UNIT III

DIGITAL MODULATION AND TRANSMISSION

Phase shift keying – BPSK, DPSK, QPSK – Principles of M-ary signaling M-ary PSK & QAM – Comparison, ISI – Pulse shaping – Duo binary encoding – Cosine filters – Eye pattern, equalizer

Phase Shift Keying (PSK) is the digital modulation technique in which the phase of the carrier signal is changed by varying the sine and cosine inputs at a particular time. PSK technique is widely used for wireless LANs, bio-metric, contactless operations, along with RFID and Bluetooth communications.

PSK is of two types, depending upon the phases the signal gets shifted. They are −

1.Binary Phase Shift Keying (BPSK)

This is also called as 2-phase PSK or Phase Reversal Keying. In this technique, the sine wave carrier takes two phase reversals such as 0° and 180°.

BPSK is basically a Double Side Band Suppressed Carrier (DSBSC) modulation scheme, for message being the digital information.

2.Quadrature Phase Shift Keying (QPSK)

This is the phase shift keying technique, in which the sine wave carrier takes four phase reversals such as 0°, 90°, 180°, and 270°.

If this kind of techniques are further extended, PSK can be done by eight or sixteen values also, depending upon the requirement.

BINARY PHASE SHIFT KEYING (BPSK)

1.(a).Demonstrate the modulation & demodulation of BPSK signal with necessary block diagram & waveform.

Or

(b).Explain in detail about BPSK transmitter and receiver and also draw the phasor and constellation diagram.

In BPSK, two phases ($2^1 = 2$) are possible for the carrier. One phase represents a logic 1, and the other phase represents a logic 0.

As the input digital signal changes state (i.e., from a 1 to a 0 or from a 0 to a 1), the phase of the output carrier shifts between two angles that are separated by 180°. Hence, other names for BPSK are phase reversal keying (PRK) and biphase modulation.

BPSK transmitter

The block diagram of BPSK transmitter is shown in the below figure. The balanced modulator acts as a phase reversing switch. Depending on the logic condition of the digital input, the carrier is transferred to the output either in phase or 180° out of phase with the reference carrier oscillator

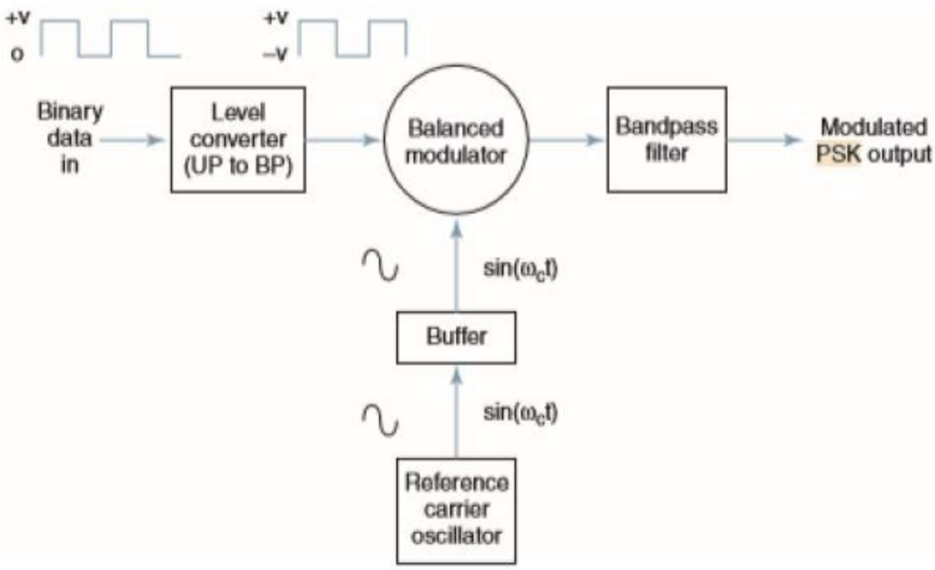


Fig. BPSK Transmitter

Balanced Ring Modulator:

The schematic diagram of a balanced ring modulator is shown in fig

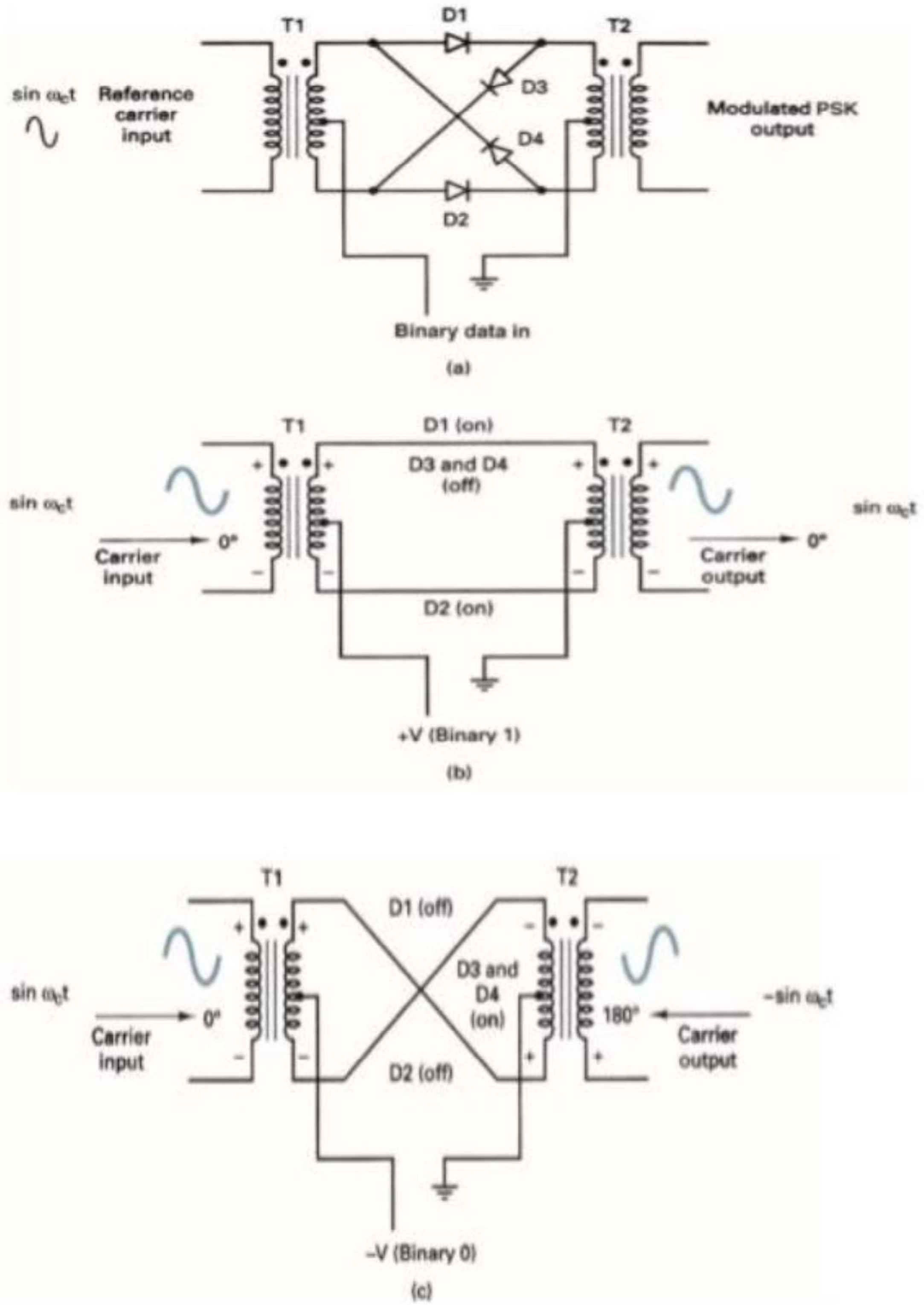


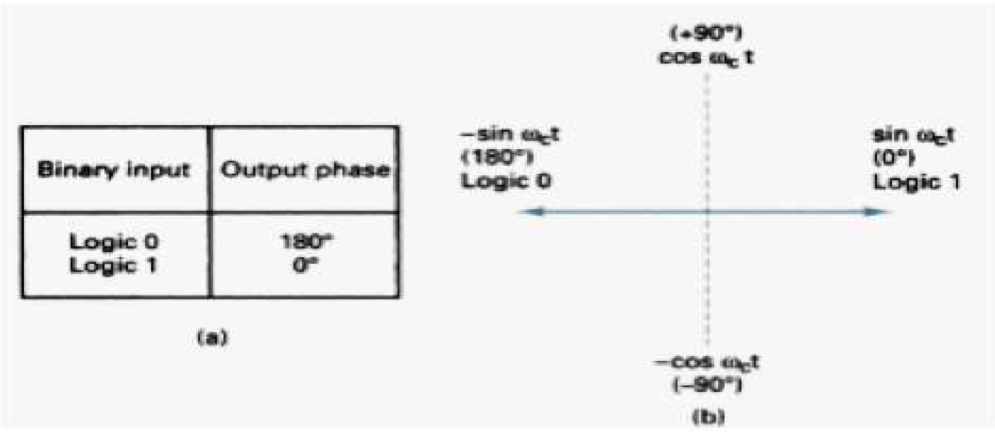
Fig. (a) Balanced Ring Modulator; (b) logic 1 input; (c) logic 0 input

The balanced modulator has two inputs: a carrier that is in phase with the reference oscillator and the binary digital data. To operate properly, the digital input voltage must be much greater than the peak carrier voltage. If the binary input is a logic 1 (positive voltage), diodes D1 and D2 are forward biased and on, while diodes D3 and D4 are reverse biased and off.

The carrier voltage is developed across transformer T2 in phase with the carrier voltage across T1. Consequently, the output signal is in phase with the reference carrier. If the binary input is a logic 0 (negative voltage), diodes D1 and D2 are reverse biased and off, while diodes D3 and D4 are forward biased and on.

Consequently, the carrier voltage is developed across transformer T2 is 180° out of phase with the carrier voltage across T1. Hence, the output signal is 180°out of phase with the reference carrier.

A constellation diagram, which is sometimes called a signal state-space diagram, is similar to a phasor diagram except that the entire phasor is not drawn. In a constellation diagram, only the relative positions of the peaks of the phasors are shown in the given figure.



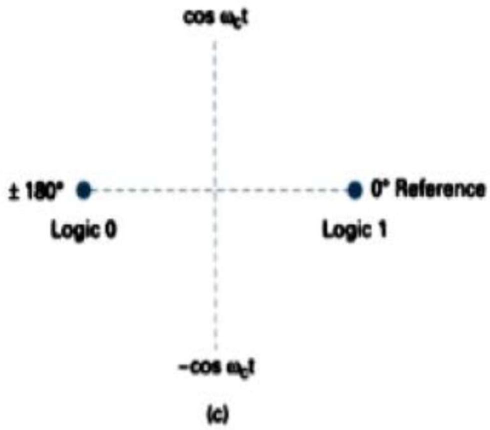


Fig. BPSK modulator: (a) Truth table; (b) Phasor diagram; (c) Constellation diagram

Bandwidth Considerations of BPSK:

In BPSK, the output rate of change (baud) is equal to the input rate of change (bps).

