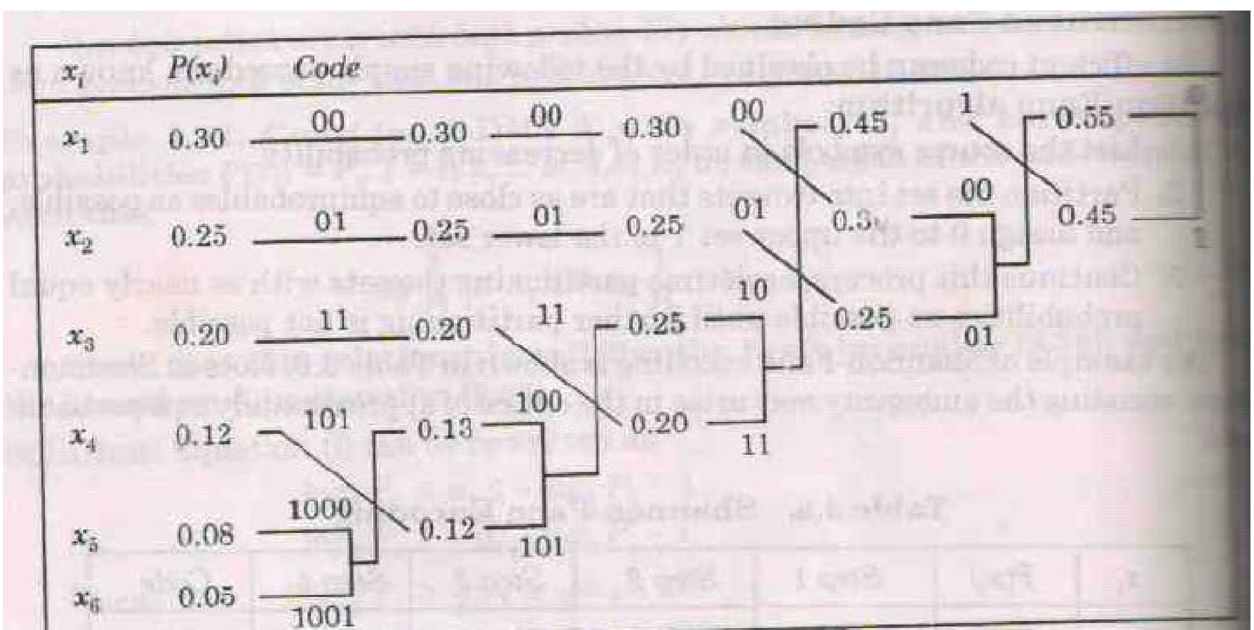
8. We can avoid signal jamming using spread spectrum technique.

b. A discrete memoryless source (DMS) has six symbols X_i and their probability of occurrence $P(X_i)$ as follows:

X_i		$P(X_i)$
X_1	$\longrightarrow$	0.30
X_2	$\longrightarrow$	0.25
X_3	$\longrightarrow$	0.20
X_4	$\longrightarrow$	0.12
X_5	$\longrightarrow$	0.08
X_6	$\longrightarrow$	0.05

Using Huffman coding algorithm, find the Huffman codes for the symbols. Calculate the coding efficiency. (8)

Answer:



$$H(X) = 2.36 \text{ b/symbol}$$
$$L = 2.38 \text{ b/symbol}$$
$$\eta = 0.99$$

Q.3 a. State and explain sampling theorem. (6)

Answer:

Sampling Theorem: A bandlimited signal can be reconstructed exactly if it is sampled at a rate at least twice the maximum frequency component in it."

b. g(t) is a band limited signal with bandwidth f_m Hz and its spectrum is given in Fig (a). Mathematically show that the spectrum of the sampled (at sampling frequency of f_s Hz) version of signal g(t) is as given in Fig (b). (10)

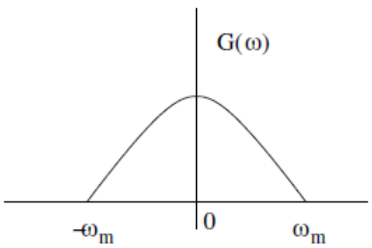


Fig (a)

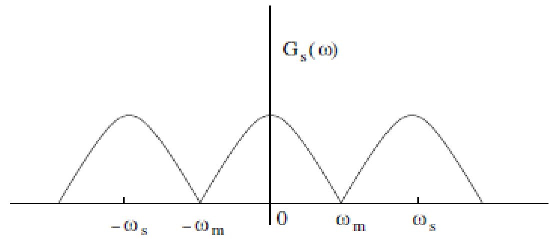
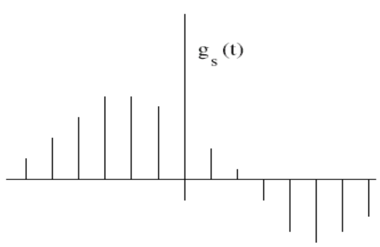
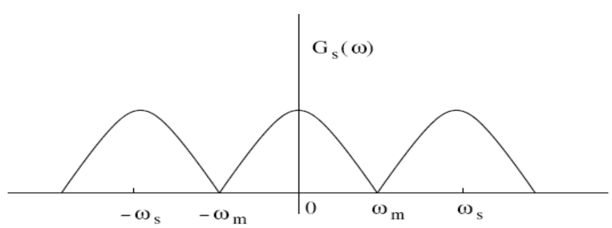


Fig (b)

Answer:

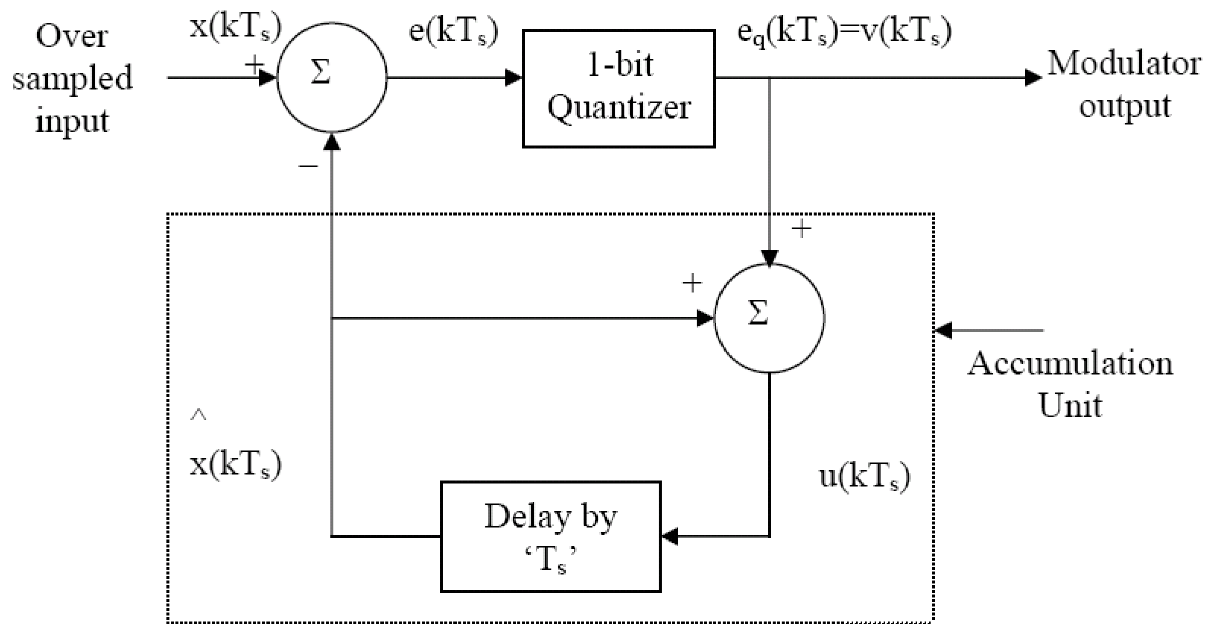
$$
\begin{aligned}
\mathcal{F}(g_s(t)) &= \mathcal{F}\left[g(t)\delta_T(t)\right] \\
&= \mathcal{F}\left[g(t)\sum_{n=-\infty}^{+\infty}\delta(t-nT)\right] \\
&= \frac{1}{2\pi}\left[G(\omega) * \omega_0 \sum_{n=-\infty}^{+\infty}\delta(\omega-n\omega_0)\right] \\
G_s(\omega) &= \frac{1}{T}\sum_{n=-\infty}^{+\infty} G(\omega) * \delta(\omega-n\omega_0) \\
G_s(\omega) &= \mathcal{F}\left[g(t) + 2g(t)\cos(\omega_0 t) + 2g(t)\cos(2\omega_0 t) + \cdots\right] \\
G_s(\omega) &= \frac{1}{T}\sum_{n=-\infty}^{+\infty} G(\omega-n\omega_0)
\end{aligned}
$$

