Objectives:

- Need of Line Coding
- Introduction of Line Coding
- Properties of Line Coding
- Types of Line Coding
- Advantages and Disadvantages
- Power Spectral Density
- PSD of Line Coding
- Comparison of Line Coding

Need Of Line Coding:

- Various Techniques
- Other Way: From Computers
- Information: Inherently discrete in nature
- Transmitted over band-limited channel: Signal gets Dispersed
- Causes: Overlap and Distortion
- Distortion: Intersymbol Interference(ISI)

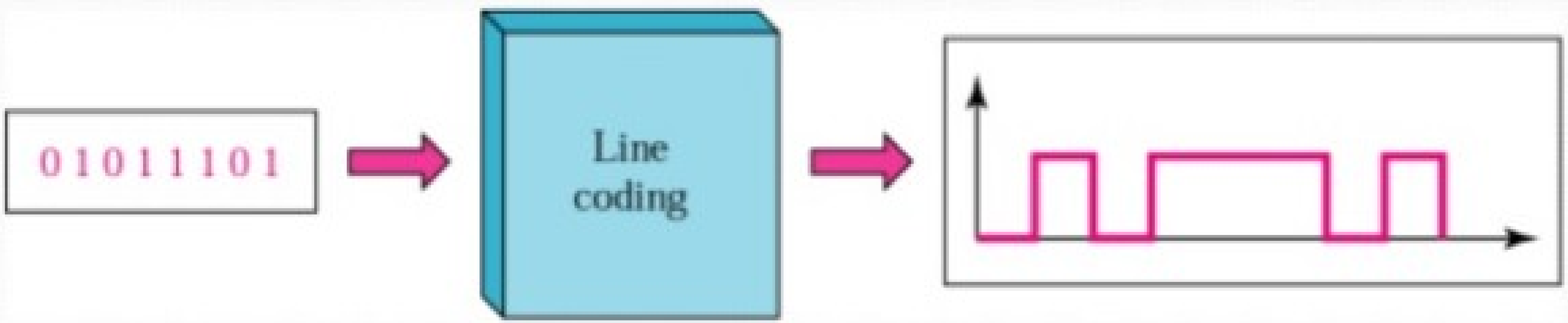
To avoid all these problems we are going for

Line Coding

Introduction:

- Binary Data: Pulses

- Line Coding: A pair of pulses to represent symbols 1 and 0



transmitted, and is ...

2.1.1 Desirable Properties of Line Codes

The line codes often use the terminology non-return-to-zero (NRZ) or return-to-zero (RZ). As the name suggests,

- *Non-Return-to-Zero (NRZ)* indicates that the binary pulse used to represent the bit does not necessarily return to zero or neutral level during the bit period.
- *Return-to-Zero (RZ)* implies that the binary pulse used to represent the bit always returns to zero or neutral level usually at the middle of the bit period.

Before we discuss different types of line-coding techniques, we address their common characteristics and desirable properties of a good line code for a specified application. Upon reading this subsection, you will be able to answer how to optimize the performance of digital data transmission.

-able
f Line

There are certain desirable properties which must be considered for line coding.

- Transmission power efficiency
- Duty cycle
- The dc components
- Baseline wandering
- Bandwidth considerations
- Self-clocking capability or self-synchronization
- Immunity to noise and interference
- Error detection capability
- Ease of detection and decoding

Transmission power efficiency, or transmission voltage levels, can be categorized as either unipolar (UP), or polar, as given below:

- In *unipolar voltage levels*, only one nonzero voltage level is specified. For example, positive voltage level is used to represent binary data 1 and zero (ground) voltage level is used to represent binary data 0.
- In *polar voltage levels*, two distinct nonzero symmetrical but opposite voltage levels are specified. For example, positive voltage level is used to represent binary data 1 and negative voltage level is used to represent binary data 0.

Proper Transmis Power Effi

Dear student... *After reading this material, you should be able to answer how power efficiency can be increased by using polar voltage levels. This is illustrated with the help of the following example.*

SOLVED EXAMPLE 2.1.1 **Transmission Power Efficiency—Polar *versus* Unipolar**

Show that there is 200% improvement in power efficiency in case of a polar line code using symmetrical voltage levels as compared to that of a unipolar line code for the same voltage level. Assume a load resistor of 1 Ω.

Solution

Case I: Unipolar Line-Coding Waveform

Step 1: Let binary data 1 be represented by +5 V (peak) and binary data 0 is represented by 0 V. [Assume]

Step 2: Therefore, average power in case of unipolar waveform,

$$P_{av}(\text{UP}) = \frac{V_{rms}^2(1)}{R} + \frac{V_{rms}^2(0)}{R} = \frac{1}{R}\left[V_{rms}^2(1) + V_{rms}^2(0)\right]$$

Substituting the values as $R = 1\ \Omega$, $V_{rms} = V_p/\sqrt{2}$; we have

$$P_{av}(\text{UP}) = \frac{1}{1\ \Omega}\left[\left(\frac{5}{\sqrt{2}}\right)^2 + 0\right] = 12.5\ \text{W}$$ [Compute it carefully]

Case II: Polar Symmetrical Line-Coding Waveform

Step 3: Let binary data 1 be represented by +5 V (peak) and binary data 0 is represented by –5 V (peak). [Assume]

Step 4: Therefore, average power in case of polar waveform,

$$P_{av}(\text{UP}) = \frac{V_{rms}^2(1)}{R} + \frac{V_{rms}^2(0)}{R} = \frac{1}{R}\left[V_{rms}^2(1) + V_{rms}^2(0)\right]$$

Substituting the values as $R = 1\ \Omega$, $V_{rms} = V_p/\sqrt{2}$; we have

$$P_{av}(\text{Polar}) = \frac{1}{1\ \Omega}\left[\left(\frac{+5}{\sqrt{2}}\right)^2 + \left(\frac{-5}{\sqrt{2}}\right)^2\right] = 25\ \text{W}$$ [Compute it carefully]

Step 5: Improvement in Transmission Power Efficiency

$$\frac{P_{av}(\text{Polar})}{P_{av}(\text{UP})} \times 100 = \frac{25}{12.5} \times 100 = 200\%$$ **[Hence, Proved]**

Duty cycle is defined as the ratio of the bit duration for which the binary pulse has defined transmission voltage to the entire bit duration.

For example,

- In **non-return-to-zero (NRZ) line-encoding format**, the binary pulse is maintained high for binary data 1 for the entire bit duration, or the binary pulse is maintained low for binary data 0 for the entire bit duration. So, the duty cycle is 100%. We will see the details of NRZ line-encoding format in Section 2.1.2.
- In **return-to-zero (RZ) line-encoding format**, the binary pulse is maintained high for binary data 1 for 50% of the entire bit duration only, and the binary pulse is maintained low for binary data 0 for the entire bit duration. So, the average duty cycle is less than 100% of specified bit duration. We will see the details of RZ line-encoding format in Section 2.1.2.

Some communication systems like a telephone system (a telephone line cannot pass frequencies below 200 Hz), and a long-distance link using transformers (to isolate different parts of the line electrically) do not allow transmission of frequencies around zero, called direct-current (dc) components. This situation occurs when the voltage level in a digital signal is constant for a while, the frequency spectrum shows very low-frequency components. Therefore, we need a line-coding technique with no dc component. Figures 2.1.3 and 2.1.4 show a signal without dc component and with dc component respectively.

without Dc.