Bit rate = Baud

The bandwidth for BPSK is given as, B=2fa ; fa=f_b/2 =2f_b/2=f_b

BPSK Waveform

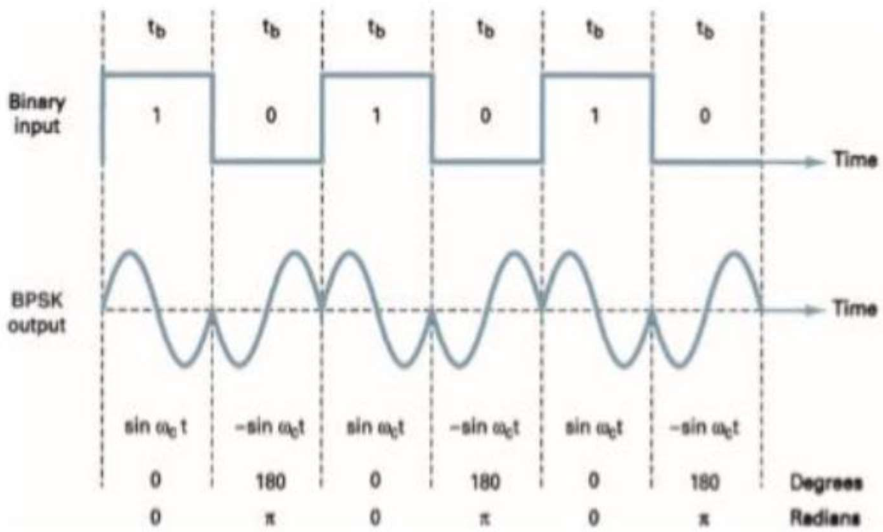


Fig. Output phase-versus-time relationship for a BPSK modulator

BPSK Receiver:

The input signal may be +sin $\omega_c t$ or - sin $\omega_c t$. The signal is detected by coherent carrier recovery circuit and it regenerates a carrier signal that is both frequency and phase coherent with the original transmit carrier signal

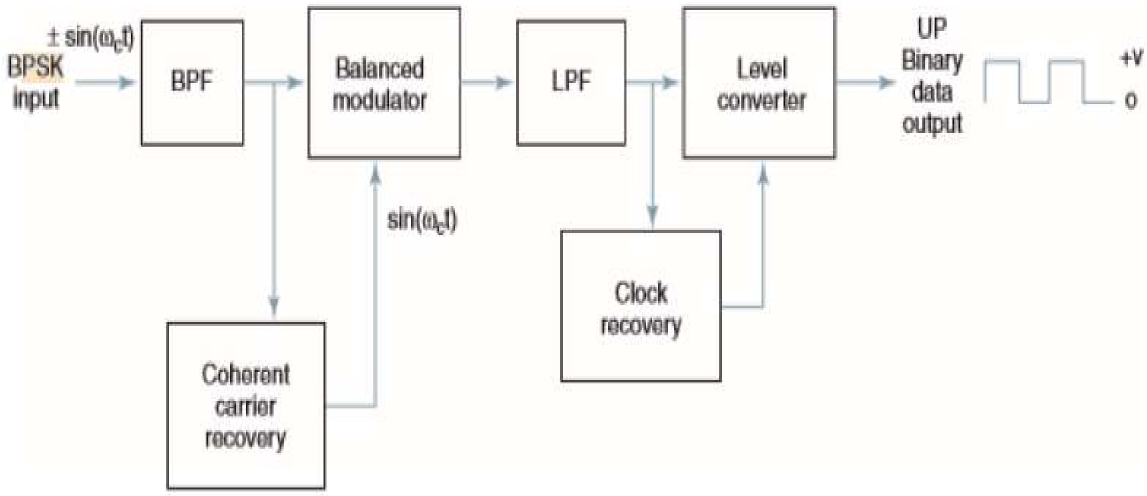


Fig. Block diagram of a BPSK receiver

The balanced modulator is a product detector; the output is the product of the two inputs (the BPSK signal and the recovered carrier). The low-pass filter (LPF) separates the recovered binary data from the complex demodulated signal.

For a BPSK input signal of + sin $\omega_c t$ (logic 1), the output of the balanced modulator is,

Output=(sin $\omega_c t$)(sin $\omega_c t$)=sin2$\omega_c t$ =1/2(1−cos2$\omega_c t$) =1/2−1/2cos2$\omega_c t$ =+1/2V(logic 1)

For a BPSK input signal of -sin ωct (logic 0), the output of the balanced modulator is, Output=(− sin $\omega_c t$)(sin $\omega_c t$)=−sin2$\omega_c t$ =−1/2(1−cos2$\omega_c t$) =−1/2+1/2cos2$\omega_c t$ =−1/2V(logic 0)

Probability Of Error:

The Probability of error of a BPSK signal, Pe=1/2 erfc[√Eb/N]

Application of BPSK

1. It is used for high bit rates transmission even higher tha 1800 bits/sec
2. Due to low bandwidth requirement the BPSK modems are preferred over the FSK modems,at higher operating speeds.

Advantage

1. It has lower bandwidth than BFSK signal

2. Best performance in presence of noise.

3. It gives minimum probability of error

Disadvantage

1.Detection of BPSK signal is difficult because of carrier synchronization

Different Phase Shift Keying

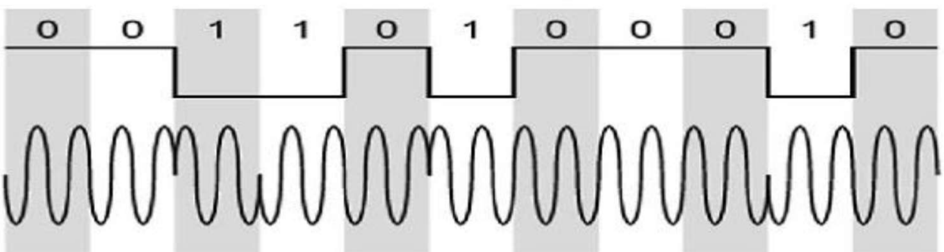
2.(a)Explain in detail about Differential Phase Shift Keying.

Or

(b)With block diagram,Explain DPSK transmitter and Receiver

In **Differential Phase Shift Keying** , the phase of the modulated signal is shifted relative to the previous signal element. No reference signal is considered here. The signal phase follows the high or low state of the previous element. This DPSK technique doesn't need a reference oscillator.

The following figure represents the model waveform of DPSK

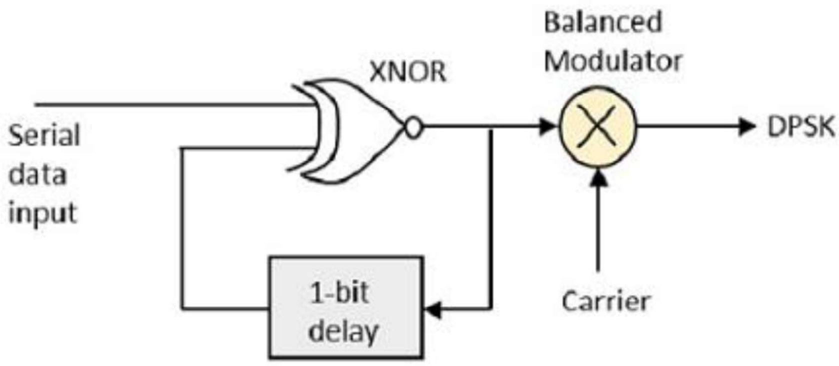


It is seen from the above figure that, if the data bit is Low i.e., 0, then the phase of the signal is not reversed, but continued as it was. If the data is a High i.e., 1, then the phase of the signal is reversed, as with NRZI, invert on 1 a form of differential encoding.

If we observe the above waveform, we can say that the High state represents an **M** in the modulating signal and the Low state represents a **W** in the modulating signal.

DPSK Modulator

DPSK is a technique of BPSK, in which there is no reference phase signal. Here, the transmitted signal itself can be used as a reference signal. Following is the diagram of DPSK Modulator.

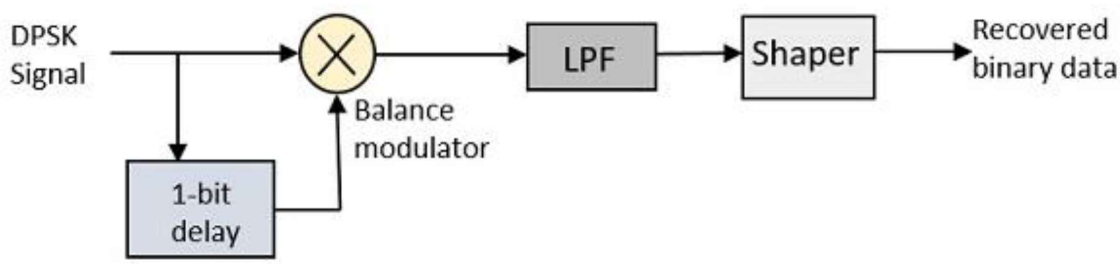


DPSK Modulator

DPSK encodes two distinct signals, i.e., the carrier and the modulating signal with 180° phase shift each. The serial data input is given to the XNOR gate and the output is again fed back to the other input through 1-bit delay. The output of the XNOR gate along with the carrier signal is given to the balance modulator, to produce the DPSK modulated signal.

DPSK Demodulator

In DPSK demodulator, the phase of the reversed bit is compared with the phase of the previous bit. Following is the block diagram of DPSK demodulator.



From the above figure, it is evident that the balance modulator is given the DPSK signal along with 1-bit delay input. That signal is made to confine to lower frequencies with the help of LPF. Then it is passed to a shaper circuit, which is a comparator or a Schmitt trigger circuit, to recover the original binary data as the output.

Bandwidth of DPSK

Transmission Bandwidth

Phase shift of DPSK signal is dependent on the existing bit and one previous bit.

$\therefore$ Symbol duration $T_s = 2T_b$

Where T_b is one bit duration

$$\text{Bandwidth} = \frac{2}{T_s}$$

$$= \frac{2}{2T_b}$$

$$\text{Bandwidth} = f_b$$

Bandwidth in terms of baud rate

$$\text{Bandwidth} = f_b$$

Advantage of DPSK

1. No Carrier synchronization takes place here
2. Requires lower bandwidth

Disadvantages of DPSK

1. Bit determination is made on the basis of the signal received in two successive bit intervals. Hence noise in one bit interval may cause error to two bit determinations.

2. Probability of error is higher

3. Effect of noise is higher.

Quadrature Phase shift Keying

3.(a)Explain in detail about QPSK transmitter and Receiver.
Or
(b)With neat diagram,explain Quadrature Phase Shift Keying.
Or
(c)Briefly Explain about QPSK and also draw the phasor and constellation diagram

- Quadrature Phase shift Keying also called as Quarternary phase shift keying
- QPSK is an M ary encoding scheme where N=2 and M = 4
- Four phases (o/p) are possible for a single carrier frequency.since,there are four output phases,there must be four different input conditions.
- Product modulator requires more than single i/p bit to determine the o/p condition .
- Four possible conditions : 00,01,10,11
- The binary i/p data are combined into group of two bits called **dibits.**
- In the modulator, each dibit code generates one of the four possible o/p phases ($+45^0$,

$+135^0$, -45^0 , -135^0).

- For each two dibit clocked into the modulator single o/p change occurs.
- Baud rate = ½ bit i/p rate (two i/p bit produce one o/p phase change).