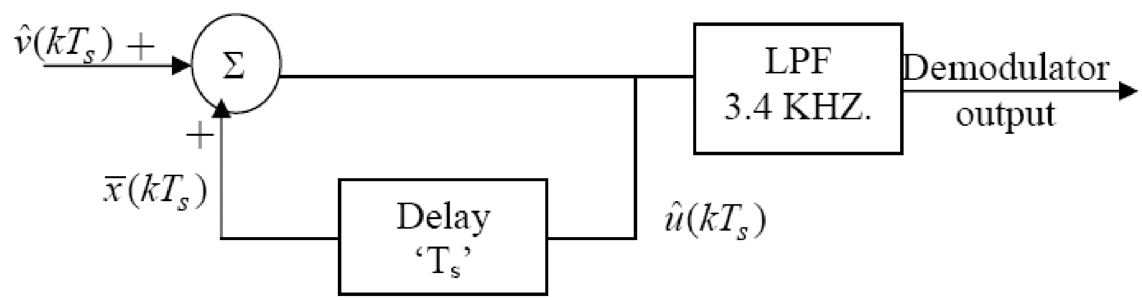




Q.4 a. Explain Delta modulation with transmitter and receiver block diagram. What are the major limitations of Delta modulation? (10)

Answer:

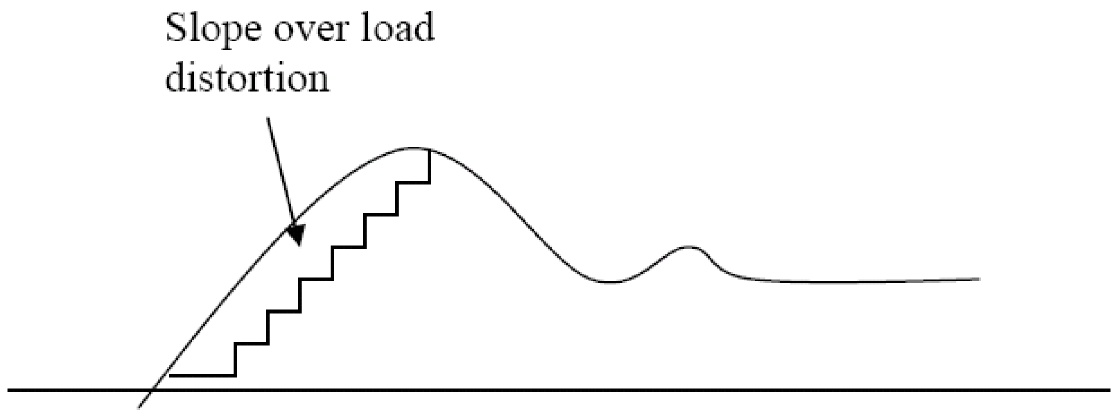




Delta Modulation is a special case of DPCM. In DPCM scheme if the base band signal is sampled at a rate much higher than the Nyquist rate purposely to increase the correlation between adjacent samples of the signal, so as to permit the use of a simple quantizing strategy for constructing the encoded signal, Delta modulation (DM) is precisely such as scheme. Delta Modulation is the one-bit (or two-level) versions of DPCM.

DM provides a staircase approximation to the over sampled version of an input base band signal. The difference between the input and the approximation is quantized into only two levels, namely, $\pm\delta$ corresponding to positive and negative differences, respectively, Thus, if the approximation falls below the signal at any sampling epoch, it is increased by δ. Provided that the signal does not change too rapidly from sample to sample, we find that the stair case approximation remains within $\pm\delta$ of the input signal. The symbol δ denotes the absolute value of the two representation levels of the one-bit quantizer used in the DM.

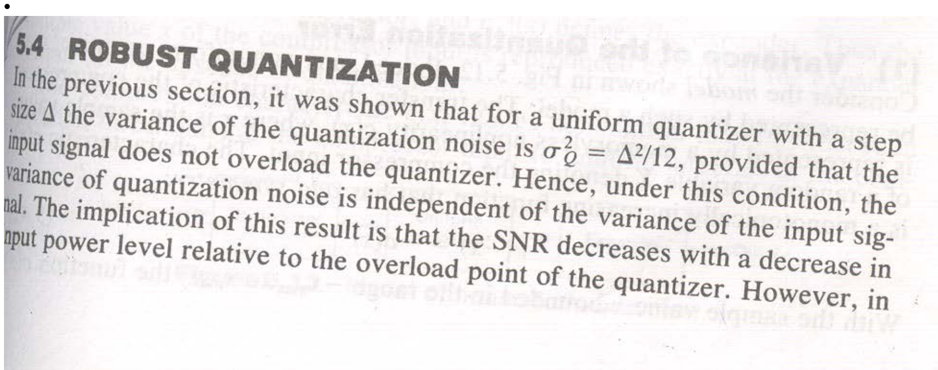
However DM also suffers from a few limitations such as the following: a) Slope over load distortion: If the input signal amplitude changes fast, the step by-step accumulation process may not catch up with the rate of change. This happens initially when the demodulator starts operation from cold-start but is usually of negligible effect for speech.



Granular noise: If the step-size is made arbitrarily large to avoid slope-overload distortion, it may lead to 'granular noise'. Imagine that the input speech signal is fluctuating but very close to zero over limited time duration. This may happen due to pauses between sentences or else. During such moments, our delta modulator is likely to produce a fairly long sequence of 101010…., reflecting that the accumulator output is close but alternating around the input signal. This phenomenon is manifested at the output of the delta demodulator as a small but perceptible noisy background. This is known as 'granular noise'. An expert listener can recognize the crackling sound. This noise should be kept well within a tolerable limit while deciding the step-size. Larger step-size increases the granular noise while smaller step size increases the degree of slope-overload distortion.

b. Explain the concepts of Robust Quantization in detail. (6)

Answer:



5.4 ROBUST QUANTIZATION

In the previous section, it was shown that for a uniform quantizer with a step size Δ the variance of the quantization noise is $\sigma_Q^2 = \Delta^2/12$, provided that the input signal does not overload the quantizer. Hence, under this condition, the variance of quantization noise is independent of the variance of the input signal. The implication of this result is that the SNR decreases with a decrease in input power level relative to the overload point of the quantizer. However, in certain applications, notably in the use of PCM for the transmission of speech signals, the same quantizer has to accommodate input signals with widely varying power levels. For example, the range of voltages covered by speech signals, from the peaks of loud talk to the weak passages of weak talk, is on the order of 1000 to 1. It would therefore be highly desirable from a practical viewpoint for the signal-to-quantization noise ratio to remain essentially constant for a wide range of input power levels. A quantizer that satisfies this requirement is said to be *robust*.

The provision for such a robust performance necessitates the use of a *nonuniform quantizer*, characterized by a step size that increases as the separation from the origin of the transfer characteristic is increased. Accordingly, in the case of speech signals, the weak passages (which generally occur with high probability and therefore require extra protection) are assigned more representation levels and thereby favored at the expense of the loud passages (which occur relatively infrequently). The result is that a nearly uniform percentage precision is achieved through the greater part of the amplitude range of the input signal with a smaller number of representation levels than would be possible by means of a uniform quantizer. Also, the nonuniform quantizer exploits a characteristic of human hearing, namely that large amplitudes mask quantization noise to some extent.