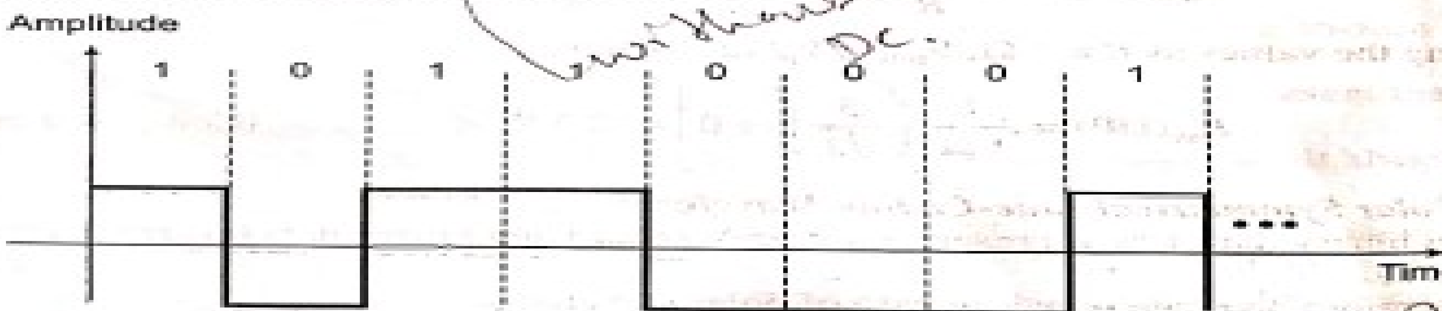


Figure 2.1.3 A Digital Signal without dc component

Zero DC component

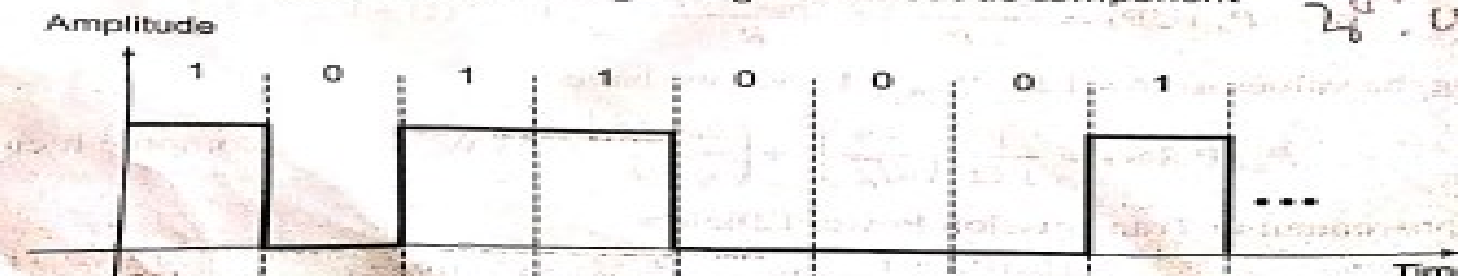


Figure 2.1.4 A Digital Signal with dc component

Property 4: Baseline Wandering

In digital transmission, a long sequence of either 1s or 0s produces a condition in which a receiver may lose its amplitude reference. This reference is needed for optimum determination of received 1s and 0s with clear-cut discrimination between them. Similar conditions may also arise when there is a significant imbalance in the number of 1s and 0s transmitted. This condition causes a drift in the baseline (a running average received signal power calculated by the receiver while decoding a digital signal), called *baseline wandering*. This makes it difficult for the receiver to decode received data correctly. Therefore, a good line-coding scheme needs to prevent baseline wandering.

Property 5: Bandwidth Consideration

Obviously, it is desirable that the bandwidth of a line code should be as small as possible. This allows more information to be transmitted per unit channel bandwidth.

Property 6: Self-clocking Capability or Se synchronizatic

A self-synchronizing digital signal includes timing information in the data being transmitted. This can be achieved if there are transitions in the line-encoded transmitted digital signal that alert the receiver to the beginning, middle, or end of the pulse waveform. If the receiver's clock is out of synchronization, these transitions can reset the clock. We will discuss later that there are some line coding techniques which exhibit self-clocking capability.[3]

Figure 2.1.5 shows the effect of lack of self-synchronization in which the receiver has a shorter bit duration.

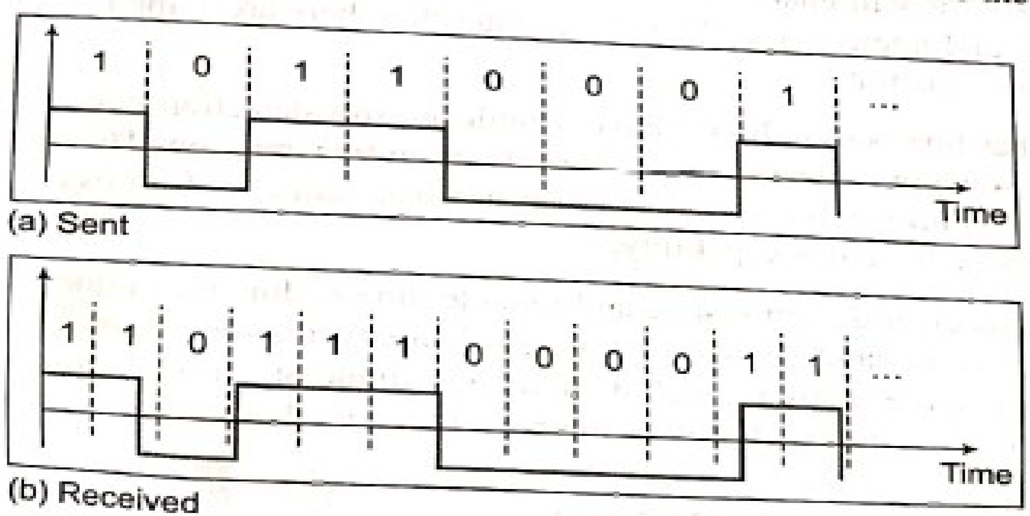


Figure 2.1.5 Need of Synchronization between Transmitted and Received Data

As it can be seen from the figure, the receiver receives extra bits (11011100000111) ... time duration...

sed to represent the binary
three types, as depicted in Figure 2.1.7.

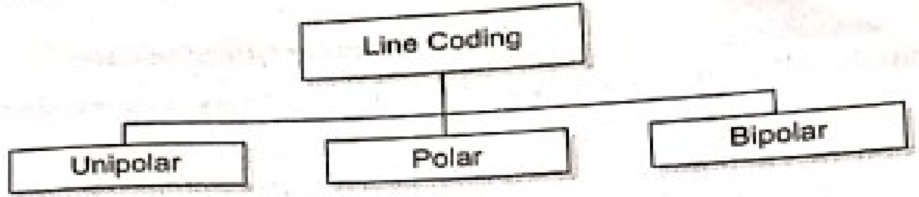


Figure 2.1.7 Classification of Line-Coding Techniques

Unipolar (represented by only one level +V or –V). Thus, Unipolar line coding can be Unipolar Non-Return-to-Zero (**UP-NRZ**) and Unipolar Return-to-Zero (**UP-RZ**).

Polar (represented by two distinct non-zero symmetrical but opposite voltage levels, +V and –V).[4]

Bipolar (also known as pseudo-ternary +V, –V, and 0 V or alternate mark inversion).

Polar line codes can be further classified as depicted in Figure 2.1.8.

Figure 2.1.8 Classification of Polar Line Codes

List of Types of Line Codes

Any one of several line coding techniques can be used for the electrical representation of a binary data stream:

- Unipolar Non-Return-to-Zero (NRZ) Line Code
- Unipolar Return-to-Zero (RZ) Line Code
- Polar Non-Return-to-Zero (NRZ) Line Code
- Manchester Polar Line Code
- Differential Manchester Polar Line Code
- Bipolar Non-Return-to-Zero Alternate Mark Inversion (BP-NRZ-AMI) Line Code
- Bipolar RZ Line Code
- Bipolar RZ-AMI Line Code
- High-Density Bipolar (HDB) NRZ-AMI Line Code
- Binary Eight Zeros Substitution (B8ZS) RZ-AMI Line Code

Now, we describe each one of these line codes briefly in order to develop better understanding.