QPSK transmitter

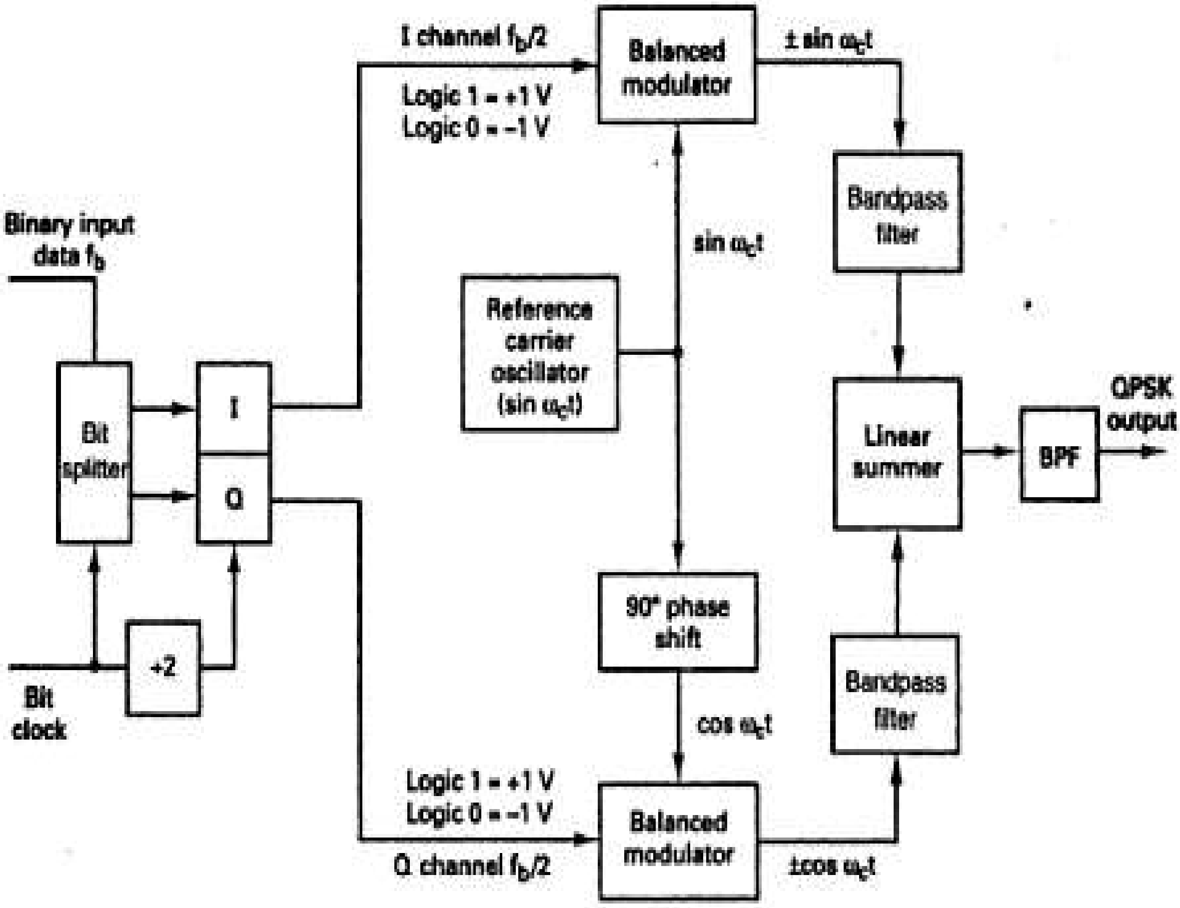
- The block diagram of a QPSk modulator is shown in figure.Two bits are clocked into bit splitter. After both bits are serially inputted, they are simultaneously parallel outputted.one bit is passed to the I channel and the other bit passed to the Q channel
- The I bit modulates a carrier signal is inphase with the reference oscillator &

Q bit modulates a carrier signal is 90^0 out of phase (or) in quadrature with reference carrier

- Basically, a QPSK modulator is two BPSK modulators combined in parallel
- For a logic1=+1V and for logic 0= -1V is assigned. Two phases are occurs at the output of I balanced modulator($+\sin\omega_c t$,$-\sin\omega_c t$) similarly two phases are occurs at the out put of Q balanced modulator($+\cos\omega_c t$,$-\cos\omega_c t$)
- The linear summer adds the two quadrature (90^0 out of phase)signals, resulting in four phasors.

(i.e) ($+\sin\omega_c t+\cos\omega_c t$, $+\sin\omega_c t-\cos\omega_c t$, $-\sin\omega_c t+\cos\omega_c t$, $-\sin\omega_c t-\cos\omega_c t$)

- Since the phasor diagram has four possible output phasors,which are equal in amplitude,the binary information must be encoded in the phase of the output signal.
- Angular separation between any two adjacent phasors in QPSK is 90^0.Hence,a QPSK signal can undergo almost $+45^0$ or -45^0 shift in phase during transmission & still retain the correct encoded binary information,when demodulated at receiver.



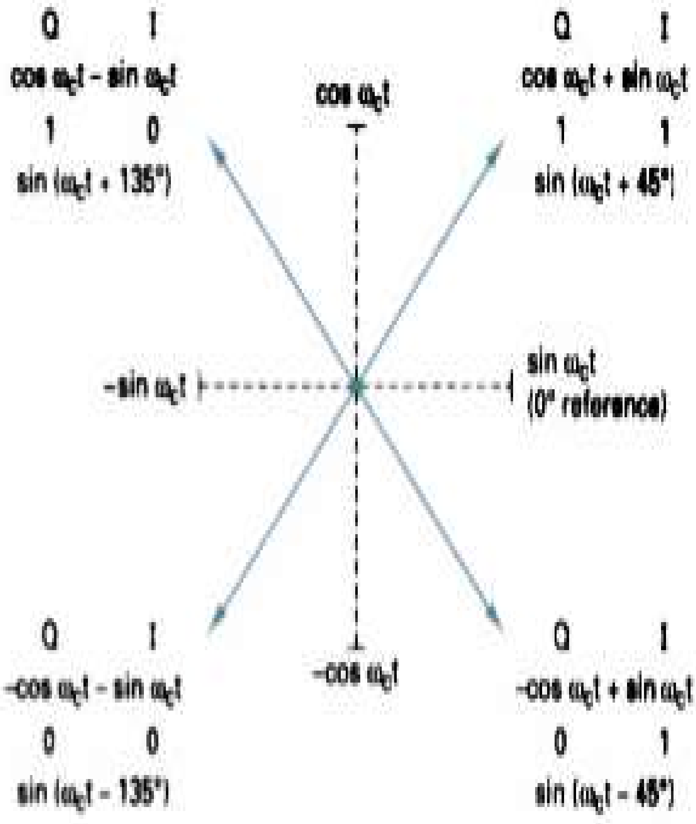
Linear summer o/p = $(+\sin\omega_c t + \cos\omega_c t, +\sin\omega_c t - \cos\omega_c t, -\sin\omega_c t + \cos\omega_c t, -\sin\omega_c t - \cos\omega_c t)$

Each of the four possible o/p phasors has exactly the same amplitude.

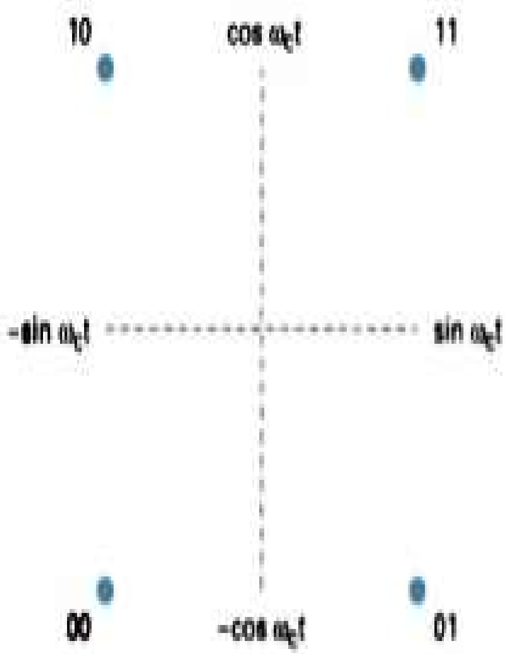
Binary information must be encoded entirely in the phase of the o/p signal.

Binary input Q	Binary input I	QPSK output phase
0	0	-135°
0	1	-45°
1	0	+135°
1	1	+45°

(a)



(b)

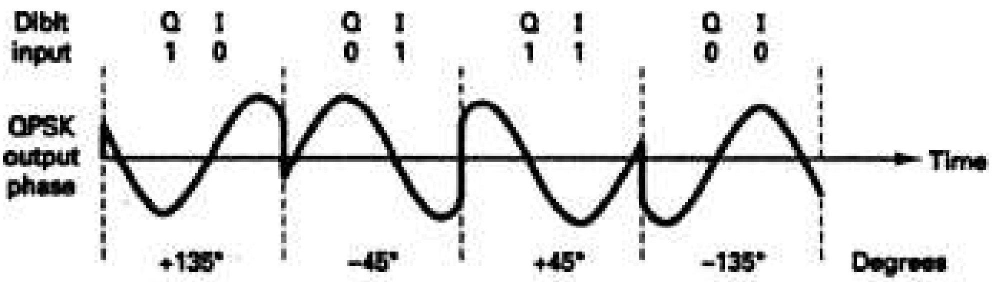


(c)

Fig. QPSK modulator: (a) Truth table; (b) Phasor diagram; (c) Constellation diagram

Basically, a QPSK modulator is two BPSK modulators combined in parallel. For a logic 1 =+1 V and a logic 0 = -1 V is assigned. Two phases are occur at the output of the I balanced modulator (+sin $\omega_c t$, - sin $\omega_c t$). Similarly two phases are occurs at the output of the Q balanced modulator (+cos $\omega_c t$ and -cos $\omega_c t$). When the linear summer adds the two quadrature (90° out of phase) signals, there are four possible resultant phasors,

(+sin $\omega_c t$ + cos $\omega_c t$, + sin $\omega_c t$ - cos $\omega_c t$, - sin $\omega_c t$ + cos $\omega_c t$, - sin $\omega_c t$ - cos $\omega_c t$)



Output phase-versus-time relationship for a QPSK modulator

Bandwidth consideration of QPSK

- **Bit rate (I or Q) equal to i/p data rate (f_b / 2).**
- **Highest fundamental frequency present at the data i/p to I or Q = ¼ of data rate .**
- Twice (I & Q) $f_b/4 = f_b/2$ (nyquist BW).
- Bandwidth compression is realized in QPSK (min BW < incoming bit rate).
- QPSK o/p signal doesnot change phase until two bits have been clocked into bit splitter.**o/p rate (baud) = ¼ of i/p bit rate.**

Output of the balanced modulator

Output = (sin ω_at)(sin ω_ct)

$\omega_a t = 2\pi f_b/4\ t$ **{modulating signal}** $\omega_a t = 2\pi f_c$ **{carrier} Thus,**

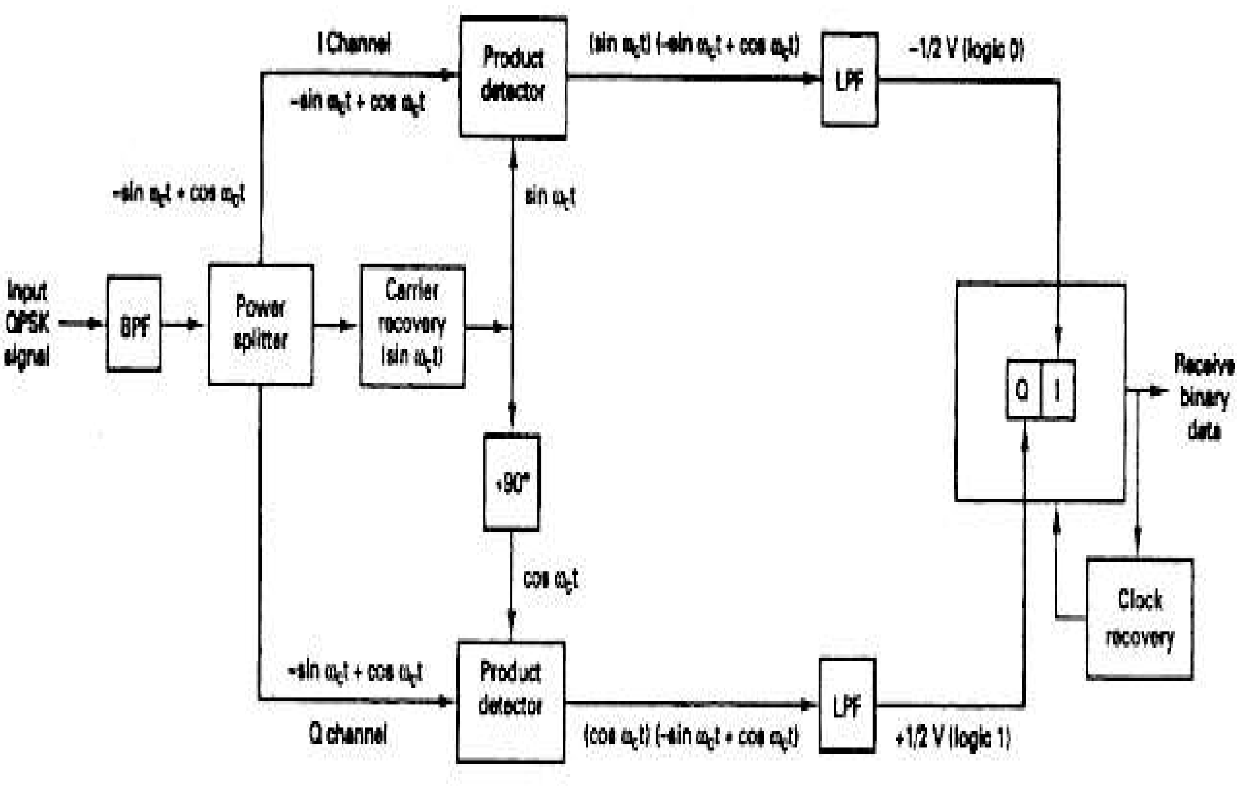
Output = $(\sin 2\pi f_b/4\ t)(\sin 2\pi f_c t)$

½ **cos** $2\pi(f_c - f_b/4)\ t$ - ½ **cos** $2\pi(f_c + f_b/4)\ t$

The output frequency spectrum extends from $f_c + f_b/4$ to $f_c - f_b/4$, and the **minimum bandwidth**
(f_N) is

$(f_c + f_b/4) - (f_c - f_b/4) = 2\ f_b/4 = \mathbf{f_b/2}$

QPSK receiver



- The power splitter directs the i/p QPSK signal to the I & Q product detectors and carrier recovery circuit.
- Carrier recovery circuit recover / reproduce the original transmit carrier.
- QPSK signal is demodulated in I& Q product detector generate original I & Q data.
- The output of the product detectors given to the combining circuit,where they are converted from parallel I & Q data channels to a single binary output data stream.