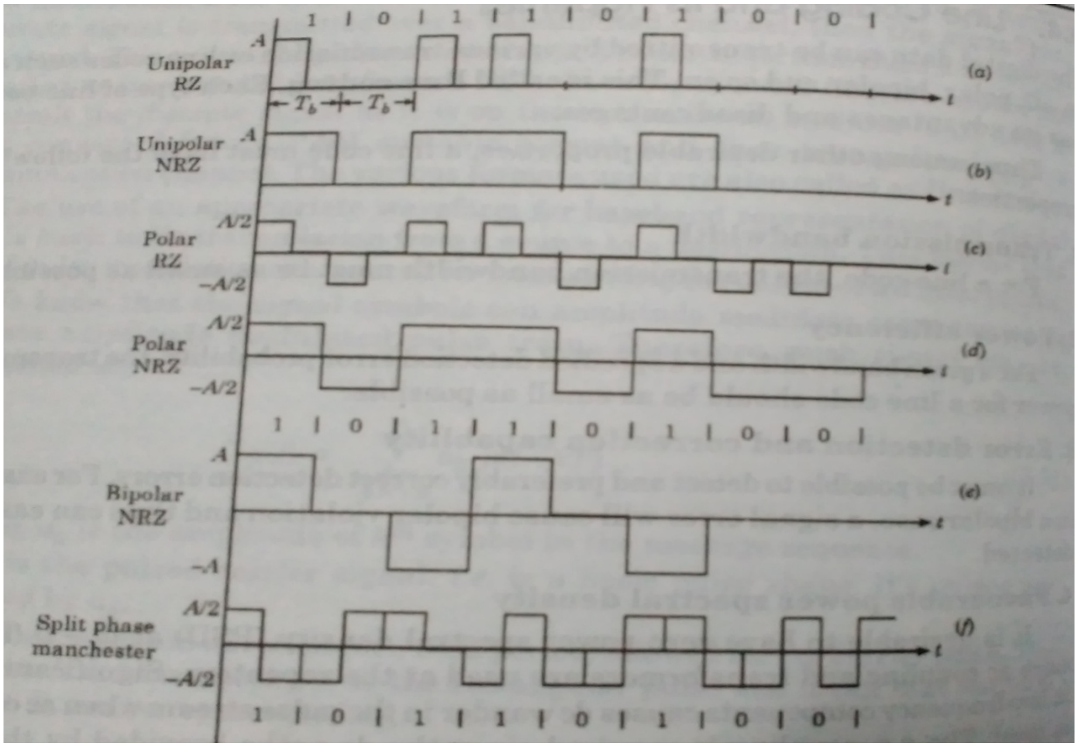
The desired form of nonuniform quantization can be achieved by using a *compressor followed by a uniform quantizer*. By cascading this combination, with an *expander* complementary to the compressor, the original signal samples are restored to their correct values except for quantization errors. Ideally, the compression and expansion laws are exactly the *inverse* of each other, as illustrated in Fig. 5.11. This figure depicts the transfer characteristics of the compressor, uniform quantizer, and expander. In particular, it shows the relationship between the decision thresholds at the compressor input and the representation (reconstruction) levels at the expander output. Thus all sample values of the compressor input, which lie inside an interval $\mathcal{I}_k$ (say), are assigned the discrete value defined by the kth representation level at the expander output.

The combination of a *compressor* and an *expander* is called a *compander*. Naturally, in an actual PCM system, the combination of compressor and uniform quantizer is located in the transmitter, while the expander is located in the receiver.

Q.5 a. Consider a binary sequence 10110100. Draw the waveforms for this sequence, using the following signaling formats. (6)

(i) Unipolar RZ	**(ii) Unipolar NRZ**
(iii) Polar RZ	**(iv) Polar NRZ**
(v) Bipolar NRZ	**(vi) Split Phase Manchester**

Answer:



b. Derive the Nyquist criterion for distortionless Baseband Binary transmission.(10)

Answer:

$$p(t) = \begin{cases} 1 & t = 0 \\ 0 & t = nT_s \ (n \neq 0) \end{cases}$$

$$p_s(t) = \sum_{n=-\infty}^{\infty} p(nT_s)\delta(t - nT_s) = \delta(t)$$

$$P_S(f) = \frac{1}{T_s} \sum_{n=-\infty}^{\infty} P\left(f - \frac{n}{T_s}\right) = 1$$

Therefore $\displaystyle\sum_{n=-\infty}^{\infty} P\left(f - \frac{n}{T_s}\right) = T_s$

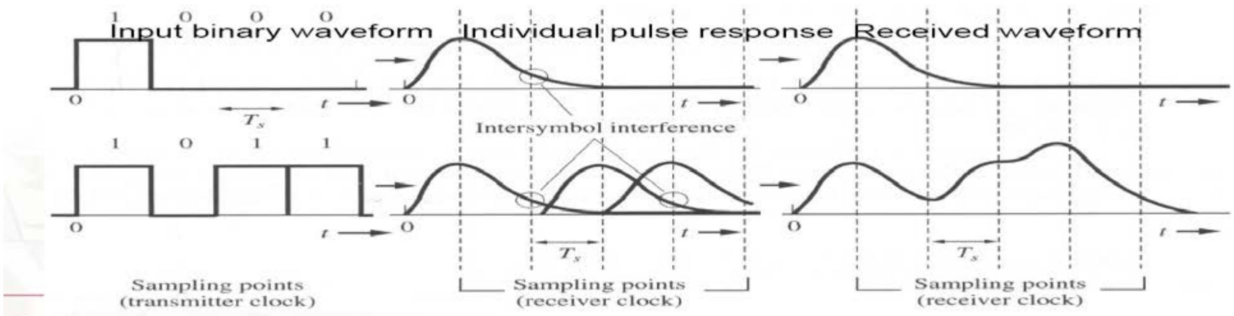
The presence of ISI in the system introduces errors in the decision device at the receiver output. Therefore, in the design of the transmitting and receiving filters, the objective is to minimize the effects of ISI, and thereby deliver the digital data to its destination with the smallest error rate possible.

Q.6 a. What is meant by Inter symbol Interference? What are the effects of Inter symbol Interference? (6)

Answer:

When a signal is band limited in the frequency domain, it is usually smeared in the time domain. This smearing results in inter symbol interference (ISI). The only way to avoid ISI is to satisfy the 1st Nyquist criterion. For an impulse response this means at sampling instants having only one nonzero sample. Rectangular pulses are suitable for infinite-bandwidth channels (practically – wideband). Practical channels are band-limited ->

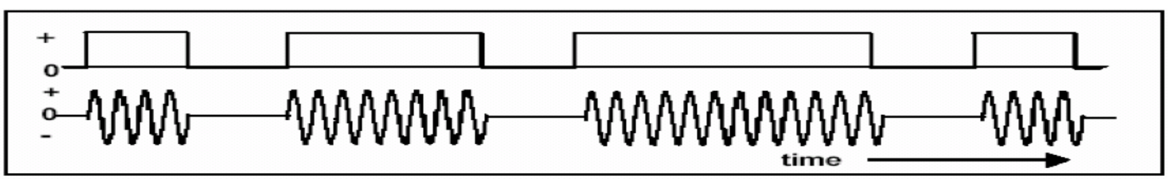
pulses spread in time and are smeared into adjacent slots. This is inter symbol interference (ISI)



b. Explain with block diagram and mathematical equations the generation and coherent demodulation of ASK signal along with its signal space diagram. (10)

Answer:

Amplitude shift keying - ASK - in the context of digital communications is a modulation process, which imparts to a sinusoid two or more discrete amplitude levels. These are related to the number of levels adopted by the digital message.



Binary Amplitude-Shift Keying (BASK)

A *binary amplitude-shift keying* (BASK) signal can be defined by

$$s(t) = A\, m(t) \cos 2\pi f_c t, \qquad 0 \le t \le T \qquad (22.1)$$

where A is a constant, $m(t) = 1$ or 0, f_c is the carrier frequency, and T is the bit duration. It has a power $P = A^2/2$, so that $A = \sqrt{2P}$. Thus equation (22.1) can be written as

$$\begin{aligned} s(t) &= \sqrt{2P} \cos 2\pi f_c t, && 0 \le t \le T \\ &= \sqrt{PT}\sqrt{\frac{2}{T}} \cos 2\pi f_c t, && 0 \le t \le T \\ &= \sqrt{E}\sqrt{\frac{2}{T}} \cos 2\pi f_c t, && 0 \le t \le T \qquad (22.2) \end{aligned}$$

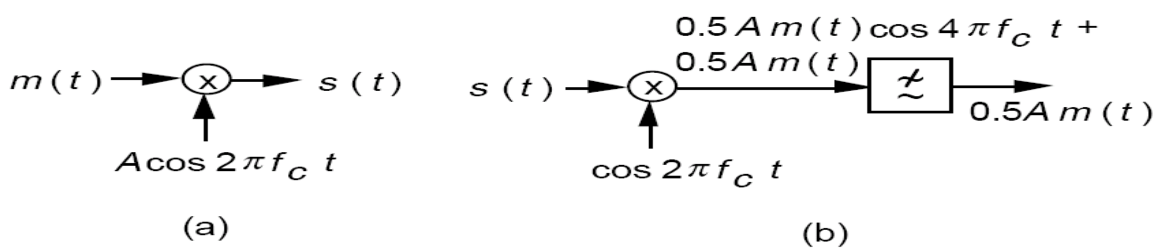


Figure 22.4 (a) BASK modulator and (b) coherent demodulator.