Unipolar Signaling:

•On-Off keying ie OOK

•Pulse 0: Absence of pulse

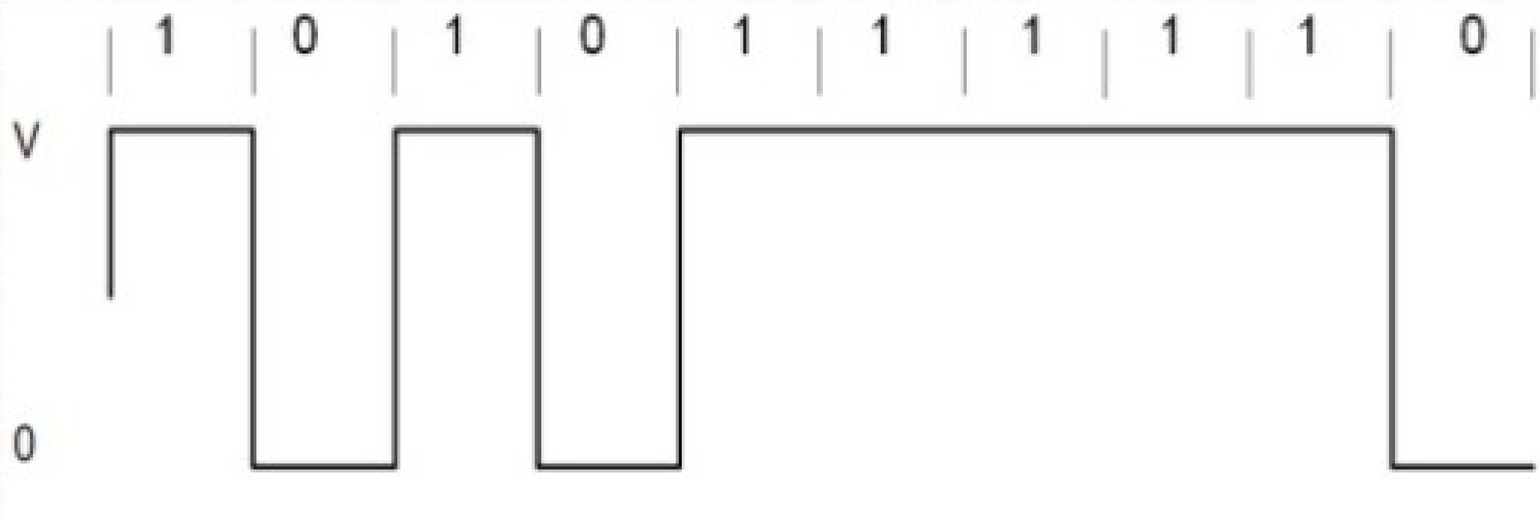
•Pulse1 : Presence of pulse

There are two common variations of unipolar signalling:

1. Non-Return to Zero (NRZ)
2. Return to Zero (RZ)

Unipolar Non-Return to Zero (NRZ):

•Duration of the MARK pulse (τ) is equal to the duration (T_o) of the symbol slot.



Advantages:

•Simplicity in implementation

• Doesn't require a lot of bandwidth for transmission.

Disadvantages:

•Presence of DC level (indicated by spectral line at 0 Hz).

•Contains low frequency components. Causes "Signal Droop"

•Does not have any error correction capability.

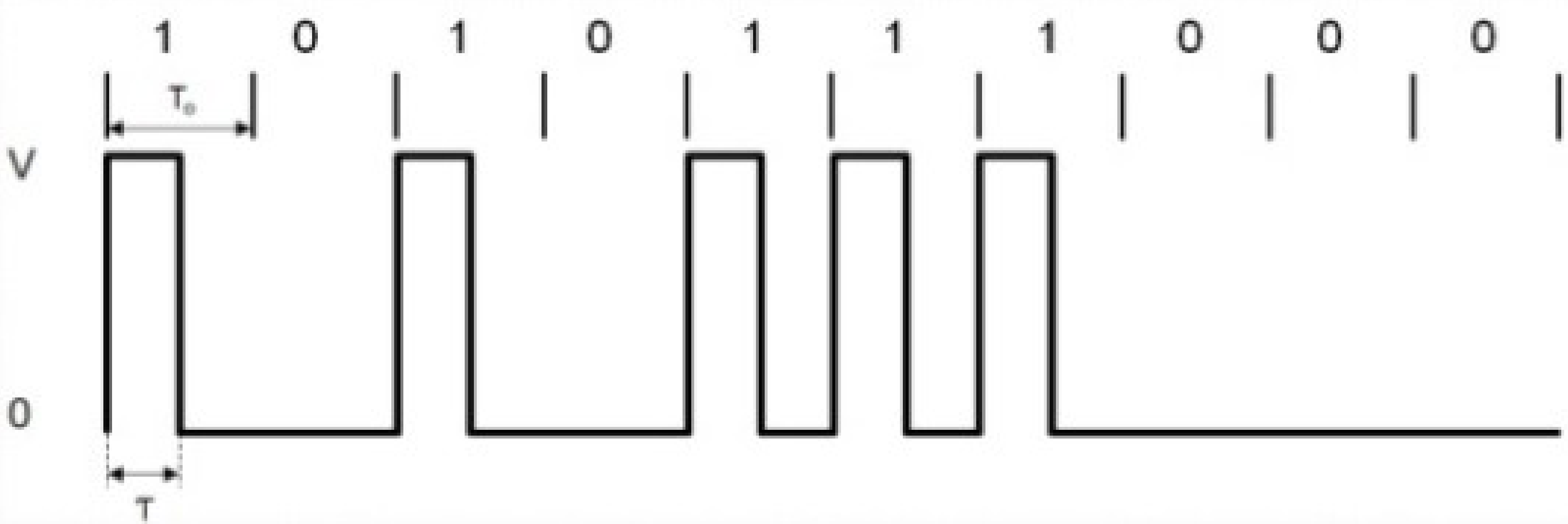
•Does not posses any clocking component for ease of synchronisation.

•Is not Transparent. Long string of zeros causes loss of synchronisation.

Unipolar Return to Zero (RZ):

•MARK pulse (τ) is **less** than the duration (T_o) of the symbol slot.

•Fills only the first half of the time slot, returning to zero for the second half.



Advantages:

- Simplicity in implementation.
- Presence of a spectral line at symbol rate which can be used as symbol timing clock signal.

Disadvantages:

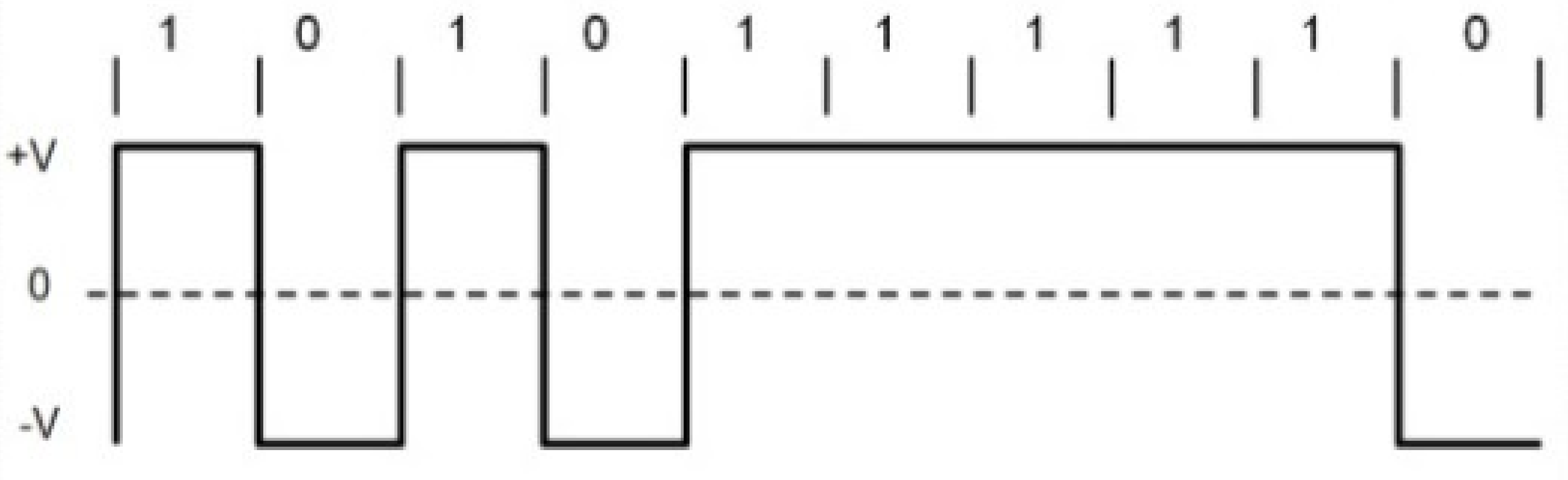
- Presence of DC level (indicated by spectral line at 0 Hz).
- Continuous part is non-zero at 0 Hz. Causes “Signal Droop”.
- Does not have any error correction capability.
- Occupies twice as much bandwidth as Unipolar NRZ.
- Is not Transparent

Polar Signalling:

- Polar RZ
- Polar NRZ

Polar NRZ:

- A binary 1 is represented by a pulse $g_1(t)$
- A binary 0 by the opposite (or antipodal) pulse $g_0(t) = -g_1(t)$.