- Four possible o/p phases ($+45^0$,$+135^0$, -45^0 , -135^0).

I product detector

The received QPSK signal (**$-\sin\omega_c t + \cos\omega_c t$**) is one of the input to the **I**

product detector other i/p (recovered carrier) **sinω_ct.**

$I = (-\sin\omega_c t + \cos\omega_c t)(\sin\omega_c t)$

$= (-\sin\omega_c t)(\sin\omega_c t) + (\cos\omega_c t)(\sin\omega_c t)$

$= -\sin^2\omega_c t + (\cos\omega_c t)(\sin\omega_c t)$

$= -½ (1 - \cos 2\omega_c t) + ½ \sin(\omega_c + \omega_c)t + ½ \sin(\omega_c - \omega_c)t$

$= -½ + ½ \cos 2\omega_c t + ½ \sin 2\omega_c t + ½ \sin 0$

Filteredout

equals 0

$= -½ \quad v$

(logic 0)

Q product

Again, the receive QPSK signal is one of the input to the **Q product detector** .the other input is recovered carrier shifted 90^0 in phase.

$I = (-\sin\omega_c t + \cos\omega_c t)(\cos\omega_c t)$

QPSK signal carrier

$= \cos^2\omega_c t - (\sin\omega_c t)(\cos\omega_c t)$

$= ½ (1 - \cos 2\omega_c t) - ½ \sin(\omega_c + \omega_c)t - ½ \sin(\omega_c - \omega_c)t$

$= ½ + ½ \cos 2\omega_c t - ½ \sin 2\omega_c t - ½ \sin 0$

Filtered out equals 0

$= ½$ v (logic 1)

Adv:

- limited phase shift.
  - For the same Bit Error rate(BER) the bandwidth required by QPSK is reduced to half as compared to BPSK
  - Due to reduced bandwidth,the information transmission rate of QPSK is high

- Amplitude remains constant,hence the carrier power is also remains constant

Disadvantage:

- Generation and detection is complex

Principles of M-ary signaling

M-ary PSK

4.(a).Explain M-ary PSK.

Or

(b).Briefly explain the principle of M-ary PSK transmitter and receiver.

In M-ary PSK, the phase of the carrier takes one of M possible values, namely, $\theta_i = 2i\pi/M$, where $i = 0, 1, \ldots, M-1$. Accordingly, during each signaling interval of duration T, one of the M possible signals

$$s_i(t) = \sqrt{\frac{2E}{T}}\cos\left(2\pi f_c t + \frac{2\pi i}{M}\right) \qquad i = 0,1,\ldots\ldots,M-1$$

is sent, where E is the signal energy per symbol. The carrier frequency $f_c = n_c/T$ for some fixed integer n_c

Each $s_i(t)$ is expanded in terms of two basis functions ϕ_1 and ϕ_2 defined as

$$\phi_1(t) = \sqrt{\frac{2}{T}}\cos(2\pi f_c t) \qquad 0 \le t \le T$$

$$\phi_2(t) = \sqrt{\frac{2}{T}}\sin(2\pi f_c t) \qquad 0 \le t \le T$$

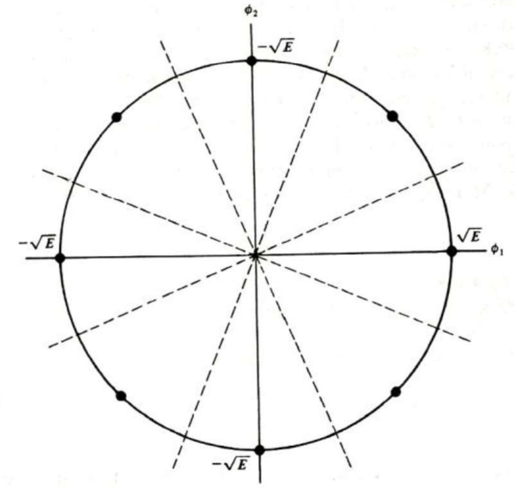


Figure : Signal constellation for octaphase-shift-keying (i.e., M = 8).

The decision boundaries are shown as dashed lines.

Both ϕ_1 and ϕ_2 have unit energy. The signal constellation of M-ary PSK is therefore two-dimensional. The M message points are equally spaced on a circle of radius $\sqrt{E}$ and center at the origin, as illustrated in Figure for *octaphase-shift-keying* (i.e., M = 8). This figure also includes the corresponding decision boundaries indicated by dashed lines.

The optimum receiver for coherent M-ary PSK (assuming perfect synchronization with the transmitter) is shown in block diagram form in Figure.

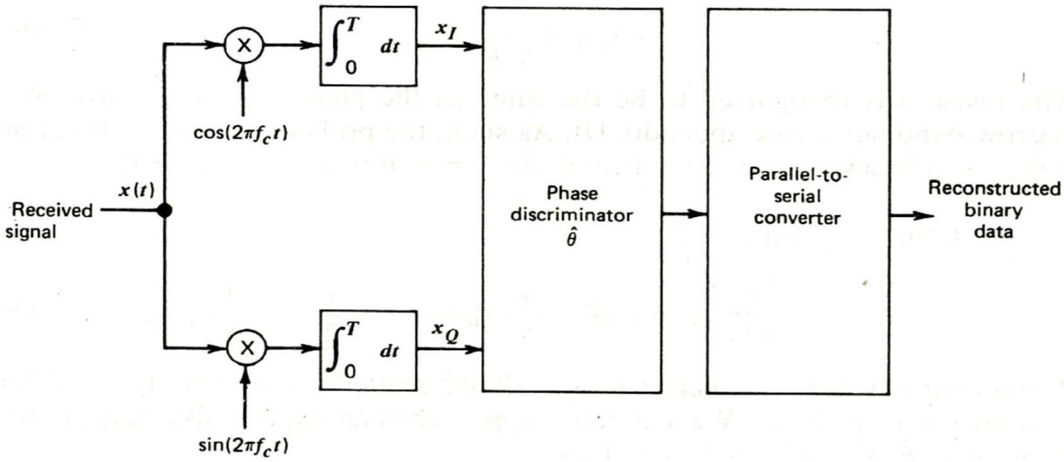


Figure :Receiver for coherent M-ary PSK.

It includes a pair of correlators with reference signals in phase quadrature. The two correlator outputs, denoted as x_I *and* x_Q, are fed into a **phase discriminator** that first computes the phase estimate

$$\hat{\theta} = \tan^{-1}\left(\frac{x_Q}{x_I}\right)$$

The phase discriminator then selects from the set $\{s_i(t), i = 0,..,M - I\}$ that particular signal whose phase is closest to the estimate $\hat{\theta}$

In the presence of noise, the decision-making process in the phase discriminator is based on the noisy inputs

$$x_I = \sqrt{E}\cos\left(\frac{2\pi i}{M}\right) + w_I \qquad i = 0,1,\dots\dots, M-1$$

and

$$x_Q = -\sqrt{E}\sin\left(\frac{2\pi i}{M}\right) + w_Q \qquad i = 0,1,\dots\dots, M-1$$

where w_I *and* w_Q are samples of two independent Gaussian random variables W_I *and* W_Q whose mean is zero and common variance equals

$$\sigma^2 = \frac{N_0}{2}$$

In Figure, message points exhibit circular symmetry. Also, the random variables W_I *and* W_Q have a symmetric probability density function. Thus, in an

M-ary PSK system, the average probability of symbol error,P_e is independent of the transmitted signal $s_i(t)$. Therefore, we simplify the calculation of P_e by setting $\theta = 0$, which corresponds to the message point whose coordinates along the $\phi_1(t) -$ *and* $\phi_2(t) -axes$ are $+\sqrt{E}$ *and* 0, respectively. The decision region related to this message point [i.e., the signal $s_o(t)$] is bounded by the threshold $\hat{\theta} = -\pi M$ below the $\phi_1(t) - axis$ and the threshold $\hat{\theta} = \pi M$ above the $\phi_1(t) -$ axis. Therefore the probability of correct reception is

$$P_c = \int_{-\pi/M}^{\pi/M} f_\Theta(\hat{\theta})\, d\hat{\theta}$$

where $f_\theta(\hat{\theta})$ is the probability density function of the random variable Θ whose sample value equals the phase discriminator output θ produced in response to a received signal that consists of the signal $s_o(t)$ plus AWGN. That is,

$$\hat{\theta} = \tan^{-1}\left(\frac{W_Q}{\sqrt{E} + W_I}\right)$$

The phase $\hat{\theta}$ is same as the phase of a sine wave plus narrow-band noise. As such, the probability density function $f_\theta(\hat{\theta})$ has a known value. Specifically, for$-\pi \leq \hat{\theta} \leq -\pi$, we can write

$$f_\Theta(\hat{\theta}) = \frac{1}{2\pi}\exp\left(-\frac{E}{N_0}\right) + \sqrt{\frac{E}{\pi N_0}}\cos\hat{\theta}\exp\left(-\frac{E}{N_0}\sin^2\hat{\theta}\right)\left[1 - \frac{1}{2}erfc\left(\frac{E}{N_0}\cos\hat{\theta}\right)\right] \quad --(1)$$

The probability density function $f_\theta(\hat{\theta})$ is plotted versus θ shown in Fig. 3.21 for various values of E/N_o. It approaches an impulse-like appearance about θ as E/N_o goes high.

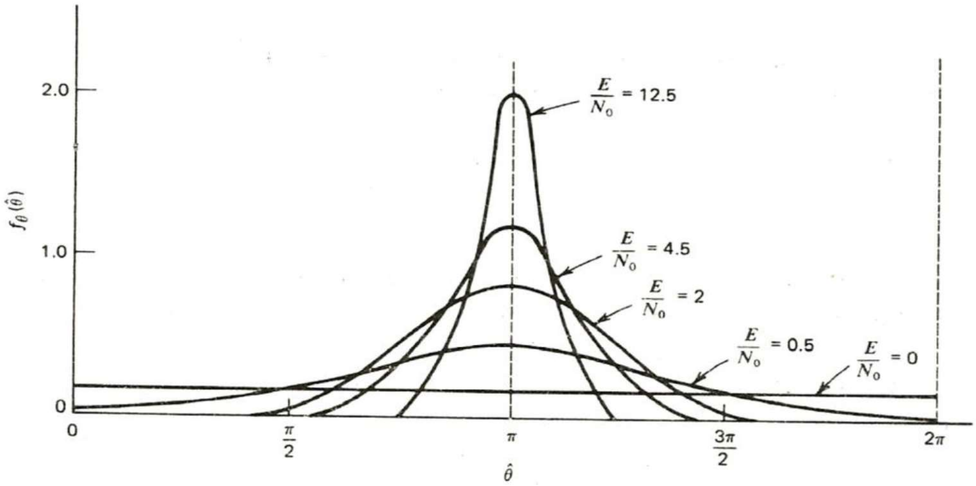


Figure: Probability density function of phase estimate

A decision error is made if the angle $\hat{\theta}$ falls outside $-\pi/M \leq \hat{\theta} \leq -\pi/M$.

The probability of symbol error is therefore

$$P_e = 1 - P_c$$

$$= 1 - \int_{-\pi/M}^{\pi/M} f_\Theta(\hat{\theta})\,d\hat{\theta} \qquad --(2)$$

The integral in eqn (2) does not reduce to a simple form, except for $M = 2$ and $M = 4$. Hence, for $M > 4$, it must be evaluated by using numerical integration.