Q.7 a. What is a matched filter? Derive the expression for impulse response of the Matched filter. (8)

Answer:

Maximization of output Signal –to-Noise Raito:

Let, h(t) be the impulse response of a linear filter and $x(t) = \varphi(t) + \omega(t), 0 \le t \le T$: is the input to the filter where $\varphi(t)$ is a known signal and $\omega(t)$ is an additive white noise sample function with zero mean and psd of $(N_0/2)$ Watt/Hz. Let, $\varphi(t)$ be one of the orthonormal basis functions. As the filter is linear, its output can be expressed as, $y(t) = \varphi_0(t) + n(t)$, where $\varphi_0(t)$ is the output due to the signal component $\varphi(t)$ and $n(t)$ is the output due to the noise component $\omega(t)$.[**Fig. 4.20.2**].

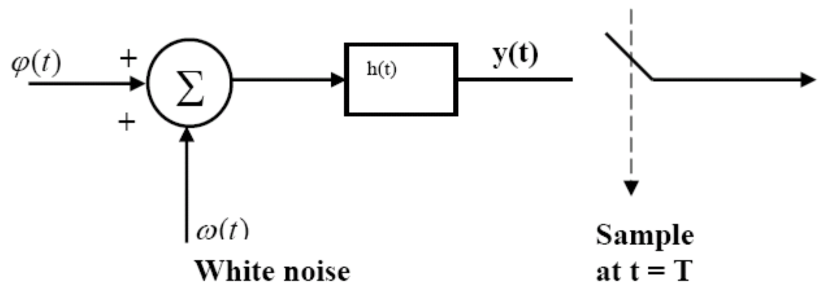


Fig. 4.20.2: *A matched filter is fed with a noisy basis function to which it is matched*

We can now re-frame the requirement of minimum probability of error (or maximum likelihood detection) as: The filter should make power of $\varphi_0(t)$ considerably greater (in fact, as large as possible) compared to the power of n(t) at t = T. That is, the filter should maximize the output signal-to-noise power ratio $[(SNR)_0]$

$$\triangleq \left.\frac{|\varphi_0(T)|^2}{E[n^2(t)]}\right]\max$$

The following discussion shows that the SNR is indeed maximized when h(t) is matched to the known input signal $\varphi(t)$.

Let, $\Phi(f)$: F.T. of known signal $\varphi(t)$

$\mathrm{H}(f)$: Transfer function of the linear filter.

$\therefore \Phi_0(f) = \mathrm{H}(f)\Phi(f)$

and $$\Phi_0(t) = \int_{-\infty}^{\infty} \mathrm{H}(f)\Phi(f)\exp(j2\pi ft)df \qquad 4.20.15$$

The filter output is sampled at t = T. Now,

$$|\varphi_0(T)|^2 = \left|\int_{-\infty}^{\infty} \mathrm{H}(f)\Phi(f)\exp(j2\pi fT)df\right|^2 \qquad 4.20.16$$

Let, $S_N(f)$: Power spectral density of noise at the output of the linear filter. So,

$$S_N(f) = \frac{N_0}{2}.|\mathrm{H}(f)|^2 \qquad 4.20.17$$

Now, the average noise power at the output of the filter

$$= E[n^2(t)] = \int_{-\infty}^{\infty} S_N(f)df$$

$$= \frac{N_0}{2} \int_{-\infty}^{\infty} |\mathrm{H}(f)|^2 \, df \qquad 4.20.18$$

Form Eq. 4.20.16 and 4.20.18, we can write an expression of the output SNR as:

$$(SNR)_0 = \frac{|\varphi^2(T)|^2}{E[n^2(t)]} = \frac{\left| \int_{-\infty}^{\infty} \mathrm{H}(f).\varphi(f) \exp(j2\pi fT) df \right|^2}{\frac{N_0}{2} \int_{-\infty}^{\infty} |\mathrm{H}(f)|^2 \, df} \qquad 4.20.19$$

Our aim now is to find a suitable form of H(f) such that $(SNR)_0$ is maximized. We use Schwarz's inequality for the purpose.

Schwarz's Inequality

Let $\overline{x}(t)$ and $\overline{y}(t)$ denote any pair of complex-valued signals with finite energy, i.e. $\int_{-\infty}^{\infty} |\overline{x}(t)|^2 \, dt < \infty$ & $\int_{-\infty}^{\infty} |\overline{y}(t)|^2 \, dt < \infty$. Schwarz's Inequality states that,

$$\left| \int_{-\infty}^{\infty} \overline{x}(t)\overline{y}(t) dt \right|^2 \leq \int_{-\infty}^{\infty} |\overline{x}(t)|^2 \, dt . \int_{-\infty}^{\infty} |\overline{y}(t)|^2 \, dt. \qquad 4.20.20$$

The equality holds if and only if $\overline{y}(t) = k.\overline{x}^*(t)$, where 'k' is a scalar constant. This implies, $\overline{y}(t)\overline{x}(t) = k.\overline{x}(t)\overline{x^*}(t) \rightarrow$ a real quantity.

Now, applying Schwarz's inequality on the numerator of (Eq.4.20.19), we may write,

$$\left| \int_{-\infty}^{\infty} \mathrm{H}(f)\Phi(f) \exp(j2\pi fT) df \right|^2 \leq \int_{-\infty}^{\infty} |\mathrm{H}(f)|^2 \, df \int_{-\infty}^{\infty} |\Phi(f)|^2 \, df \qquad 4.20.21$$

Using inequality (4.20.21), equation (4.20.19) may be expressed as,

$$(SNR)_0 \leq \frac{2}{N_0} \int_{-\infty}^{\infty} |\varphi(f)|^2 \, df \qquad 4.20.22$$

Now, from Schwarz's inequality, the SNR is maximum i.e. the equality holds, when
$\mathrm{H}_{opt}(f) = \Phi^*(f).\exp(-j2\pi fT)$. [Assuming k = 1, a scalar)

We see, $h_{opt}(t) = \int_{-\infty}^{\infty} \Phi^*(f) \exp[-j2\pi(T-t)] df$ 4.20.23

Now, $\varphi(t)$ is a real valued signal and hence,

$$\Phi^*(f) = \Phi(-f) \qquad 4.20.24$$

Using Eq. 4.20.24 we see,

$$h_{opt}(t) = \int_{-\infty}^{\infty} \Phi(-f)\exp[-j2\pi)(T-t)]df \ = \varphi(T-t)$$

$$\therefore h_{opt}(t) = \Phi(T-t)$$

This relation is the same as we obtained previously for a matched filter receiver. So, we can infer that, *SNR maximization is an operation, which is equivalent to minimization of average symbol error (P_e) for an AWGN Channel.*

b. Explain Gram-Schmidt Orthogonalization Procedure. (8)

Answer:

■ Gram-Schmidt Orthogonalization Procedure

Having demonstrated the elegance of the geometric representation of energy signals, how do we justify it in mathematical terms? The answer lies in the *Gram-Schmidt orthogonalization procedure,* for which we need a *complete orthonormal set of basis functions.* To proceed with the formulation of this procedure, suppose we have a set of M energy signals denoted by $s_1(t), s_2(t), \ldots, s_M(t)$. Starting with $s_1(t)$ chosen from this set arbitrarily, the first basis function is defined by

$$\phi_1(t) = \frac{s_1(t)}{\sqrt{E_1}} \tag{5.19}$$

where E_1 is the energy of the signal $s_1(t)$. Then, clearly, we have

$$\begin{aligned} s_1(t) &= \sqrt{E_1}\phi_1(t) \\ &= s_{11}\phi_1(t) \end{aligned} \tag{5.20}$$

where the coefficient $s_{11} = \sqrt{E_1}$ and $\phi_1(t)$ has unit energy, as required.