Advantages:

•Simplicity in implementation.

•No DC component.

Disadvantages:

•Continuous part is non-zero at 0 Hz. Causes "Signal Droop".

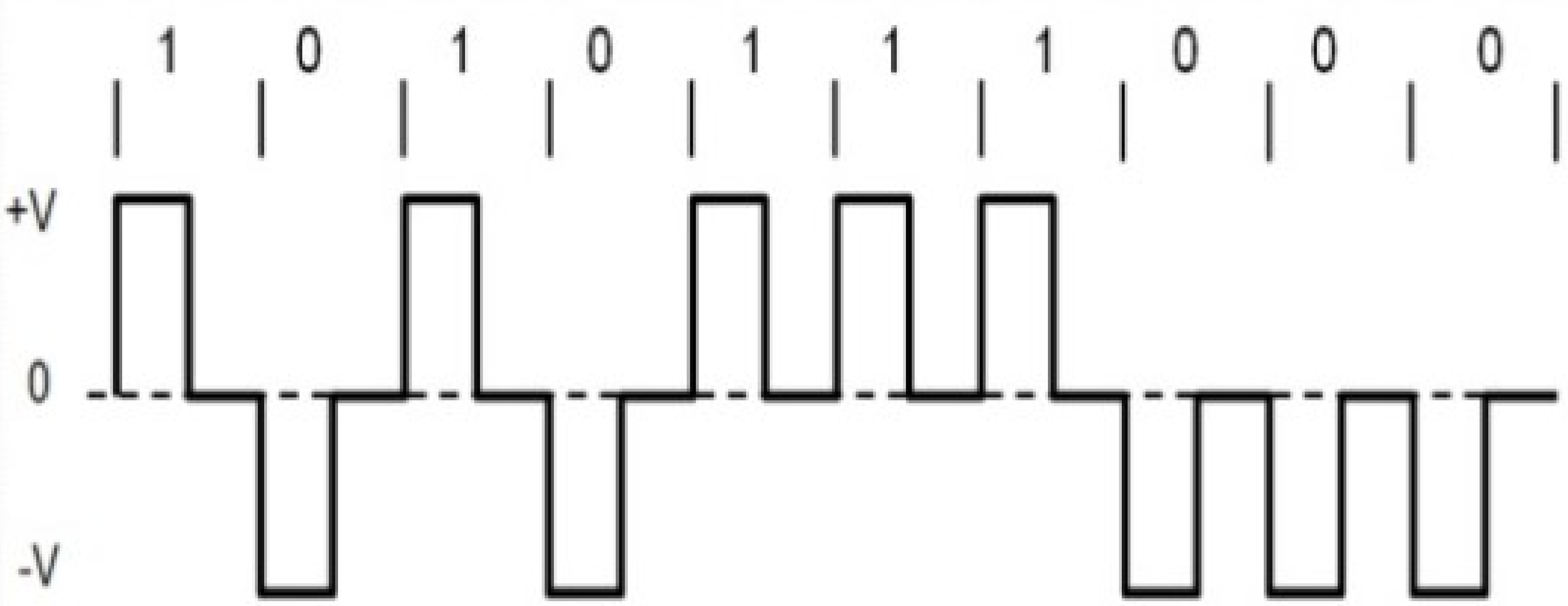
•Does not have any error correction capability.

•Does not posses any clocking component for ease of synchronisation.

•Is not transparent.

Polar RZ:

•A binary 1: A pulse $g_1(t)$
•A binary 0: The opposite (or antipodal) pulse $g_0(t) = -g_1(t)$.
•Fills only the first half of the time slot, returning to zero for the second half.



Advantages:

•Simplicity in implementation.

•No DC component.

Disadvantages:

•Continuous part is non-zero at 0 Hz. Causes "Signal Droop".

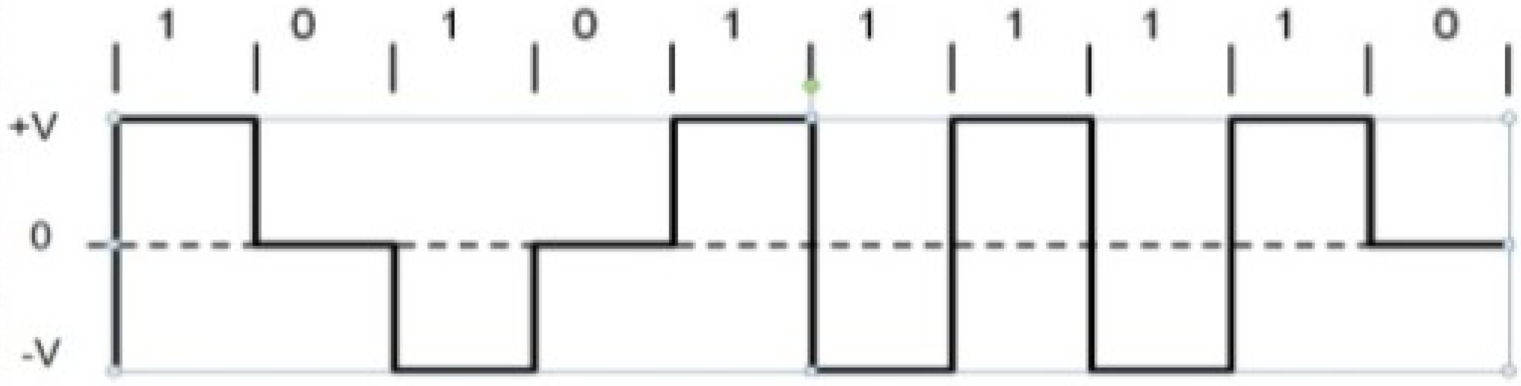
•Does not have any error correction capability.

•Occupies twice as much bandwidth as Polar NRZ.

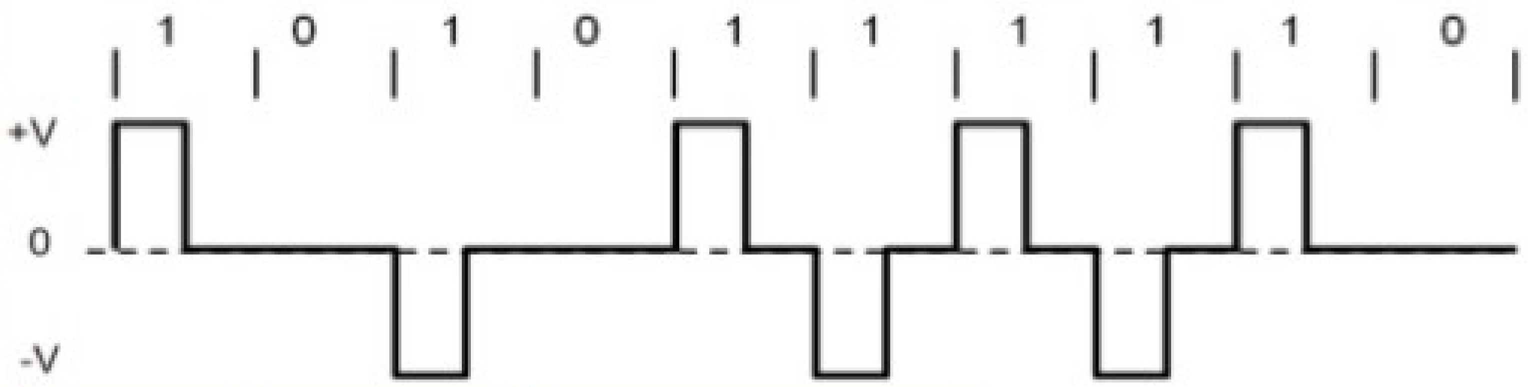
Bipolar Signalling:

- Alternate mark inversion (AMI)
- Uses three voltage levels (+V, 0, -V)
- 0: Absence of a pulse
- 1: Alternating voltage levels of +V and –V

Bipolar NRZ:



Bipolar RZ:



Advantages:

•No DC component.