However, for large M and high values of E/N_o, we can derive an approximate formula for P_e. For high values of E/N_o and for $\hat{\theta} < \pi/2$, we use the approximation

$$erfc\left(-\frac{E}{N_0}\cos\hat{\theta}\right) \approx \sqrt{\frac{N_0}{\pi E}}\frac{1}{\cos\hat{\theta}}\exp\left(-\frac{E}{N_0}\cos^2\hat{\theta}\right)$$

Hence, using this approximation for the complementary error function in Eq. (1) and simplifying terms, we get

$$f_\Theta(\hat{\theta}) \approx \sqrt{\frac{E}{\pi N_0}}\cos\hat{\theta}\exp\left(-\frac{E}{N_0}\sin^2\hat{\theta}\right) \qquad |\hat{\theta}| < \frac{\pi}{2} \qquad --(3)$$

Thus, substituting Eq. (3) in Eq. (2), we get

$$P_e \approx 1 - \sqrt{\frac{E}{\pi N_0}}\int_{-\pi/M}^{\pi/M} \cos\hat{\theta}\exp\left(-\frac{E}{N_0}\sin^2\hat{\theta}\right) \qquad --(4)$$

Changing the variable of integration from $\hat{\theta}$ to

$$z = \sqrt{\frac{E}{N_0}}\sin\hat{\theta}$$

Rewriting eqn (4)

$$P_e \approx 1 - \frac{2}{\sqrt{\pi}}\int_0^{\sqrt{E/N_0}\sin(\pi/M)} \exp(-z^2)\,dz$$

$$= erfc\left(\sqrt{\frac{E}{N_0}}\sin(\pi/M)\right) \qquad --(5)$$

This is the desired approximate formula for the probability of symbol error for coherent M-ary PSK ($M \geq 4$). The approximation becomes fixed, for fixed M, as E/N_o is increased.

Coherent M-ary PSK requires exact knowledge of the carrier frequency and phase for the receiver to be accurately synchronized to the transmitter. When carrier recovery at the receiver is impossible, we can use differential encoding based on the phase difference between successive symbols with some degradation in performance. If the incoming data are encoded by a phase shift rather than by absolute phase, the receiver performs detection by comparing the phase of one symbol with that of the previous symbol, and the need for a coherent reference is thereby eliminated. This procedure is the same as binary DPSK. The exact calculation of probability of symbol error for the differential detection of differential M-ary PSK (commonly referred to as *M-ary* DPSK) is too complicated for $M > 2$. However, for large values of E/N_o and $M \geq 4$, the probability of symbol error is approximately given by,

$$P_e \approx erfc\left(\sqrt{\frac{2E}{N_0}}\sin\left(\frac{\pi}{2M}\right)\right) \qquad M \geq 4 \qquad\qquad --(6)$$

Comparing Eqs. (5) and (6), for $M \geq 4$ an M-ary DPSK system attains the same probability of symbol error as the corresponding M-ary PSK system provided that the transmitted energy per symbol is increased by the following factor:

$$k(M) = \frac{\sin^2\left(\frac{\pi}{M}\right)}{2\sin^2\left(\frac{\pi}{2M}\right)} \qquad M \geq 4$$

For example, $k(4) = 1.7$. That is, differential QPSK (which is noncoherent) is approximately $2.3\,dB$ poorer in performance than coherent QPSK.

M-ary QAM

5.(a)Explain in detail about M-ary QAM with neat block diagram and derive the power spectral characteristics.

Or

(b).Explain in detail about M-ary QAM transmitter and receiver.

In an M-ary PSK system, in-phase and quadrature components of the modulated signal are interrelated in such a way that the envelope is constrained to remain constant. This constraint marked itself in a circular constellation for the message points. However, if this constraint is removed a new modulation scheme called *M-ary quadrature amplitude modulation* (QAM) is obtained. In this modulation scheme, the carrier experiences amplitude as well as phase modulation.

The signal constellation for M-ary QAM consists of a square lattice of message points, as illustrated in Figure for $M = 16$.

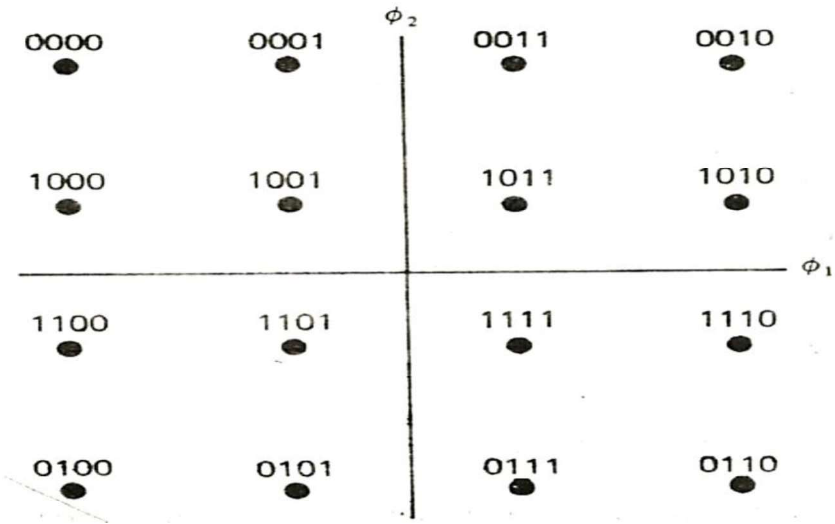


Figure : Signal-constellation of M-ary QAM for M = 16

The corresponding signal constellations for the in-phase and quadrature components of the amplitude-phase modulated wave are shown in Figs. *a* and *b,* respectively.

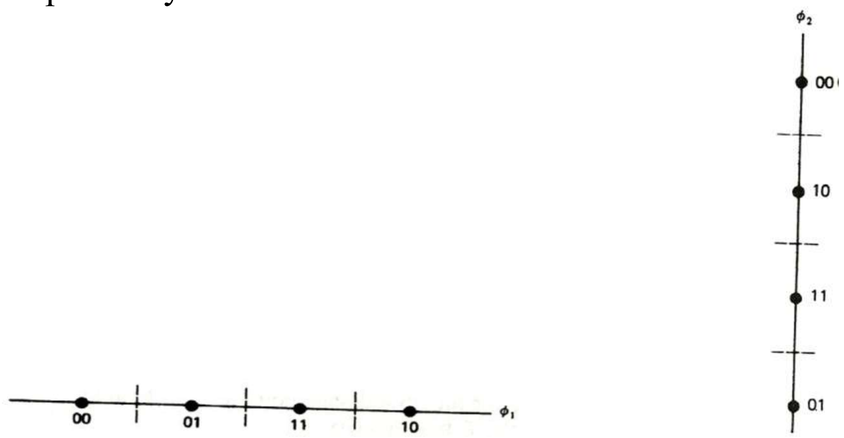


Figure: Decomposition of signal constellation of M-ary QAM (for *M* = 16) into two signal-space diagrams for *(a)* in-phase comment $\boldsymbol{\phi_1(t)}$, and *(b)* quadrature comment $\boldsymbol{\phi_1(t)}$.

In general, an M-ary QAM scheme transmits $M = L^2$ independent symbols over the same channel bandwidth. The general form of M-ary QAM is defined by the transmitted signal

$$s_i(t) = \sqrt{\frac{2E_0}{T}}a_i \cos(2\pi f_c t) + \sqrt{\frac{2E_0}{T}}b_i \sin(2\pi f_c t) \qquad 0 \le t \le T$$

where E_o is the energy of the signal with the lowest amplitude, and a_i *and* b_i are a pair of independent integers chosen according to the location of the message point. The signal $s_i(t)$ consists of two phase-quadrature carriers, each of which is modulated by a set of discrete amplitudes; hence, it is called as "quadrature amplitude modulation." (QAM). The signal $s_i(t)$ can be expanded in terms of a pair of basis functions

$$\phi_1(t) = \sqrt{\frac{2}{T}}\cos(2\pi f_c t) \qquad 0 \le t \le T$$

and

$$\phi_2(t) = \sqrt{\frac{2}{T}}\sin(2\pi f_c t) \qquad 0 \le t \le T$$

The coordinates of the i^{th} message, point are $a_i\sqrt{E}$ *and* $b_i\sqrt{E_o}$, where (a_i, b_i) is an element of the *L-by-L* matrix:

$$\{a_i, b_i\} = \begin{bmatrix} (-L+1, L-1) & (-L+3, L-1) & \dots & (L-1, L-1) \\ (-L+1, L-3) & (-L+3, L-3) & \dots & (L-1, L-3) \\ \vdots & \vdots & & \vdots \\ (-L+1, -L+1) & (-L+3, -L+1) & \dots & (L-1, -L+1) \end{bmatrix}$$

where

$$L = \sqrt{M} \qquad --(1)$$

For example, for the 16-QAM whose signal constellation is shown in Figure, where $L = 4$, we have the matrix

$$\{a_i, b_i\} = \begin{bmatrix} (-3,3) & (-1,3) & (1,3) & (3,3) \\ (-3,1) & (-1,1) & (1,1) & (3,1) \\ (-3,-1) & (-1,-1) & (1,-1) & (3,-1) \\ (-3,-3) & (-1,-3) & (1,-3) & (3,-3) \end{bmatrix}$$

To calculate the probability of symbol error for M-ary QAM, following steps are followed,

1. Since the in-phase and quadrature components of M-ary QAM are independent, the probability of correct detection for such a scheme can be written as

$$P_c = (1 - P_e')^2$$

where P_e' is the probability of symbol error for both component.

2. The signal constellation for the in-phase or quadrature component has geometry similar to that for discrete pulse-amplitude modulation (PAM) with a corresponding number of amplitude levels. Therefore we write

$$P'_e = \left(1 - \frac{1}{L}\right) erfc\left(\sqrt{\frac{E_0}{N_0}}\right) \qquad --(2)$$

where L is the square root of M.

3. The probability of symbol error for M-ary QAM is given by

$$\begin{aligned} P_e &= 1 - P_c \\ &= 1 - (1 - P'_e)^2 \\ &\approx 2P'_e \qquad --(3) \end{aligned}$$

where it is assumed that P'_e is small compared to unity. Putting Eqn.(1) and eqn (2) in Eq, (3), we find that the probability of symbol error for M-ary, QAM is given by

$$P_e \approx 2\left(1 - \frac{1}{\sqrt{M}}\right) erfc\left(\sqrt{\frac{E_0}{N_0}}\right) \qquad --(4)$$

The transmitted energy in M-ary QAM is variable so, its instantaneous value depends on the particular symbol transmitted. Thus, P_e can be expressed in terms of the *average* value of the transmitted energy rather than E_o. Assuming that the L amplitude levels of the in-phase or quadrature component are equally likely, we have

$$E_{av} = 2\left[\frac{2E_0}{L}\sum_{i=1}^{L/2}(2i-1)^2\right] \qquad --(5)$$

where the multiplying factor of 2 is for the equal contributions by the in-phase and quadrature components. Summing the series in Eq. (5), we get

$$\begin{aligned} E_{av} &= \frac{2(L^2-1)E_0}{3} \\ &= \frac{2(M-1)E_0}{3} \end{aligned}$$

Accordingly, we can rewrite Eq. (4) in terms of E_{av} as

$$P_e \approx 2\left(1 - \frac{1}{\sqrt{M}}\right) erfc\left(\sqrt{\frac{2E_{av}}{2(M-1)N_0}}\right)$$

which is the desired result.

In the special case is $M = 4$. The signal constellation for this value of M is same as that for QPSK..