Next, using the signal $s_2(t)$, we define the coefficient s_{21} as

$$s_{21} = \int_0^T s_2(t)\phi_1(t)dt \tag{5.21}$$

We may thus introduce a new intermediate function

$$g_2(t) = s_2(t) - s_{21}\phi_1(t) \tag{5.22}$$

which is orthogonal to $\phi_1(t)$ over the interval $0 \leq t \leq T$ by virtue of Equation (5.21) and the fact that the basis function $\phi_1(t)$ has unit energy. Now, we are ready to define the second basis function as

$$\phi_2(t) = \frac{g_2(t)}{\sqrt{\int_0^T g_2^2(t)dt}} \quad (5.23)$$

Substituting Equation (5.22) into (5.23) and simplifying, we get the desired result

$$\phi_2(t) = \frac{s_2(t) - s_{21}\phi_1(t)}{\sqrt{E_2 - s_{21}^2}} \quad (5.24)$$

where E_2 is the energy of the signal $s_2(t)$. It is clear from Equation (5.23) that

$$\int_0^T \phi_2^2(t)dt = 1$$

and from Equation (5.24) that

$$\int_0^T \phi_1(t)\phi_2(t)dt = 0$$

That is to say, $\phi_1(t)$ and $\phi_2(t)$ form an orthonormal pair, as required.

Continuing in this fashion, we may in general define

$$g_i(t) = s_i(t) - \sum_{j=1}^{i-1} s_{ij}\phi_j(t) \quad (5.25)$$

where the coefficients s_{ij} are themselves defined by

$$s_{ij} = \int_0^T s_i(t)\phi_j(t)dt, \qquad j = 1, 2, \ldots, i-1 \quad (5.26)$$

Equation (5.22) is a special case of Equation (5.25) with $i = 2$. Note also that for $i = 1$, the function $g_i(t)$ reduces to $s_i(t)$.

Given the $g_i(t)$, we may now define the set of basis functions

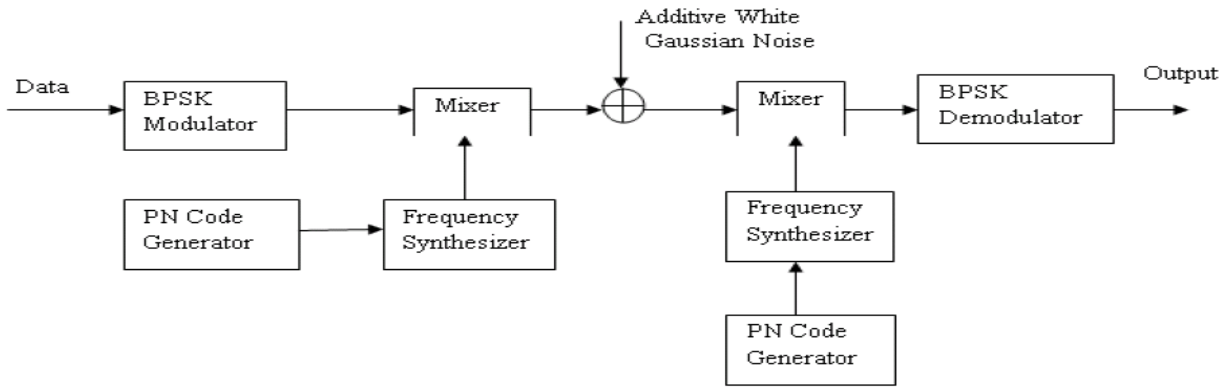
$$\phi_i(t) = \frac{g_i(t)}{\sqrt{\int_0^T g_i^2(t)dt}}, \qquad i = 1, 2, \ldots, N \quad (5.27)$$

which form an orthonormal set. The dimension N is less than or equal to the number of given signals, M, depending on one of two possibilities:

- The signals $s_1(t), s_2(t), \ldots, s_M(t)$ form a *linearly independent set*, in which case $N = M$.
- The signals $s_1(t), s_2(t), \ldots, s_M(t)$ are *not* linearly independent, in which case $N < M$, and the intermediate function $g_i(t)$ is zero for $i > N$.

Q.8 a. Explain with a block diagram the working of Frequency hopped spread Spectrum. (7)

Answer:



In FHSS, the input carrier frequency to the modulator itself varies within a fixed bandwidth. With respect to time, the frequency assigned for modulation is changed with a central frequency but with a fixed bandwidth. As the frequencies change from one to another and the allocation of frequency is pseudorandom, i.e., not in order (dependent upon a PN sequence), the term 'hopping' comes to represent the allocation of frequency with respect to time.

b. What is the difference between Fast frequency hopping and Slow Frequency Hopping? (3)

Answer:

- Slow Frequency Hopping (SFH)

  In this case one or more data bits are transmitted within one hop. An advantage is that coherent data detection is possible. Often, systems using slow hopping also employ (burst) error control coding to restore loss of (multiple) bits in one hop.

- Fast Frequency Hopping (FFH)

  One data bit is divided over multiple hops. In fast hopping, coherent signal detection is difficult, and seldom used. Mostly, FSK or MFSK modulation is used.

c. Calculate the processing gain of FHSS system if the total hopping bandwidth is 2 MHz and instantaneous bandwidth is 40 kHz. (3)

Answer:

Processing gain = Total hopped band width/ Instantaneous band width = 2 MHz/40 KHz

d. Mention any two advantages and disadvantages of FHSS System. (3)

Answer:

Advantage of FHSS:

o Fundamentally much simpler to implement
o Better range, due to lower receiver sensitivity
o Good rejection of in band interference
o Good performance in multipath environments
o No "near/far" problems

Disadvanges Of FHSS:

o Long latency time o Slow Lock-In, must search a channel

o Must resynchronization with other after every hop
o short outdoor range o Lower overall data throughput

Q.9 **Write a short note on applications of :** **(8×2)**
(i) Waveform Coding Techniques
(ii) Spread Spectrum Techniques

(i)

5.8 APPLICATIONS

In this section of the chapter, we describe two related applications: (1) hierarchy of *digital multiplexers*, whereby digitized voice and video signals as well as digital data are combined into one final data stream, and (2) *light wave transmission link* that is well-suited for use in a long-haul telecommunication network.

(1) Digital Multiplexers

In Chapter 4 we introduced the idea of time-division multiplexing whereby a group of analog signals (e.g., voice signals) are sampled sequentially in time at a *common* sampling rate and then multiplexed for transmission over a common line. In this section we consider the multiplexing of digital signals at different bit rates.† This enables us to combine several digital signals, such as computer

* The use of mean opinion score as a subjective measure of quality is discussed in the following references: Daumer (1982) and Jayant and Noll (1985, pp. 10–12). For a comparison of digital speech coders, based on MOS ratings, see Jayant (1986).

† For additional information on digital multiplexers, see Bell Telephone Laboratories (1970, Chapter 26).

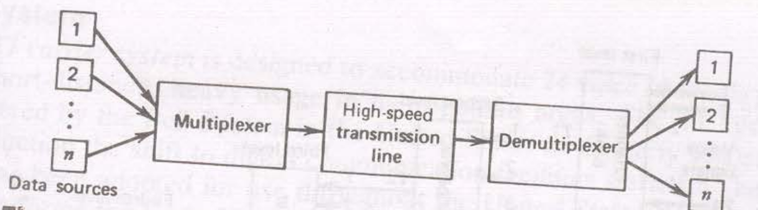


Figure 5.25 Conceptual diagram of multiplexing–demultiplexing.

outputs, digitized voice signals, digitized facsimile and television signals, into a single data stream (at a considerably higher bit rate than any of the inputs). Figure 5.25 shows a conceptual diagram of the digital multiplexing–demultiplexing operation.

The multiplexing of digital signals may be accomplished by using *a bit-by-bit interleaving procedure* with a selector switch that sequentially takes a bit from each incoming line and then applies it to the high-speed common line. At the receiving end of the system the output of this common line is separated out into its low-speed individual components and then delivered to their respective destinations.