•Occupies less bandwidth than unipolar and polar NRZ schemes.

•Does not suffer from signal droop (suitable for transmission over AC coupled lines).

•Possesses single error detection capability.

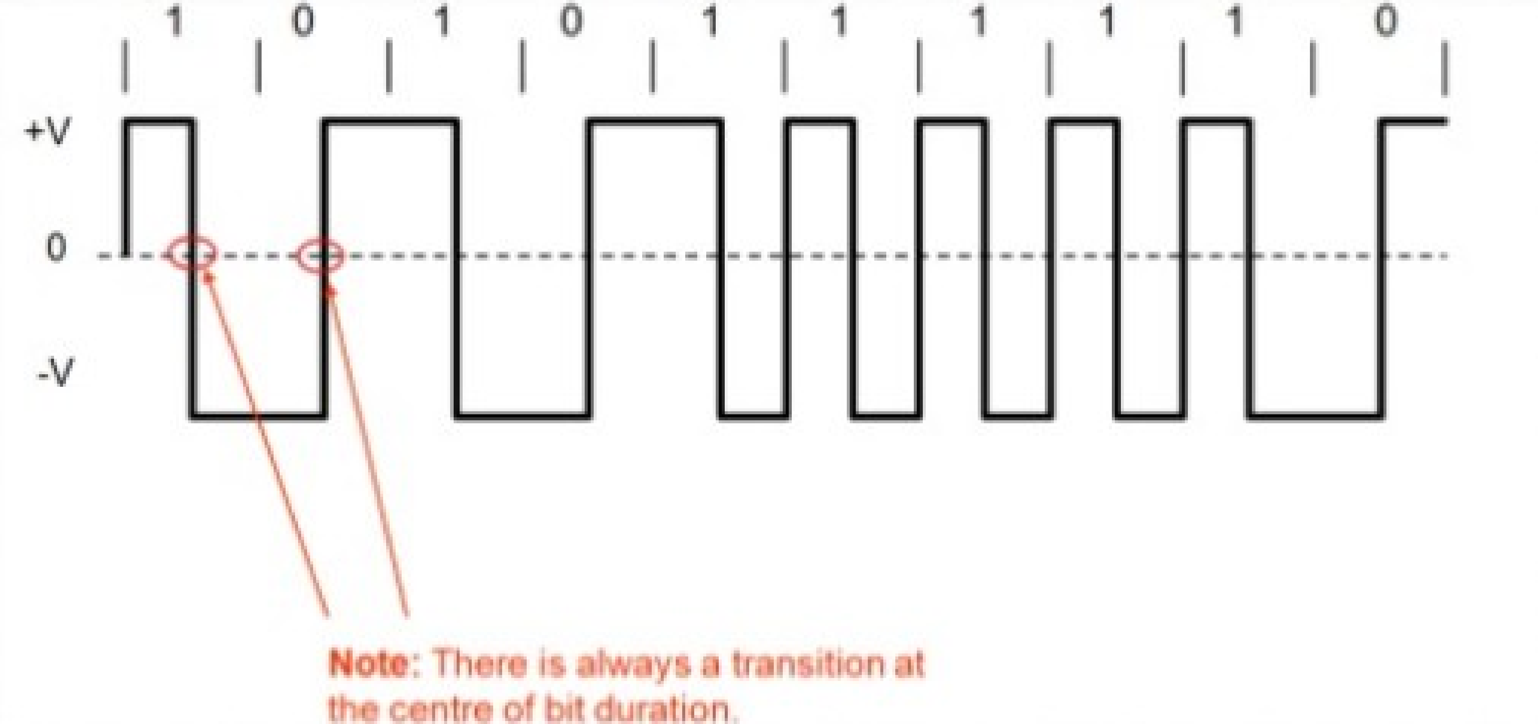
Disadvantages:

•Does not posses any clocking component for ease of synchronisation.

•Is not Transparent.

Manchester Signalling:

- The duration of the bit is divided into two halves
- A 'One' is +ve in 1^{st} half and -ve in 2^{nd} half.
- A 'Zero' is -ve in 1^{st} half and +ve in 2^{nd} half.



Advantages:

- No DC component.
- Does not suffer from signal droop (suitable for transmission over AC coupled lines).
- Easy to synchronise.
- Is Transparent.

Disadvantages:

- Because of the greater number of transitions it occupies a significantly large bandwidth.
- Does not have error detection capability.

In *Manchester polar line code*, a binary symbol 0 is represented by a $-V$ pulse during first half of the bit period followed by a $+V$ pulse during the second half of the bit period, and a binary symbol 1 is represented by a $+V$ pulse during the first half of the bit period followed by a $-V$ pulse during the second half of the bit period. Due to this reason, this type of line code is also known as *split phase* or *biphase*.

Mathematically, Manchester polar line coding waveform can be expressed as

For symbol '0';
$$v(t) = \begin{cases} -V \\ +V \end{cases} \quad \text{for} \begin{cases} 0 \le t < (T_b/2) \\ (T_b/2) \le t < T_b \end{cases}$$

For symbol '1';
$$v(t) = \begin{cases} +V \\ -V \end{cases} \quad \text{for} \begin{cases} 0 \le t < (T_b/2) \\ (T_b/2) \le t < T_b \end{cases}$$

Figure 2.1.12 depicts the Manchester polar line-coding waveform for the given binary data seque[nce] 0 1 0 0 1 1 1 0 1 0.