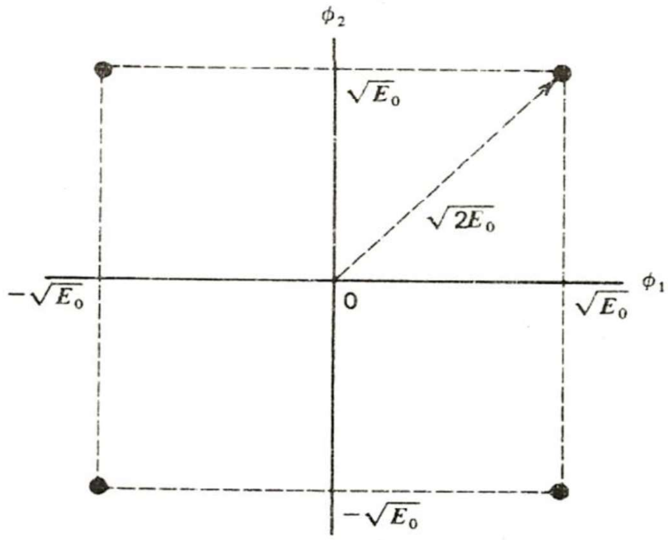
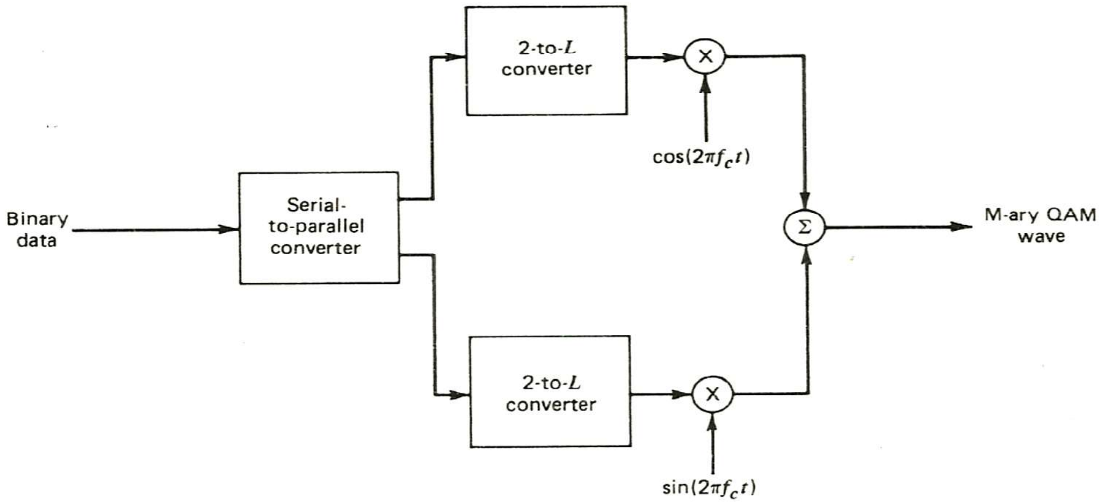
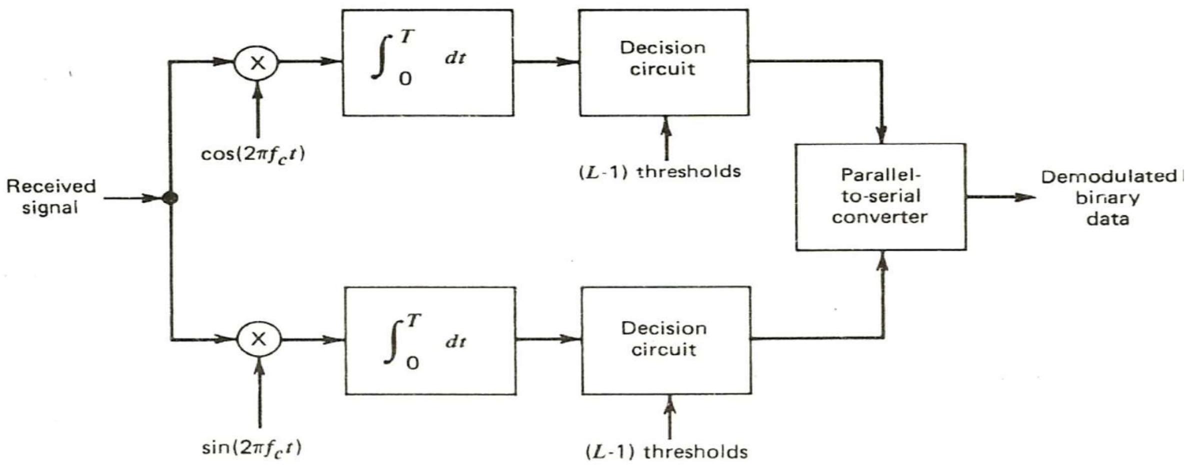


Figure Signal constellation for the special case of M-ary QAM for M = 4

Figure *a* shows the block diagram of an M-ary QAM transmitter.



(a)



(b)

Figure : Block diagrams of M-ary QAM system. *(a)* Transmitter. *(b)* Receiver.

The serial-to-parallel converter accepts a binary sequence at a bit rate $R_b = 1/T_b$

and produces two parallel binary sequences whose bit rates are $R_b/2$ each. The 2-to- L level converters, where $L = \sqrt{M}$, generate polar L-level signals in response to the respective in-phase and quadrature channel inputs. Quadrature-carrier multiplexing of the two polar L-level signals so generated produces the desired M-ary QAM signal.

Figure shows the block diagram of the corresponding receiver. Decoding of each baseband channel is done at the output of the decision circuit. The decision circuit compares the L-level signals against L - 1 decision thresholds. The detected two binary sequences are then combined in the parallel-to-serial converter to reproduce the original binary sequence

6.(a).Give the Comparison of M-ary Digital Modulation Techniques

Or

(b).Compare M-ary PSK and M-ary QAM.

In Table typical values of power-bandwidth requirements for coherent binary and M-ary PSK schemes are summarized, assuming an average probability of symbol error equal to 10^{-4} and that the systems operate in identical noise environments. This table shows that, among M-ary PSK signals, QPSK (corresponding to $M = 4$) offers the best trade-off between power and bandwidth requirements. For $M > 8$, power requirements become high; accordingly, M-ary PSK schemes with $M > 8$ are not as widely used in practice. Also, coherent M-ary PSK schemes require more complex equipment than coherent binary PSK schemes for signal generation or detection, especially when $M > 8$. Basically, M-ary PSK and M-ary QAM have similar spectral and bandwidth characteristics. For $M > 4$, however, the two schemes have different signal constellations. For M-ary PSK the signal constellation is circular, whereas for M-ary QAM it is rectangular. Moreover, a comparison of these two constellations shows that the distance between the message points of M-ary PSK is smaller than the distance between the message points of M-ary QAM, for a fixed peak transmitted power. This basic difference between the two schemes is illustrated in Figure for $M = 16$. Accordingly, in an AWGN channel, M-ary QAM outperforms the corresponding M-ary PSK in error performance for $M > 4$. However, the superior performance of M-ary QAM can be realized only if the channel is free from nonlinearities.

Comparison of Power-Bandwidth Requirements for M-ary PSK with Binary PSK. Probability of Symbol Error = 10^{-4}

Value of M	$\frac{(\text{Bandwidth})_{\text{M-ary}}}{(\text{Bandwidth})_{\text{Binary}}}$	$\frac{(\text{Average power})_{\text{M-ary}}}{(\text{Average power})_{\text{Binary}}}$
4	0.5	0.34 dB
8	0.333	3.91 dB
16	0.25	8.52 dB
32	0.2	13.52 dB

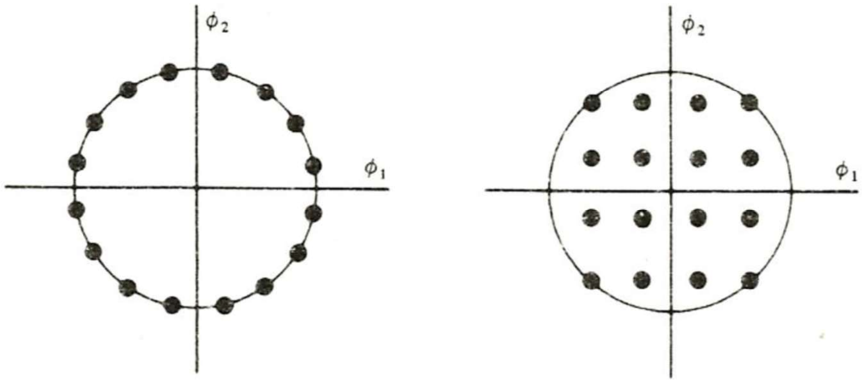


Figure : Signal constellations. (a) M-ary QPSK, and *(b)* M-ary QAM for M = 16

As for M-ary FSK, we find that for a fixed probability of error, increasing M results in a reduced power requirement. However, this reduction in transmitted power is achieved with increased channel bandwidth.

Duobinary encoding

7.(a)Write the expression for duobinary encoding

Or

(b).Explain in detail about Duobinary encoder with neat expression.

Or

(c).Describe Duobinary Signaling technique and its performance by illustrating its frequency and impulse responses.

In duobinary signaling "duo" implies doubling of the transmission capacity of a straight binary system. Consider a binary input sequence $\{bk\}$ consisting of uncorrelated binary digits each having duration T_b seconds, with symbol 1 represented by a pulse of amplitude $+1\ volt$, and symbol 0 by a pulse of amplitude $-1\ volt$. When this sequence is applied to a duobinary encoder, it is converted into a three-level output, namely, $-2, 0, and\ +2\ volts$. This scheme is illustratd in Figure.

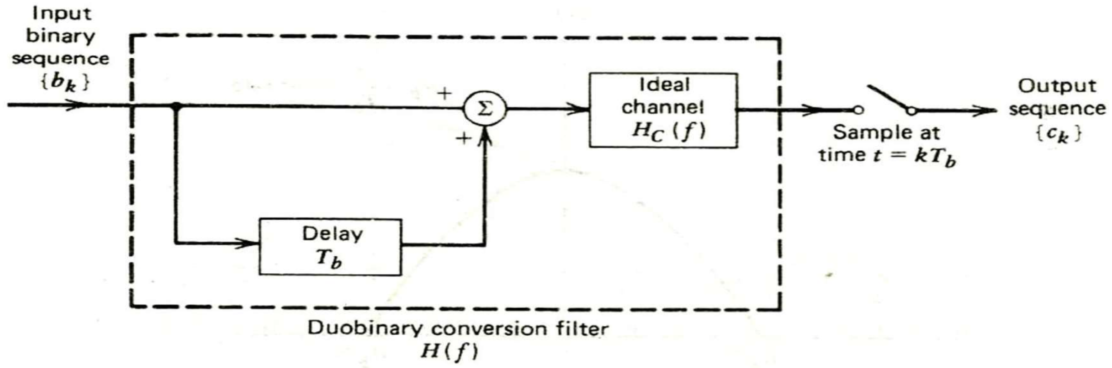


Figure Duobinary signaling scheme

The binary sequence $\{bk\}$ is first passed through a simple filter involving a single delay element. For every unit impulse applied to the input of this filter, we get two unit impulses spaced T_b seconds at the filter output. Thus the digit c_k at the duobinary coder output is the sum of the present binary digit b_k, and its previous value b_{k-1}, as shown by

$$c_k = b_k + b_{k-1} \qquad --(1)$$

In Eq. (1) the input sequence $\{bk\}$ of uncorrelated binary digits is transformed into a sequence $\{c_k\}$ of correlated digits. This correlation between the adjacent transmitted levels is viewed as introducing intersymbol interference into the transmitted signal in an artificial manner. An ideal delay element, producing a delay of T_b seconds, has the transfer function $exp(-j2\pi f\ T_b)$, so that the transfer function of simple filter shown in figure is $1 + exp(-j2\pi f\ T_b)$. Hence, the overall transfer function of this filter connected in cascade with the ideal channel $H_C(f)$ is

$$\begin{aligned} H(f) &= H_C(f)[1 + \exp(-j2\pi f T_b)] \\ &= H_C(f)[\exp(j\pi f T_b) + \exp(-j\pi f T_b)]\exp(-j\pi f T_b) \\ &= 2H_C(f)\cos(\pi f T_b)\exp(-j\pi f T_b) \end{aligned}$$

For an ideal channel of bandwidth $B_o = R_b/2$, we have

$$H_C(f) = \begin{cases} 1 & |f| \leq \dfrac{R_b}{2} \\ 0 & otherwise \end{cases} \qquad --(2)$$

Thus the overall frequency response has the form of a half-cycle cosine function, as shown by

$$H(f) = \begin{cases} 2\cos(\pi f T_b)\exp(-j\pi f T_b) & |f| \leq \dfrac{R_b}{2} \\ 0 & otherwise \end{cases}$$

for which the amplitude response and phase response are as shown in Fig.*2a* and Fi*2b*, respectively. An advantage of this frequency response is that it can be easily approximated in practice.