Two major groups of digital multiplexers are used in practice:

1. One group of multiplexers is designed to combine relatively low-speed digital signals, up to a maximum rate of 4800 bits per second, into a higher speed multiplexed signal with a rate of up to 9600 bits per second. These multiplexers are used primarily to transmit data over voice-grade channels of a telephone network. Their implementation requires the use of *modems* in order to convert the digital format into an analog format suitable for transmission over telephone channels. The theory of a modem (*mo*dulator–*dem*odulator) is covered in Chapter 7.
2. The second group of multiplexers, designed to operate at much higher bit rates, forms part of the data transmission service generally provided by communication carriers. For example, Fig. 5.26 shows a block diagram of the digital hierarchy based on the T1 carrier, which has been developed by the Bell System. The T1 carrier system, described below, is designed to operate at 1.544 megabits per second, the T2 at 6.312 megabits per second, the T3 at 44.736 megabits per second, and the T4 at 274.176 megabits per second. The system is thus made up of various combinations of lower order T-carrier subsystems designed to accommodate the transmission of voice signals, Picturephone® service, and television signals by using PCM, as well as (direct) digital signals from data terminal equipment.

There are some basic problems involved in the design of a digital multiplexer, irrespective of its grouping:

1. Digital signals cannot be directly interleaved into a format that allows for their eventual separation unless their bit rates are locked to a common clock. Accordingly, provision has to be made for *synchronization* of the incoming digital signals, so that they can be properly interleaved.

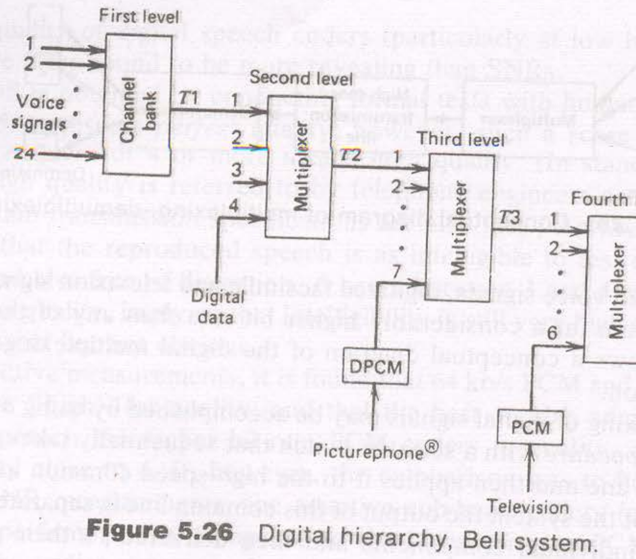


Figure 5.26 Digital hierarchy, Bell system.

2. The multiplexed signal must include some form of *framing*, so that its individual components can be identified at the receiver.
3. The multiplexer has to handle small variations in the bit rates of the incoming digital signals. For example, a 1000-kilometer coaxial cable carrying 3×10^8 pulses per second will have about one million pulses in transit, with each pulse occupying about one meter of the cable. A 0.01 percent variation in the propagation delay, produced by a 1°F decrease in temperature, will result in 100 fewer pulses in the cable. Clearly, these pulses must be absorbed by the multiplexer.

In order to cater for the requirements of synchronization and rate adjustment to accommodate small variations in the input data rates, we may use a technique known as *bit stuffing*. The idea here is to have the outgoing bit rate of the multiplexer slightly higher than the sum of the maximum expected bit rates of the input channels by stuffing in additional non-information carrying pulses. All incoming digital signals are stuffed with a number of bits sufficient to raise each of their bit rates to equal that of a locally generated clock. To accomplish bit stuffing, each incoming digital signal or bit stream is fed into an *elastic store* at the multiplexer. The elastic store is a device that stores a bit stream in such a manner that the stream may be read out at a rate different from the rate at which it is read in. At the demultiplexer, the stuffed bits must obviously be removed from the multiplexed signal. This requires a method that can be used to identify the stuffed bits. To illustrate one such method, and also show one method of providing frame synchronization, we describe the signal format of the Bell System *M12 multiplexer,* which is designed to combine four T1 bit streams into one T2 bit stream. We begin the description by considering the T1 system first and then the M12 multiplexer.

T1 System

The *T1 carrier system* is designed to accommodate 24 voice channels primarily for short-distance, heavy usage in metropolitan areas. The T1 system was pioneered by the Bell System in the United States in the early 1960s; with its introduction the shift to digital communication facilities started.* The T1 system has been adopted for use throughout the United States, Canada, and Japan. It forms the basis for a complete hierarchy of higher order multiplexed systems that are used for either long-distance transmission or transmission in heavily populated urban centers.

A voice signal (male or female) is essentially limited to a band from 300 to 3400 Hz in that frequencies outside this band do not contribute much to articulation efficiency. Indeed, telephone circuits that respond to this range of frequencies give quite satisfactory service. Accordingly, it is customary to pass the voice signal through a low-pass filter with a cutoff frequency of about 3.4 kHz prior to sampling. Hence, with $W = 3.4$ kHz, the nominal value of the Nyquist rate is 6.8 kHz. The filtered voice signal is usually sampled at a slightly higher rate, namely, 8 kHz, which is the *standard* sampling rate in telephone systems.

For companding, the T1 system uses a *piecewise-linear* characteristic (consisting of 15 linear segments) to approximate μ-law companding of Eq. 5.61 with the constant $\mu = 255$. This approximation is constructed in such a way that the segment-end points lie on the compression curve computed from Eq. 5.61, and their projections onto the vertical axis are spaced uniformly. Table 5.3 gives the projections of the segment-end points onto the horizontal axis, and the step sizes of the individual segments. The table is normalized to 8159, so that all values are represented as integer numbers. Segment 0 of the approximation is a colinear segment, passing through the origin; it contains a total of 32 uniform quantizing levels. Linear segments $1a, 2a, \ldots, 7a$ lie above the horizontal axis, whereas linear segments $1b, 2b, \ldots, 7b$ lie below the horizontal axis; each of these 14 segments contains 16 uniform representation

* For a description of the original version of the T1 PCM system, see Fultz and Penick (1965). The description given here is based on an updated version of this system; see Henning and Pan (1972).

Table 5.3 The 15-Segment Companding Characteristic ($\mu = 255$)

Linear segment number	Step size	Projections of segment-end point onto the horizontal axis
0	2	±31
$1a, 1b$	4	±95
$2a, 2b$	8	±223
$3a, 3b$	16	±479
$4a, 4b$	32	±991
$5a, 5b$	64	±2015
$6a, 6b$	128	±4063
$7a, 7b$	256	±8159

levels. For colinear segment 0 the representation levels at the compressor input are $\pm1, \pm3, \ldots, \pm31$, and the corresponding compressor output levels are 0, $\pm1, \ldots, \pm15$. For linear segments $1a$ and $1b$, the representation levels at the compressor input are $\pm35, \pm39, \ldots, \pm95$, and the corresponding compressor output levels are $\pm16, \pm17, \ldots, \pm31$, and so on for the other linear segments.

There are a total of $31 + 14 \times 16 = 255$ output levels associated with the 15-segment companding characteristic described above. To accommodate this number of output levels, each of the 24 voice channels uses a binary code with an 8-bit word. The first bit indicates whether the input voice sample is positive or negative; this bit is a 1 if positive and a 0 if negative. The next three bits of the code word identify the particular segment inside which the amplitude of the input voice sample lies, and the last four bits identify the actual quantizing step inside that segment.

With a sampling rate of 8 kHz, each frame of the multiplexed signal occupies a period of 125 μs. In particular, it consists of twenty-four 8-bit words, plus a single bit that is added at the end of the frame for the purpose of synchronization. Hence, each frame consists of a total of $24 \times 8 + 1 = 193$ bits. Correspondingly, the duration of each bit equals 0.647 μs, and the corresponding bit rate is 1.544 megabits per second.

In addition to the voice signal, a telephone system must also pass special supervisory signals to the far end. This *signaling information* is needed to transmit dial pulses, as well as telephone off-hook/on-hook signals. In the T1 system this requirement is accomplished as follows. Every sixth frame, the least significant (that is, the eighth) bit of each voice channel is deleted and a *signaling bit* is inserted in its place, thereby yielding on average $7\frac{5}{6}$-bit operation for each voice input. The sequence of signaling bits is thus transmitted at a rate equal to the sampling rate divided by six, that is, 1.333 kilobits per second.