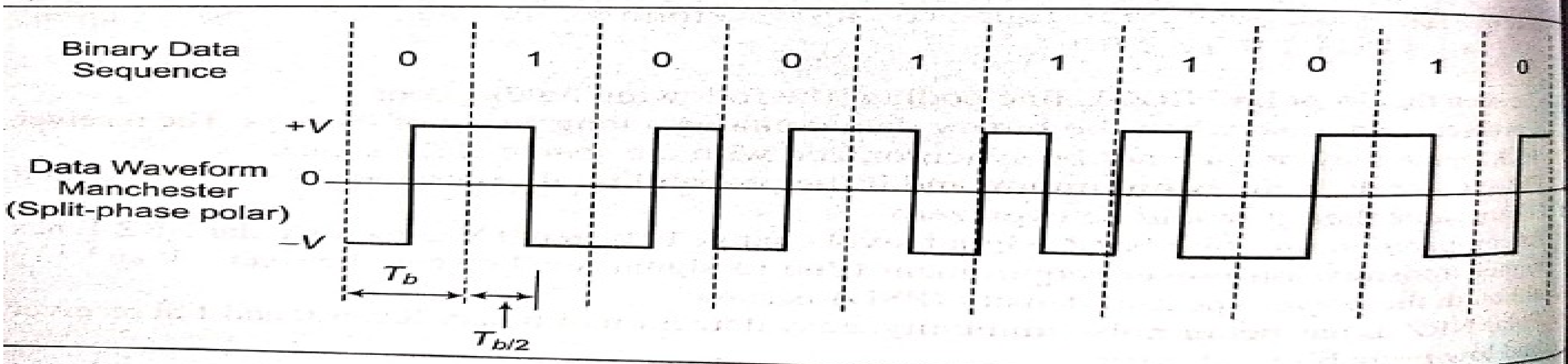


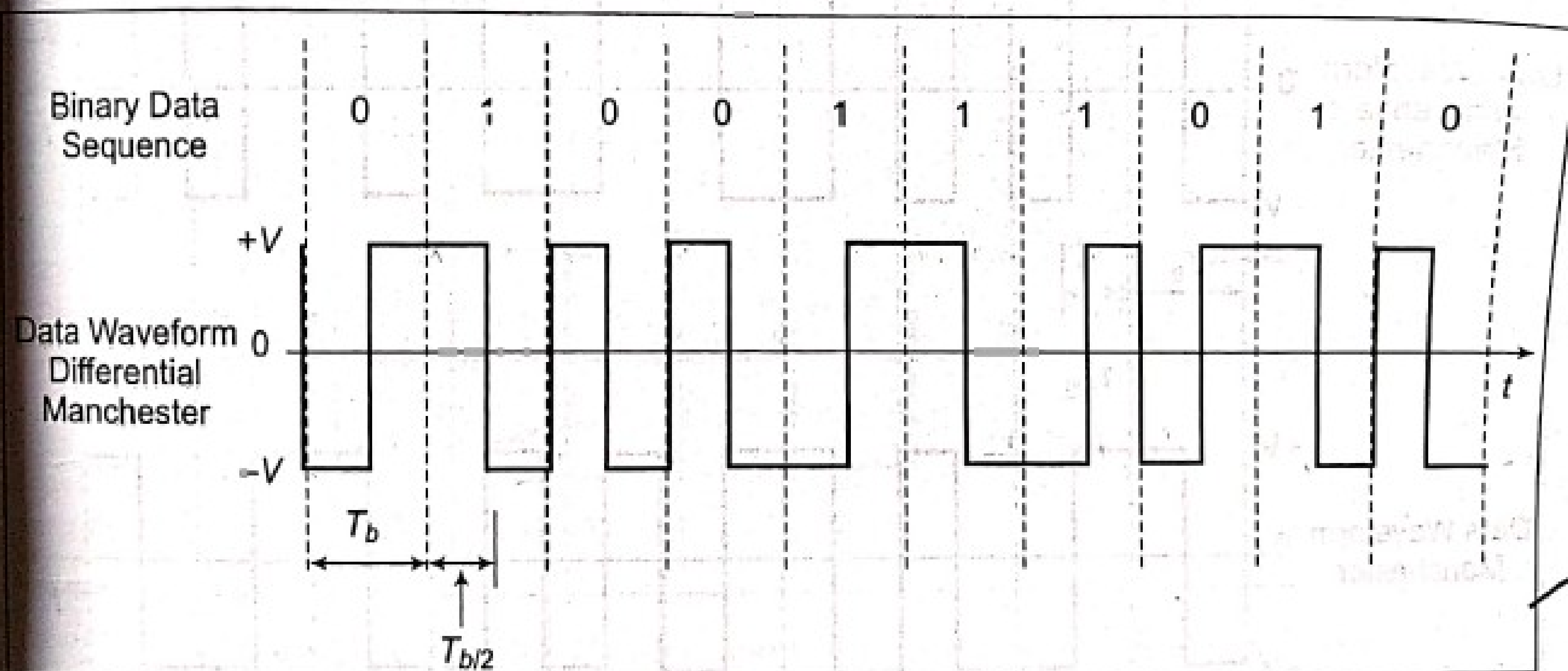
Figure 2.1.12 Manchester Polar Line-Coding Waveform for Example 2.1.10

SOLVED EXAMPLE 2.1.12 Differential Manchester Polar Line Code

Illustrate the differential Manchester polar line-coding waveform for the binary data sequence given as 0 1 0 0 1 1 1 0 1 0.

Solution We know that in the differential Manchester polar line code, the binary symbol 0 is represented by a transition at the beginning of the bit, followed by transition at the middle of the bit interval too, and the binary symbol 1 is represented by no transition at the beginning of the bit, but followed by transition at the middle of the bit interval.

Figure 2.1.13 depicts the differential Manchester polar line-coding waveform for the given binary data sequence 0 1 0 0 1 1 1 0 1 0.



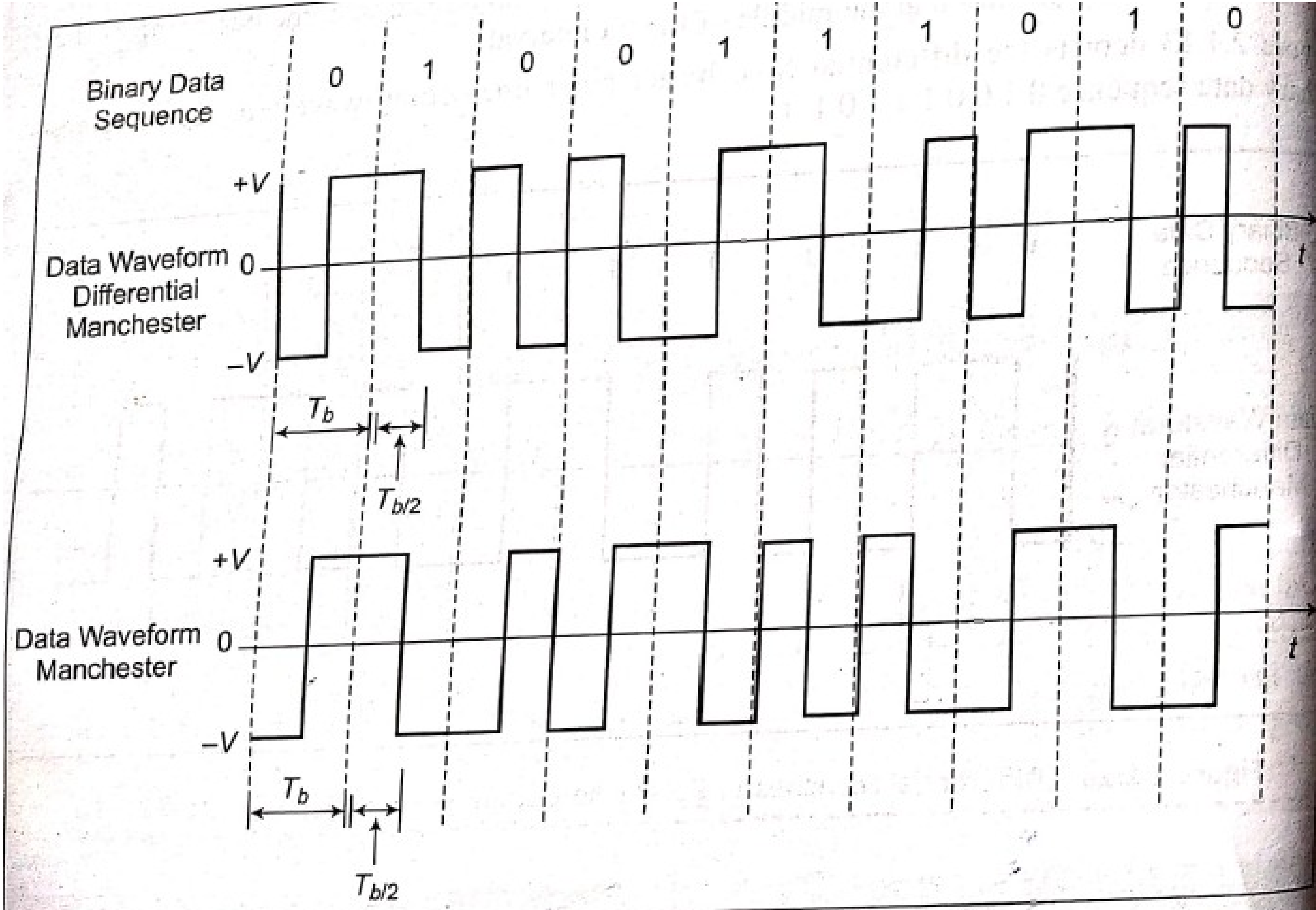


Figure 2.1.14 Manchester and Differential Manchester Line-Coding Waveforms

Illustrate the bipolar-NRZ-AMI line-coding waveform for the binary data sequence specified as 0 1 0 0 1 1 1 0 1 0.

Solution Figure 2.1.15 depicts the bipolar NRZ-AMI line-encoding waveform for the given binary data 0 1 0 0 1 1 1 0 1 0.