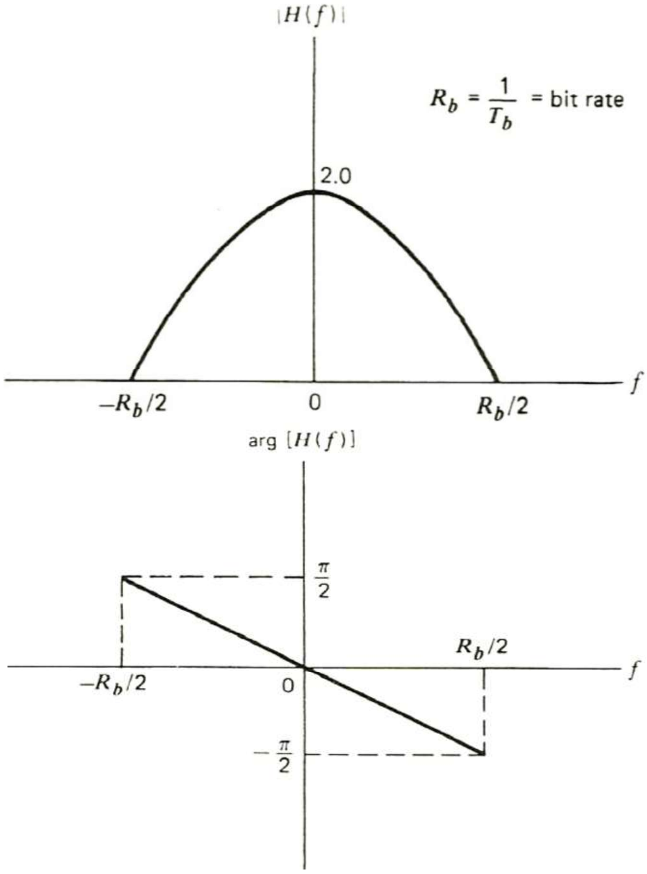


Figure: Frequency response of duobinary conversion filter. *(a)* Amplitude response. *(b)* Phase response

The corresponding value of the impulse response consists of two sine pulses, time-displaced by T_b seconds, as shown by (except for a scaling factor) which is shown plotted in Figure.

$$
\begin{aligned}
h(t) &= \frac{\sin(\pi t/T_b)}{\pi t/T_b} + \frac{\sin(\pi(t-T_b)/T_b)}{\pi(t-T_b)/T_b} \\
&= \frac{\sin(\pi t/T_b)}{\pi t/T_b} - \frac{\sin(\pi t/T_b)}{\pi(t-T_b)/T_b} \\
&= \frac{T_b^2 \sin(\pi t/T_b)}{\pi t(T_b - t)}
\end{aligned}
$$

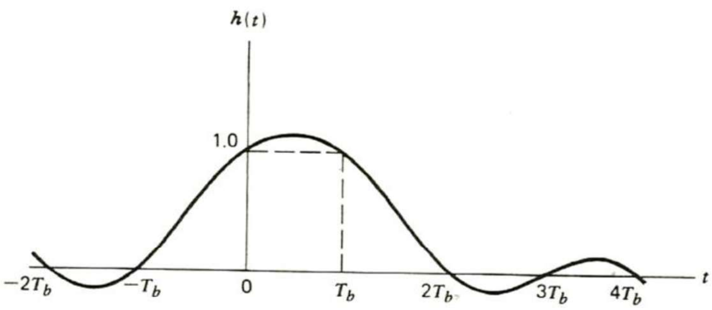


Figure: Impulse response of duobinary conversion filter.

The overall impulse response $h(t)$ has only *two* distinguishable values at the sampling instants.
The original data $\{b_k\}$ can be detected from the duobinary-coded sequence $\{c_k\}$ by subtracting the previous decoded binary digit from the currently received digit c_k in accordance with Eq. (1). Let $\hat{b}_k$ represent the *estimate* of the original binary digit b_k as received by the receiver at time *t* equal to kT_b, we have

$$\hat{b}_k = c_k - \hat{b}_{k-1}$$

If c_k is received without error and if also the previous estimate $\hat{b}_{k-1}$ at time $t = (k - I)T_b$ corresponds to a correct decision, then the current estimate b_k, will also be correct. The technique of using a stored estimate of the previous symbol is called *decision feedback.*
The detection procedure is an inverse of the operation of the simple filter at the transmitter. However, a drawback of this detection process is that once errors are made, they tend to propagate. This is due to the fact that a decision on the current binary digit b_k, depends on the correctness of the decision made on the previous binary digit b_{k-1}.
A practical means of avoiding this error propagation is to use precoding before the duobinary coding, as shown in Fig.

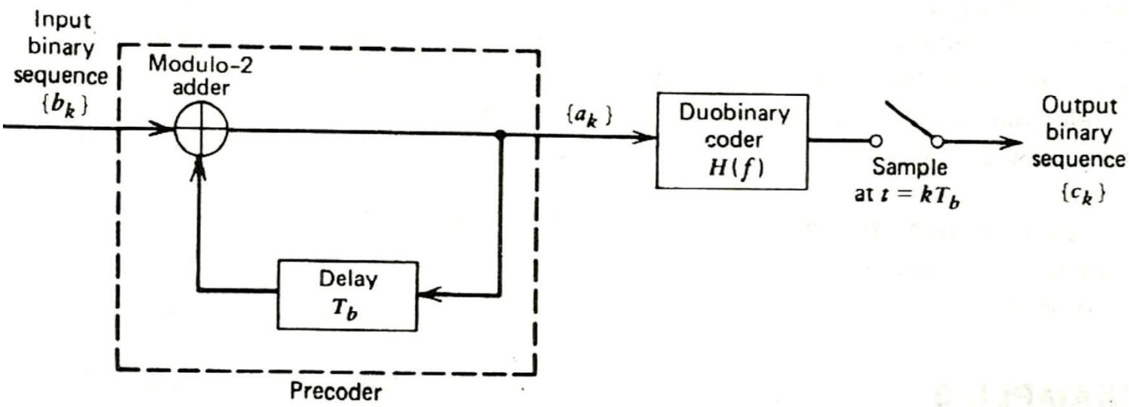


Figure: A precoded duobinary scheme. Details of the duobinary coder are given in Fig.

The precoding operation performed on the input binary sequence $\{b_k\}$ converts it into another binary sequence $\{a_k\}$ defined by

$$a_k = b_k + a_{k-1} \quad modulo - 2 \qquad --(3)$$

Module-2 addition is equivalent to the exclusive-OR (X-OR) operation. Here, if the inputs of an X-OR gate are different, then the output is 1; otherwise, the output is a 0. The resulting precoder output $\{a_k\}$ is next applied to the duobinary coder, thereby producing the sequence $\{c_k\}$ as follows:

$$c_k = a_k + a_{k-1} \qquad --(4)$$

Assume that symbol 1 at the precoder output in Fig. 3.14 is represented by $+1\ volt$ and symbol 0 by $-1\ volt$. Therefore, from Eqs. (3) and (4), we find that

$$c_k = \begin{cases} \pm 2\ volts, & if\ b_k\ is\ represented\ by\ symbol\ 0 \\ 0\ volts, & if\ b_k\ is\ represented\ by\ symbol\ 1 \end{cases} \qquad --(5)$$

From Eq. (5), we deduce the following decision rule for detecting the original input binary sequence $\{b_k\}$ from $\{C_k\}$:

$$b_k = \begin{cases} symbol\ 0 & if\ |c_k| > 1volt \\ symbol\ 1 & if\ |c_k| < 1volt \end{cases} \qquad --(6)$$

According to Eq. (6), the detector consists of a rectifier, the output of which is compared to a threshold of 1 volt, and thus the original binary sequence $\{b_k\}$ is detected.

A block diagram of the detector is shown in Fig.

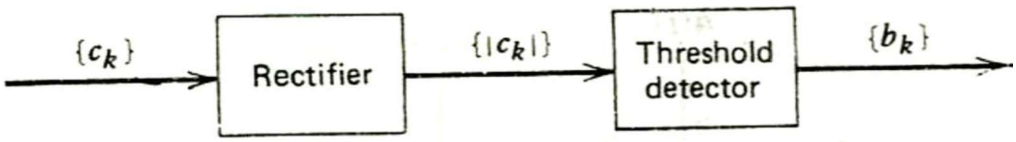


Figure : Detector for recovering original binary sequence from the precoded duobinary coder output.

In this detector only the knowledge of present input sample is required. Hence, error propagation cannot occur in the detector of Fig

INTERSYMBOL INTERREFERENCE

8.(a)Write short notes on Intersymbol interference(ISI)

or

(b).What are the causes of ISI

Fig. (a) shows input signal to an ideal minimum bandwidth, low pass filter.The input signal is a random, binary non return to zero (NPZ) sequence. Fig. (b) shows the output of a LPF that doesnot produce amplitude or phase distortion (perfect filter).The output signal reaches its peak value at the centre of the sampling instant. Fig. (c) shows the output of LPF that produce amplitude or phase distortion (imperfect filter). In such case the output signal doesn't reaches its peak value at the centre of the sampling instant. So that the ringing tails of several pulses gets overlapped and interfere with the major pulse lobe. This interference is referred to as Inter Symbol Interference (ISI). ISI causes cross talk between channels. Equalizers are inserted in the transmission path to reduce distortions. The four primary cause of ISI are.

1. Timing inaccuracies 2. Amplitude distortion
2. Insufficient bandwidth 4. Phase distortion

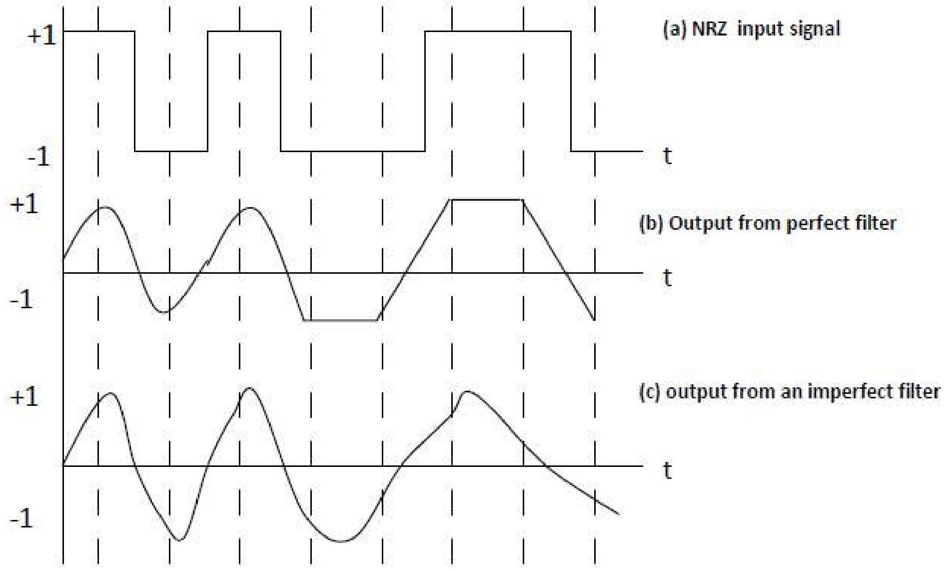


Fig.3.22 Pulse Response

ISI If the bandwidth of signal to be transmitted is less than channel bandwidth, then overlapping takes place resulting in inter symbol interference.

Timing inaccuracies It takes place if the transmission rate does not conform to the ringing frequency of the communication channel. It occurs if the bandwidth (Nyquist bandwidth) of the signal is less than channel bandwidth.

Amplitude distortion When the frequency response of a channel from the output of filter is away from expected values, pulse distortion takes place Pulse distortion occurs when peak of the pulse is reduced. It is compensated using Amplitude equalizers.