For two reasons, namely, the assignment of the eighth digit in every sixth frame to signaling, and the need for two signaling paths for some switching systems, it is necessary to identify a super frame of 12 frames in which the sixth and twelfth frames contain two signaling paths. To accomplish this identification and still allow for rapid synchronization of the receiver framing circuitry, the frames are divided into *odd* and *even* frames. In the odd-numbered frames, the 193rd digit is made to alternate between 0 and 1. Accordingly, the framing circuit searches for the pattern 101010101010 . . . to establish frame synchronization. In the even-numbered frames, the 193rd digit is made to follow the pattern 000111000111 . . . . This makes it possible for the receiver to identify the sixth and twelfth frames as those that follow a 01 transition or 10 transition of this digit, respectively. Figure 5.27 depicts the signaling format of the T1 system.

M12 Multiplexer

Figure 5.28 illustrates the signal format of the M12 multiplexer. Each frame is subdivided into four subframes. The first subframe (first line in Fig. 5.28) is transmitted, then the second, the third, and the fourth, in that order.

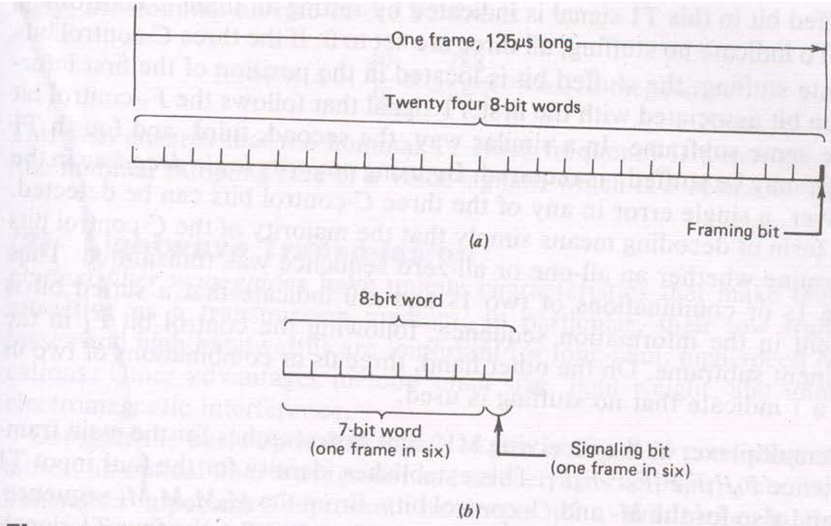


Figure 5.27 T1 bit stream format. (*a*) Coarse structure of a frame. (*b*) Frame structure of a word.

Bit-by-bit interleaving of the incoming four T1 bit streams is used to accumulate a total of 48 bits, 12 from each input. A *control bit* is then inserted by the multiplexer. Each frame contains a total of 24 control bits, separated by sequences of 48 data bits. Three types of control bits are used in the M12 multiplexer to provide synchronization and frame indication, and to identify which of the four input signals has been stuffed. These control bits are labeled as F, M, and C in Fig. 5.28. Their functions are as follows:

1. The F-control bits, two per subframe, constitute the *main* framing pulses. The subscripts on the F-control bits denote the actual bit (0 or 1) transmitted. Thus the main framing sequence is $F_0F_1F_0F_1F_0F_1F_0F_1$ or 01010101.
2. The M-control bits, one per subframe, form *secondary* framing pulses to identify the four subframes. Here again the subscripts on the M-control bits denote the actual bit (0 or 1) transmitted. Thus the secondary framing sequence is $M_0M_1M_1M_1$ or 0111.
3. The C-control bits, three per subframe are *stuffing indicators*. In particular, C_I refers to input channel I, C_{II} refers to input channel II, and so forth. For example, the three C-control bits in the first subframe following M_0 in the first subframe are stuffing indicators for the first T1 signal. The insertion of

M_0	[48]	C_I	[48]	F_0	[48]	C_I	[48]	C_I	[48]	F_1	[48]
M_1	[48]	C_{II}	[48]	F_0	[48]	C_{II}	[48]	C_{II}	[48]	F_1	[48]
M_1	[48]	C_{III}	[48]	F_0	[48]	C_{III}	[48]	C_{III}	[48]	F_1	[48]
M_1	[48]	C_{IV}	[48]	F_0	[48]	C_{IV}	[48]	C_{IV}	[48]	F_1	[48]

Figure 5.28 Signal format of Bell system M12 multiplexer.

a stuffed bit in this T1 signal is indicated by setting all three C-control bits to 1. To indicate no stuffing, all three are set to 0. If the three C-control bits indicate stuffing, the stuffed bit is located in the position of the first information bit associated with the first T1 signal that follows the F_1-control bit in the same subframe. In a similar way, the second, third, and fourth T1 signals may be stuffed, as required. By using *majority logic decoding* in the receiver, a single error in any of the three C-control bits can be detected. This form of decoding means simply that the majority of the C-control bits determine whether an all-one or all-zero sequence was transmitted. Thus three 1s or combinations of two 1s and a 0 indicate that a suffed bit is present in the information sequence, following the control bit F_1 in the pertinent subframe. On the other hand, three 0s or combinations of two 0s and a 1 indicate that no stuffing is used.

The demultiplexer at the receiving M12 unit first searches for the main framing sequence $F_0F_1F_0F_1F_0F_1F_0F_1$. This establishes identity for the four input T1 signals and also for the M- and C- control bits. From the $M_0M_1M_1M_1$ sequence, the correct framing of the C-control bits is verified. Finally, the four T1 signals are properly demultiplexed and destuffed.

The signal format just described has two safeguards:

1. It is possible, although unlikely, that with just the $F_0F_1F_0F_1F_0F_1F_0F_1$ sequence, one of the incoming T1 signals may contain a similar sequence. This could then cause the receiver to lock onto the wrong sequence. The presence of the $M_0M_1M_1M_1$ sequence provides verification of the genuine $F_0F_1F_0F_1F_0F_1F_0F_1$ sequence, thereby ensuring that the four T1 signals are properly demultiplexed.
2. The single-error correction capability built into the C-control bits ensures that the four T1 signals are properly destuffed.

The capacity of the M12 multiplexer to accommodate small variations in the input data rates can be calculated from the format of Fig. 5.28. In each M frame, defined as the interval containing one cycle of $M_0M_1M_1M_1$ bits, one bit can be stuffed into each of four input T1 signals. Each such signal has

$$12 \times 6 \times 4 = 288 \text{ positions in each } M \text{ frame}$$

Also the T1 signal has a bit rate equal to 1.544 megabits per second. Hence, each input can be incremented by

$$1.544 \times 10^3 \times \frac{1}{288} = 5.4 \text{ kilobits/s}$$

This result is much larger than the expected change in the bit rate of the incoming T1 signal. It follows therefore that the use of only one stuffed bit per input channel in each frame is sufficient to accommodate expected variations in the input signal rate.

The local clock that determines the outgoing bit rate also determines the nominal *stuffing rate S*, defined as the average number of bits stuffed per channel in any frame. The M12 multiplexer is designed for $S = 1/3$. Accordingly, the nominal bit rate of the T2 line is

$$1.544 \times 4 \times \frac{49}{48} \times \frac{288}{288\text{-}S} = 6.312 \text{ megabits/s}$$

This also ensures that the nominal T2 clock frequency is a multiple of 8 kHz (the nominal sampling rate of a voice signal), which is a desirable feature.

(2) Lightwave Transmission

Optical fiber waveguides have unique characteristics that make them highly attractive as a transmission medium. In particular, their low transmission losses and high bandwidths are important for long-haul, high-speed communications. Other advantages include small size, light weight, and immunity to electromagnetic interference.

Consider the basic optical fiber link shown in Fig. 5.29 consisting of a transmitter, an optical fiber waveguide, and a receiver. The binary data fed into the transmitter input may represent the output of a multiplexer in the digital heirarchy of Fig. 5.26. In any event, the transmitter emits pulses of optical *power*, with each pulse being "on" or "off" in accordance with the input data. Note that power is a baseband quantity that varies at the data (i.e., modulation) rate and *not* the optical frequency. For the *light source*, we may use a laser injection diode or a semiconductor light emitting diode (LED). In a system design, the choice of the light source determines the optical signal power available for transmission. The *driver* for the light source, typically, consists of a high-current–low-voltage device. The light source is thus turned on and off by switching the drive current on and off in a corresponding manner.