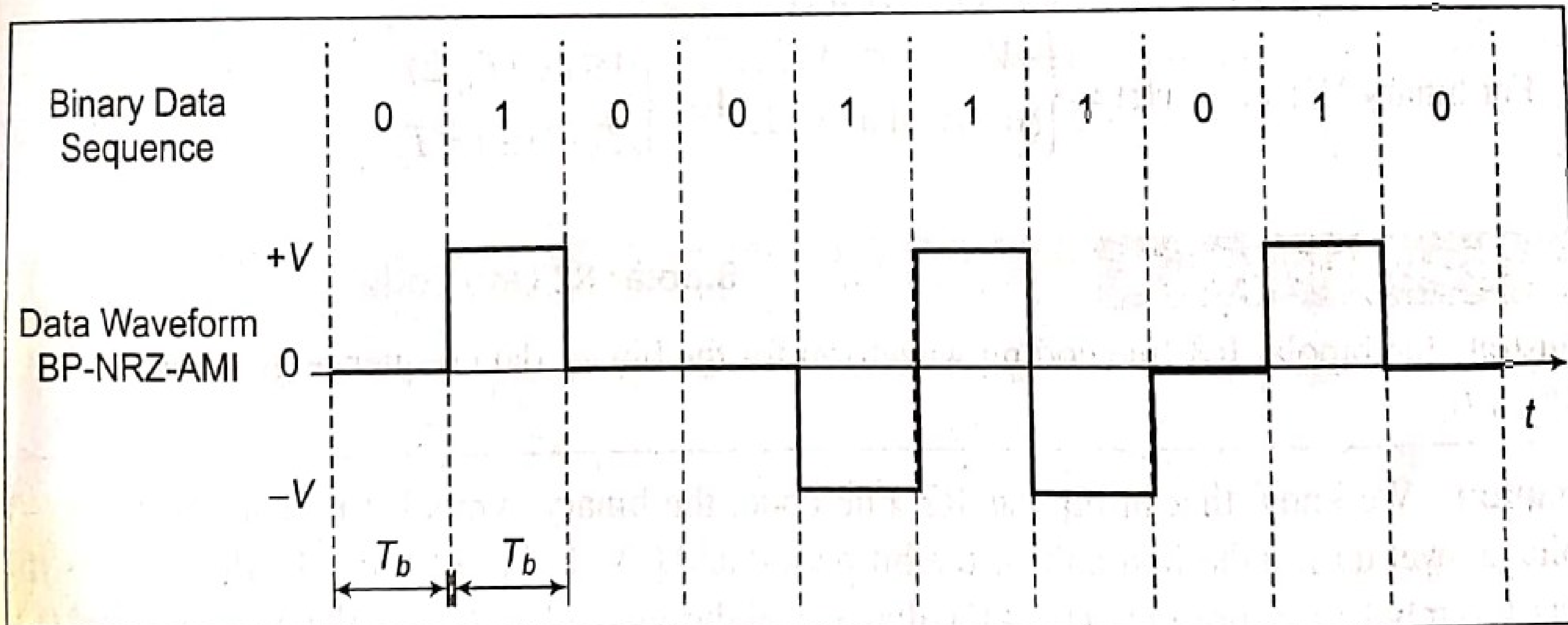


Figure 2.1.15 Bipolar NRZ-AMI Line-Coding Waveform for Example 2.1.16

period ($T_b/2$).

Mathematically, bipolar RZ-AMI waveform can be expressed as

For symbol '0'; $\quad v(t) = 0 \quad$ during $0 \le t < T_b$ interval

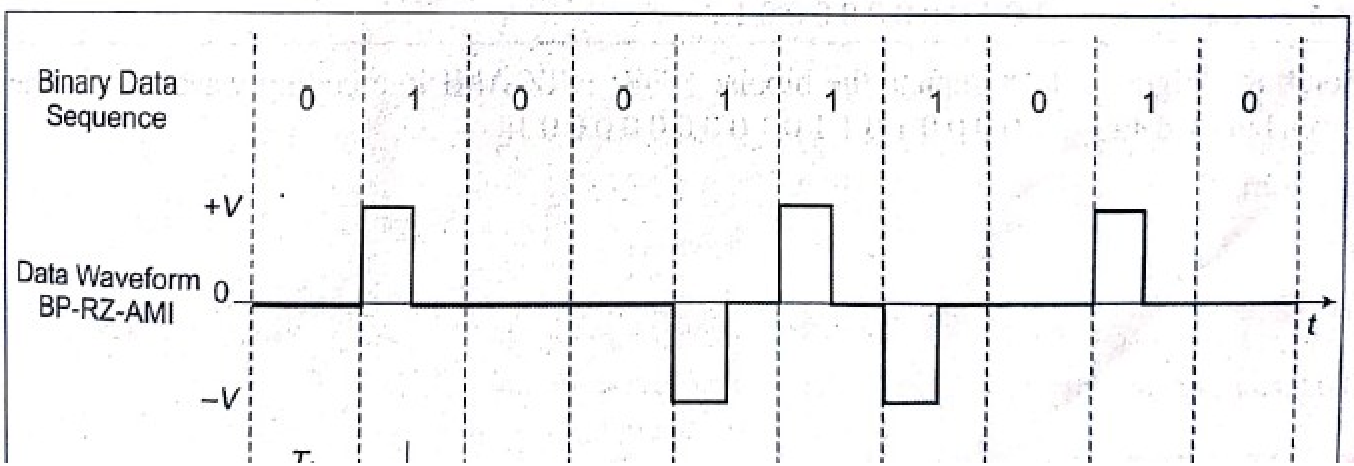
For symbol '1'; $\quad v(t) = \begin{cases} +V, \text{ or } -V \text{ alternatively} \\ 0 \end{cases} \quad$ for $\begin{cases} 0 \le t < (T_b/2) \\ (T_b/2) \le t < T_b \end{cases}$

SOLVED EXAMPLE 2.1.20 **Bipolar RZ AMI Line Code**

Illustrate the BP-RZ-AMI line coding waveform for the binary data sequence given as 0 1 0 0 1 1 1 0 1 0.

Solution We know that in the bipolar return-to-zero alternate-mark-inversion (BP-RZ-AMI) line code, the binary symbol 0 is represented by 0 V for the entire bit interval T_b, and the binary symbol 1 is represented by alternating pulses of $+V$ and $-V$ voltage levels which returns to zero (RZ) after one-half bit period ($T_b/2$).

Figure 2.1.17 depicts BP-RZ-AMI line-coding waveform for given binary data sequence 0 1 0 0 1 1 1 0 1 0.



...verting RZ-AMI to unipolar RZ.

Application ***Where does bipolar RZ-AMI line code find applications?***

BP-RZ-AMI line-coding technique is used in telephone systems as signaling scheme, and T-carrier lines with +3 V, 0 V, and –3 V voltage levels to represent binary data.

Type 9: High Density Bipolar (HDB) NRZ-AMI Line Code

Alternate mark-Inversion

Contrary to linear line codes described so far, High Density Bipolar (HDB) is a scrambling-type technique of line code, which provides synchronization without increasing the number of bits. In HDB-NRZ-AMI line coding,

- Some pre-defined number of pulses are added when the number of consecutive binary symbol 0s exceeds an integer value n.
- It is denoted as HDB_n, where $n = 1, 2, 3, ...$
- In HDB_n coding, when the input data sequence contains consecutive $(n + 1)$ zeros, this group of zeros is replaced by special $(n + 1)$ binary digit sequence.
- These special data sequences consist of some binary 1s so that they may be detected at the receiver reliably.

For example, when $n = 3$, the special binary sequences used are 000V and B 00V, where B (B stands for Bipolar) and V (V stands for Violation to AMI rule of encoding) are considered 1s. This means that four consecutive 0s are replaced with a sequence of either 000V or B00V, as explained in the following solved example.

SOLVED EXAMPLE 2.1.22 **Bipolar HDB_3 NRZ-AMI Line Code**

Illustrate the bipolar HDB_3 NRZ-AMI line-coding waveform for the binary data sequence given as 1 1 1 0 0 0 0 1 0 1 1 0 1 0 0 0 0 0 0 0 0 1.

Solution Figure 2.1.18 depicts the bipolar HDB_3 NRZ-AMI line coding waveform for the given binary data 1 1 1 0 0 0 0 1 0 1 1 0 1 0 0 0 0 0 0 0 0 1.