Phase distortion It takes place if there is a time delay in the frequency component of the signal while propagating through the transmission medium

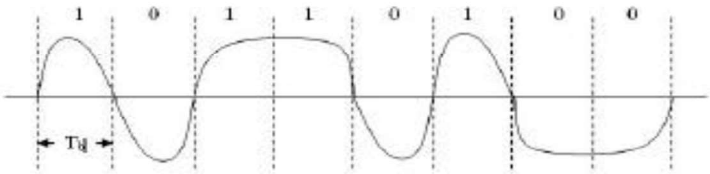
Eye pattern

9.(a).Write Short notes on Eye Pattern.

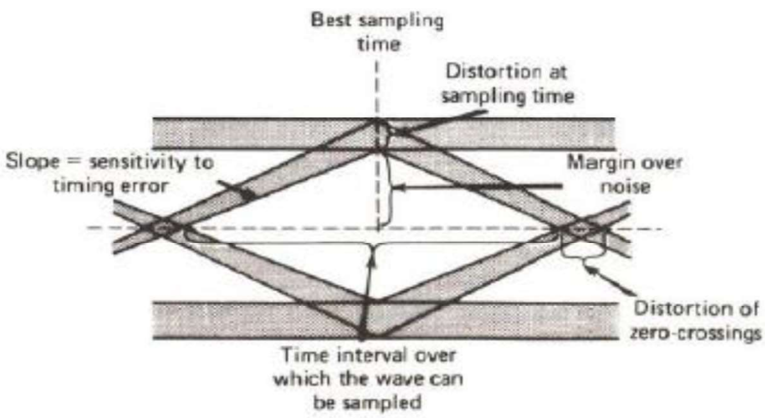
Or

(b).With neat diagram ,Explain Eye pattern.

→ The quality of digital transmission systems are evaluated using the bit error rate.Degradation of quality occurs in each process modulation, transmission, and detection.
→ The eye pattern is experimental method that contains all the information concerning thedegradation of quality.
→ Therefore, careful analysis of the eye pattern is important inanalyzing the degradation mechanism.
→ Eye patterns can be observed using an oscilloscope.
→ The received wave is appliedto the vertical deflection plates of an oscilloscope and the sawtooth wave at a rateequal to transmitted symbol rate is applied to the horizontal deflection plates,resulting display is eye pattern as it resembles human eye.
→ The interior region of eye pattern is called eye opening



We get superposition of successive symbol intervals to produce eye pattern as shown below.

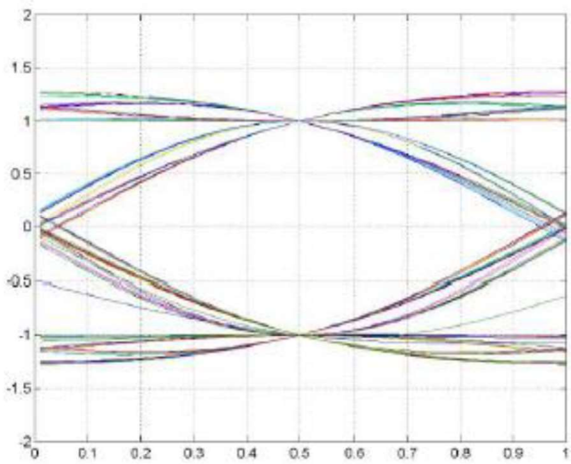


Interpretation of eye pattern

→ The width of the eye opening defines the time interval over which the received wave can be sampled without error from ISI.
→ The optimum sampling time corresponds to the maximum eye opening
→ The height of the eye opening at a specified sampling time is a measure of the margin over channel noise.
→ The sensitivity of the system to timing error is determined by the rate of closure of the eye as the sampling time is varied.
→ Any non linear transmission distortion would reveal itself in an asymmetric or squinted eye.
→ When the effected of ISI is excessive, traces from the upper portion of the eyepattern cross traces from lower portion with the result that the eye is completely closed.

Example of eye pattern:

Binary-PAM Perfect channel (no noise and no ISI)



Example of eye pattern: Binary-PAM with noise no ISI

Equalizer

10.(a).Explain in detail about Equalizer.

Or

(b).With neat diagram explain the adaptive equalization for data transmission.

This technique is another approach to minimize signal distortion in the base band data transmission. This is Nyquist third method for controlling

ISI. Equalization is essential for high speed data transmission over voice grade telephone channel which is essentially linear and band limited.

High speed data transmission involves two basic operations

i) Discrete pulse amplitude modulation:

The amplitudes of successive pulses in a periodic pulse train are varied in a discrete fashion in accordance with incoming data stream.

ii) Linear modulation

Which offers band width conservation to transmit the encoded pulse train over telephone channel.

At the receiving end of the systems, the received waves is demodulated and then synchronously sampled and quantized. As a result of dispersion of the pulse shape by the channel the number of detectable amplitude levels is limited by ISI rather than by additive noise. If the channel is known , then it is possible to make ISI arbitrarily small by designing suitable pair of transmitting and receiving filters for pulse shaping.

In switched telephone networks we find that two factors contribute to pulse distortion.

1. Differences in the transmission characteristics of individual links that may be switched together.

2. Differences in number of links in a connection

Because of these two characteristics, telephone channel is random in nature. To realize the full transmission capability of a telephone channel we need adaptive equalization.

Adaptive equalization

• An equalizer is a filter that compensates for the dispersion effects of a channel.

Adaptive equalizer can adjust its coefficients continuously during the transmission of data.

Pre channel equalization
-requires feed back channel
-causes burden on transmission.

Post channel equalization
Achieved prior to data transmission by training the filter with the guidance of a
training sequence transmitted through the channel so as to adjust the filter parameters to optimum values.
Adaptive equalization – It consists of tapped delay line filter with set of delay elements, set of adjustable multipliers connected to the delay line taps and a summer for adding multiplier outputs.

The output of the Adaptive equalizer is given by

$$y(nt)=\sum_{i=0}^{M-1} c_i\, x(nT-iT)$$

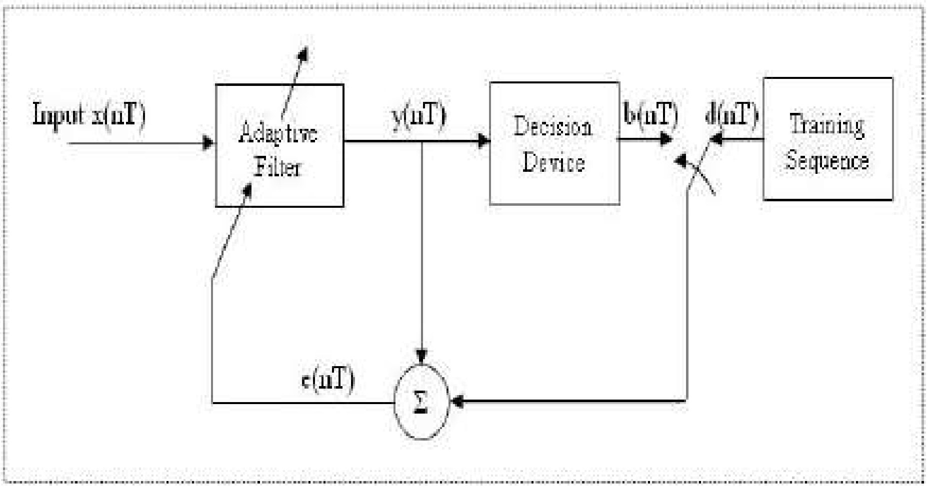
C_i is weight of the i^{th} tap Total number of taps are M .Tap spacing is equal to symbol duration T of transmitted signal

In a conventional FIR filter the tap weights are constant and particular designed response is obtained. In the adaptive equaliser the C_i's are variable and are adjusted by an algorithm

Two modes of operation

1. Training mode 2 . Decision directed mode

Mechanism of adaptation



Training mode

A known sequence d(nT) is transmitted and synchronized version of it is generated in the receiver applied to adaptive equalizer.This training sequence has maximal length PN Sequence, because it has large average power and large SNR, resulting response sequence (Impulse) is observed by measuring the filter outputs at the sampling instants.The difference between resulting response y(nT) and desired response d(nT)is error signal which is used to estimate the direction in which the coefficients of filter are to be optimized using algorithms

Methods of implementing adaptive equalizer

i) Analog

ii) Hard wired digital
iii) Programmable digital

Analog method

i)Charge coupled devices [CCD's] are used.

ii)CCD- FET's are connected in series with drains capacitively coupled to gate.The set of adjustable tap widths are stored in digital memory locations, and the multiplications of the analog sample values by the digitized tap weights done in
analog manner.
Suitable where symbol rate is too high for digital implementation.

Hard wired digital technique
• Equalizer input is first sampled and then quantized in to form that is suitable for
storage in shift registers.
• Set of adjustable lap weights are also stored in shift registers. Logic circuits are
used for required digital arithmetic operations.
• widely used technique of equalization

Programmable method
• Digital processor is used which provide more flexibility in adaptation by programming.
• Advantage of this technique is same hardware may be timeshared to perform a
multiplicity of signal processing functions such as filtering, modulation and
demodulation in modem.

Pulse Shaping

11.(a).Write short notes on pulse shaping.

Or

(b).Discuss in detail about the need for pulse shaping.

Pulse shaping is the process of changing the waveform of transmitted pulses.Its purpose is to make the transmitted signal better suited to its purpose or communication channel,typically by limiting the effective bandwidth of the transmission.By fitering the transmitted pulses this way,the intersymbol interference caused by the channel can be kept in control.

In radio frequency communication,pulse shaping is essential for making the signal fit in its frequency band.Typically pulse shaping occurs after line coding and modulation.

Need for pulse Shaping:

Transmitting a signal at high modulation rate through a bandlimited channel can create intersymbol interference.As the modulation rate increases,the signals bandwidth increases.

When the signals bandwidth becomes larger than the channel bandwidth,the channel starts to introduce distortion to the signal.This distortion usually manifests itself as intersymbol interference

The signals spectrum is determined by the modulation scheme and data rate used by the transmitter,but can be modified with a pulse shaping filter. Usually transmitted symbols are represented as a time sequence of dirac delta pulses.This thoritical signal is then filtered with thepulse shaping filter producing the transmitted signal

In many baseband Communication systems,the pulse shaping filter is implicitly a box car filter.The channel for the signal is bandlimited.Therefore better filters have been developed,which attempt to minimize the bandwidth needed for a certain symbol rate.

Pulse Shaping filter:

Not every filter can be used as a pulse shaping filter.The filter itself must not introduce intersymbol interference it needs to satisfy certain criteria.

The nyquist ISI criterion is a commonly used criterion for evaluation,because it relates the frequency spectrum of the transmitter signal to intersymbol interference.Examples of pule shaping fiters that are commonly found in communication systems are

(i)Sin C shaped filter

(ii)Raised Cosine filter

(iii)Gaussian filter.

Cosine Filter

12.(a).write short notes about Cosine filter.

Or

(b).Write the need for cosine filter.

A particularly useful class of filters is called raised cosine filters by the data communications industry. They are called sine squared filters by the television engineering community. The equivalence is seen in the following trigonometric relationship:

2sin2a=1-cos2a

The left side of the equation shows the rationale behind the television usage of sine squared terminology, while 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 analog 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 between the modulation 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.

Two Marks

1.What is meant by DPSK?

In DPSK, the input sequence is modified. Let input sequence be d(t) and output Sequence be b(t). Sequence b(t) changes level at the beginning of each interval in which d(t)=1 and it does not changes level when d(t)=0.

2.What is Signal constellation diagram?

Suppose that in each time slot of duration T seconds, one s2(t), . . (t) is transmitted with equal probability, 1/M For geometric representation, the signal si (t), = 1, 2, ..., M, is applied to a bank of correlators. The correlator outputs define the signal vector si. The set of message points corresponding to the set of transmitted signals {si(t))} i=1..M is called a signal constellation.

3.What is the purpose of using an eye pattern?

Eye pattern can be used for :
i) To determine an interval over which the received wave can be sampled without error due to ISI.
ii) To determine the sensitivity of the system to timing error.
iii) The margin over the noise is determined from eye pattern.

4.Why do you need adaptive equalization in a switched telephone network.

In switched telephone network the distortion depends upon
i) Transmission characteristics of individual links.
ii) Number of links in connection.
Hence fixed pair of transmit and receive filters will not serve the equalization problem. The transmission characteristics keep on changing. Therefore adaptive equalization is used