The on–off light pulses produced by the transmitter are launched into the optical fiber waveguide. The *collector efficiency* of the fiber depends on its core diameter and acceptance angle. We thus have to account for a *source-to-fiber coupling loss* that varies over a wide range, depending on the particular combination of light source and optical fiber selected. During the course of propagation along the fiber, a light pulse suffers an additional loss or attenuation that increases exponentially with distance; we refer to this as *fiber loss*. Another important phenomenon that occurs during propagation is *dispersion*, which causes light originally concentrated into a short pulse to spread out into a broader pulse as it propagates along the optical fiber waveguide.

At the receiver, the original input data are *regenerated* by performing three basic operations in the following order:

1. *Detection*, whereby light pulses impinging on the receiver input are converted back into pulses of electrical current.
2. *Pulse shaping and timing* that involves amplification, filtering, and equalization of the electrical pulses, as well as the extraction of timing information.
3. *Decision making*, according to which a particular received pulse is declared "on" or "off."

Typically, the detector consists of a photodiode that responds to light. The

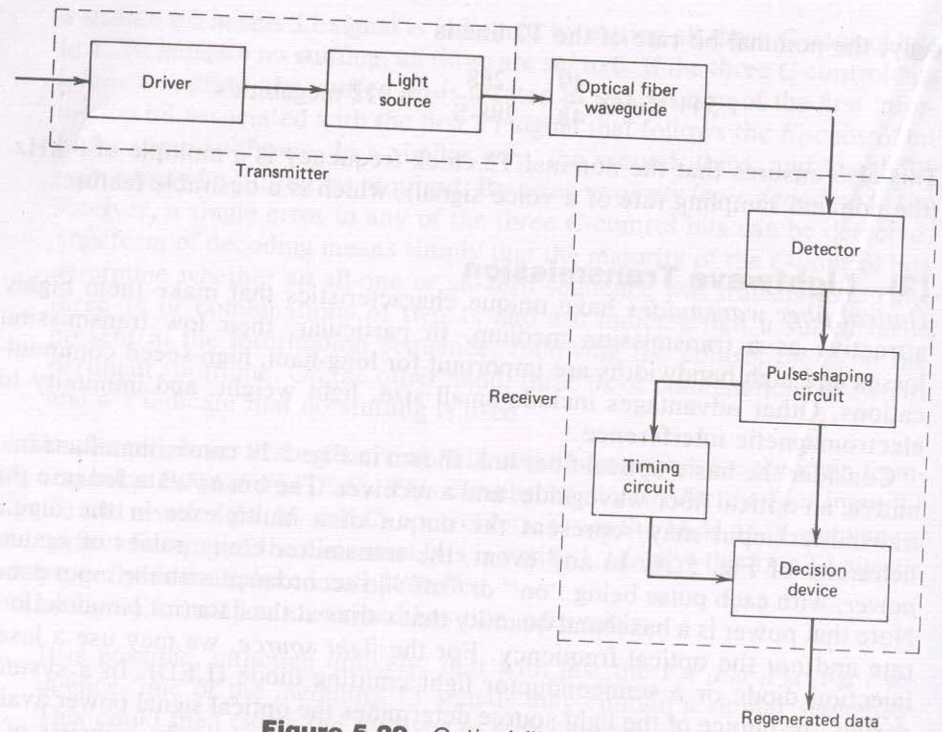


Figure 5.29 Optical fiber link.

choice of the detector and its associated circuitry determines the *receiver sensitivity*. It is also important to recognize that since the optical power is modulated at the transmitter, and since the optical fiber waveguide operates linearly on the propagation power, the detector behaves as a *linear device that converts power to current*.

From this discussion, we see that a lightwave transmission link differs from its coaxial cable counterpart in that power, rather than current (both baseband quantities), propagates through the optical fiber waveguide; otherwise, their individual block diagrams look very much alike. It differs from its microwave counterpart in that a lightwave receiver employs direct detection of the incoming signal rather than first performing a heterodyne downconversion.

In the design of a lightwave transmission link, two separate factors have to be considered: *transmission bandwidth* and *signal losses*. The transmission bandwidth of an optical fiber is determined by the dispersion phenomenon. This, in turn, limits the feasible data rate or the rate at which light pulses can propagate through the optical fiber waveguide. The signal losses are contributed by source-to-fiber coupling loss, fiber loss, fiber-to-fiber loss (due to joining the fiber by a permanent splice or the use of a demountable connector), and fiber-to-detector coupling loss. The fiber loss is insensitive to the input data rate, but it varies with wavelength. In any case, knowing these losses, and knowing the optical power available from the light source, we can determine the power available at the detector.

(ii)

9.7 APPLICATIONS

Probably the single most important application of spread-spectrum techniques is that of protection against jammers. This issue has been discussed at length in previous sections. In this section of the chapter, we briefly consider two other applications of spread-spectrum techniques, namely, *code-division multiple access*, and *multipath suppression*.

(1) Code-division Multiple Access

As mentioned previously in Section 7.13(3), the two most common multiple access techniques for satellite communications are frequency-division multiple access (FDMA) and time-division multiple access (TDMA). In FDMA, all users access the satellite channel by transmitting simultaneously but using disjoint frequency bands. In TDMA, all users occupy the same RF bandwidth of the satellite channel, but they transmit sequentially in time. When, however, all users are permitted to transmit simultaneously and also occupy the same RF bandwidth of the satellite channel, then some other method must be provided for separating the individual signals at the receiver. *Code-division multiple access* (CDMA) is the method that makes it possible to perform this separation.

To accomplish CDMA, spread spectrum is always used.* In particular, each user is assigned a code of its own, which performs the direct-sequence or frequency-hop spread-spectrum modulation. The design of the codes has to cater for two provisions:

1. Each code is approximately *orthogonal* (i.e., has low cross-correlation) with all the other codes.
2. The CDMA system operates *asynchronously*, which means that the transition times of a user's data symbols do not have to coincide with those of the other users.

The second requirement complicates the design of good codes for CDMA.†

The use of CDMA offers three attractive features over TDMA:

1. CDMA does not require an external synchronization network, which is an essential feature of TDMA.
2. CDMA offers a gradual degradation in performance as the number of users is increased. It is therefore relatively easy to add new users to the system.
3. CDMA offers an external interference rejection capability (e.g., multipath rejection or resistance to deliberate jamming).

(2) Multipath Suppression

In many radio channels, the transmitted signal reaches the receiver input via more than one path. For example, in a *mobile communication* environment, the transmitted signal is reflected off a variety of *scatterers* such as buildings, trees, and moving vehicles. Thus, in addition to the *direct path* from the transmitter to the receiver, there are several other *indirect paths* (arising from the presence of the scatterers) that contribute to the composition of the received signal.

* For analytic considerations, the evaluation of energy and bandwidth in CDMA, selective calling and identification requirements, see Cooper and McGillen (1986, pp. 378–395).

† The optimum design of good codes for CDMA systems is covered in Sawarte and Pursley (1980).

Naturally, the contributions from these indirect paths exhibit different signal attenuations and time delays relative to that from the direct path. Indeed, they may interfere with the contribution from the direct path either constructively or destructively at the receiver input. The interference caused by these indirect paths is called *multipath interference* or simply *multipath*. The variation in received signal amplitude due to this interference is called *fading*, as the signal amplitude tends to fade away when destructive interference occurs between the contributions from the direct and indirect paths. The description of multipath fading is also complicated by whether the mobile receiving unit and nearby scatterers are all standing still, whether the mobile receiving unit is standing still but some of the scatterers are moving, or whether the mobile receiving unit is moving as well as some (or all) of the scatterers.*

In a slow-fading channel, we may combat the effects of multipath by applying spread spectrum.† Specifically, in a direct-sequence spread-spectrum system, we find that if the reflected signals at the receiver input are delayed (compared with the direct-path signal) by more than one chip duration of the PN code, then the reflected signals are treated by the matched filter or correlator of the receiver in the same way as any other uncorrelated input signal. Indeed, the higher the chip rate of the PN code, the smaller will the degradation due to multipath be.

In a frequency-hop spread-spectrum system, improvement in system performance in the presence of multipath is again possible, but through a mechanism different from that in a direct-sequence spread-spectrum system. In particular, the effect of multipath is diminished, provided that the carrier frequency of the transmitted signal hops fast enough relative to the differential time delay between the desired signal from the direct path and the undesired signals from the indirect paths. Under this condition, all (or most) of the multipath energy will (on the average) fall in frequency slots that are orthogonal to the slot occupied currently by the desired signal, and degradation due to multipath is thereby minimized.

Text

Digital Communications, Wiley Student Edition, Simon Haykin