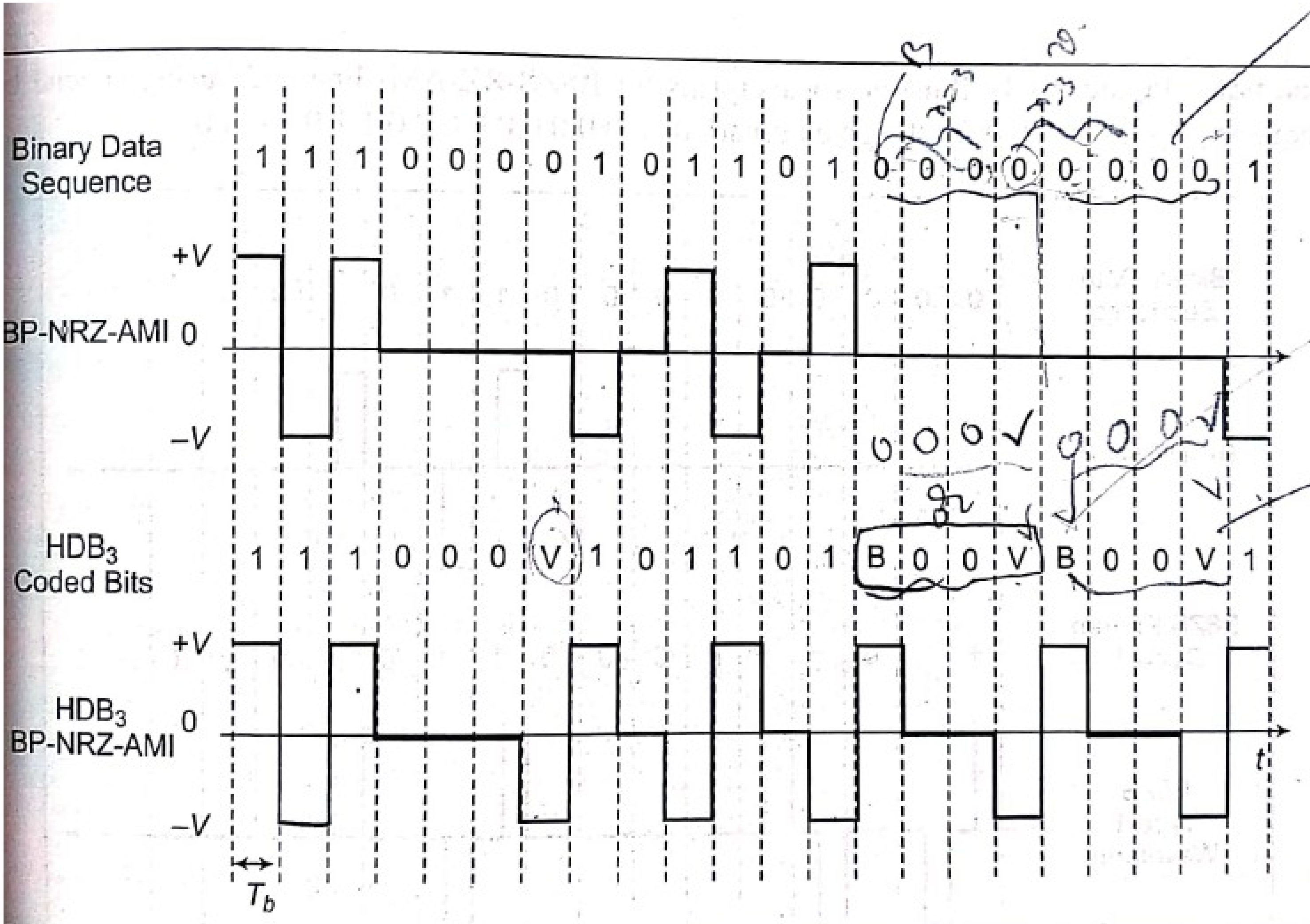


Figure 2.1.18 HDB$_3$ NRZ-AMI Line-Coding Waveform for Example 2.1.22

Just like HDB_3, B8ZS is another scrambling-type technique of line code, which provides synchronization without increasing the number of bits. In B8ZS-RZ-AMI line coding, whenever eight consecutive binary 0s appear in given binary data sequence, one of two special bit sequence – + – 0 – + 0 0 0 or – + 0 + – 0 0 0 (where + and – represent bipolar logic 1 conditions) is substituted for eight consecutive 0s.

Type 10: Binary Eight Zeros Substitution (B8ZS) RZ-AMI Line Code

SOLVED EXAMPLE 2.1.24 **B8ZS-RZ-AMI Line Code Type I**

Illustrate the B8ZS-RZ-AMI line-coding waveform for the binary data sequence given as 0 0 0 0 0 0 0 0 0 1 1 0 1 0 0 0. Use special bit sequence as + – 0 – + 0 0 0.

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[6]High Density Bipolar (HDB) NRZ-AMI line code is a scrambling-type technique of line code, which provides synchronization without increasing the number of bits. HDB_3 line code is used in E1 digital carrier systems.

Solution Figure 2.1.19 illustrates waveforms for B8ZS-RZ-AMI line code using special bit sequence + – 0 – + 0 0 0 for the given binary data 0 0 0 0 0 0 0 0 0 1 1 0 1 0 0 0.

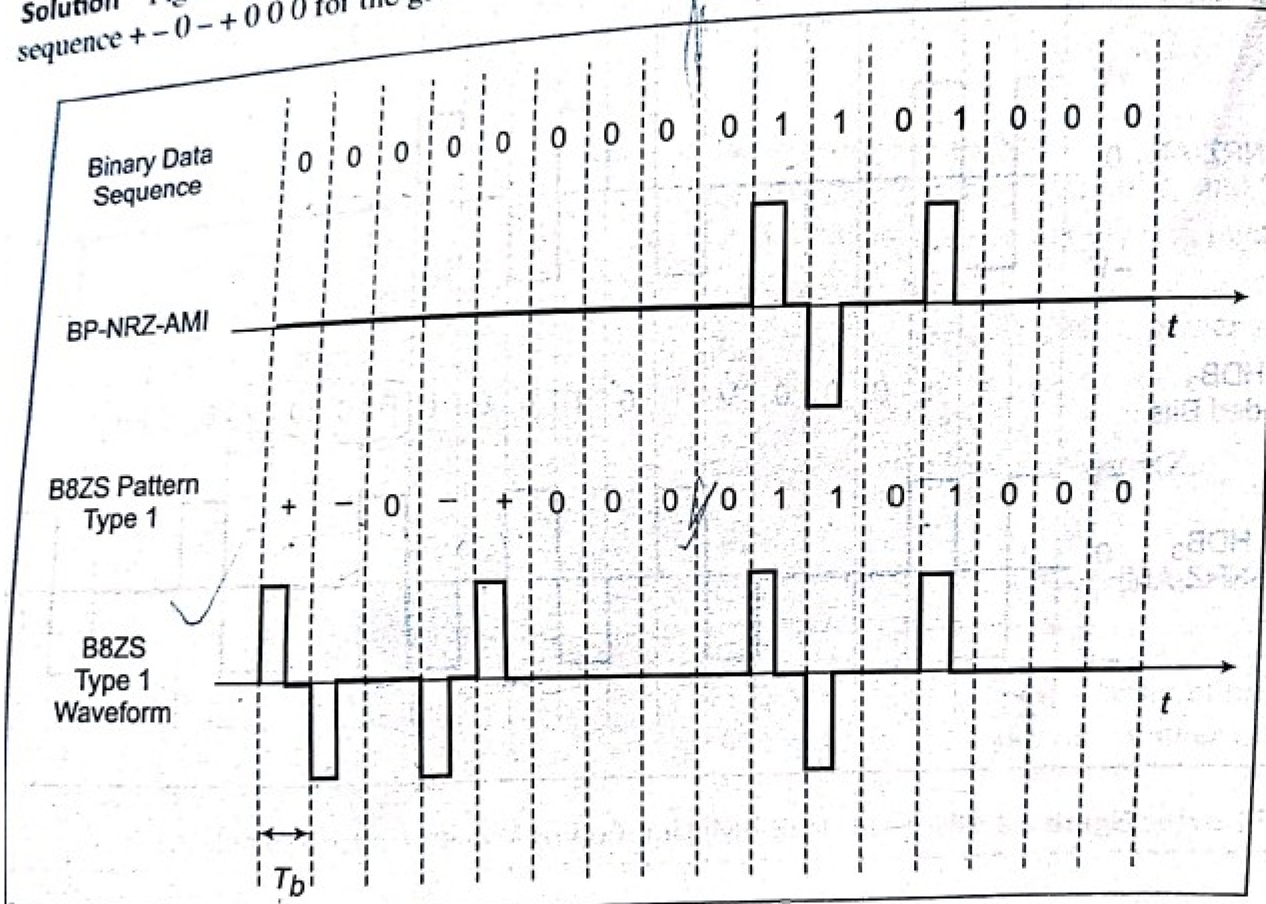


Figure 2.1.19 B8ZS Line-Coding Waveform for Example 2.1.24

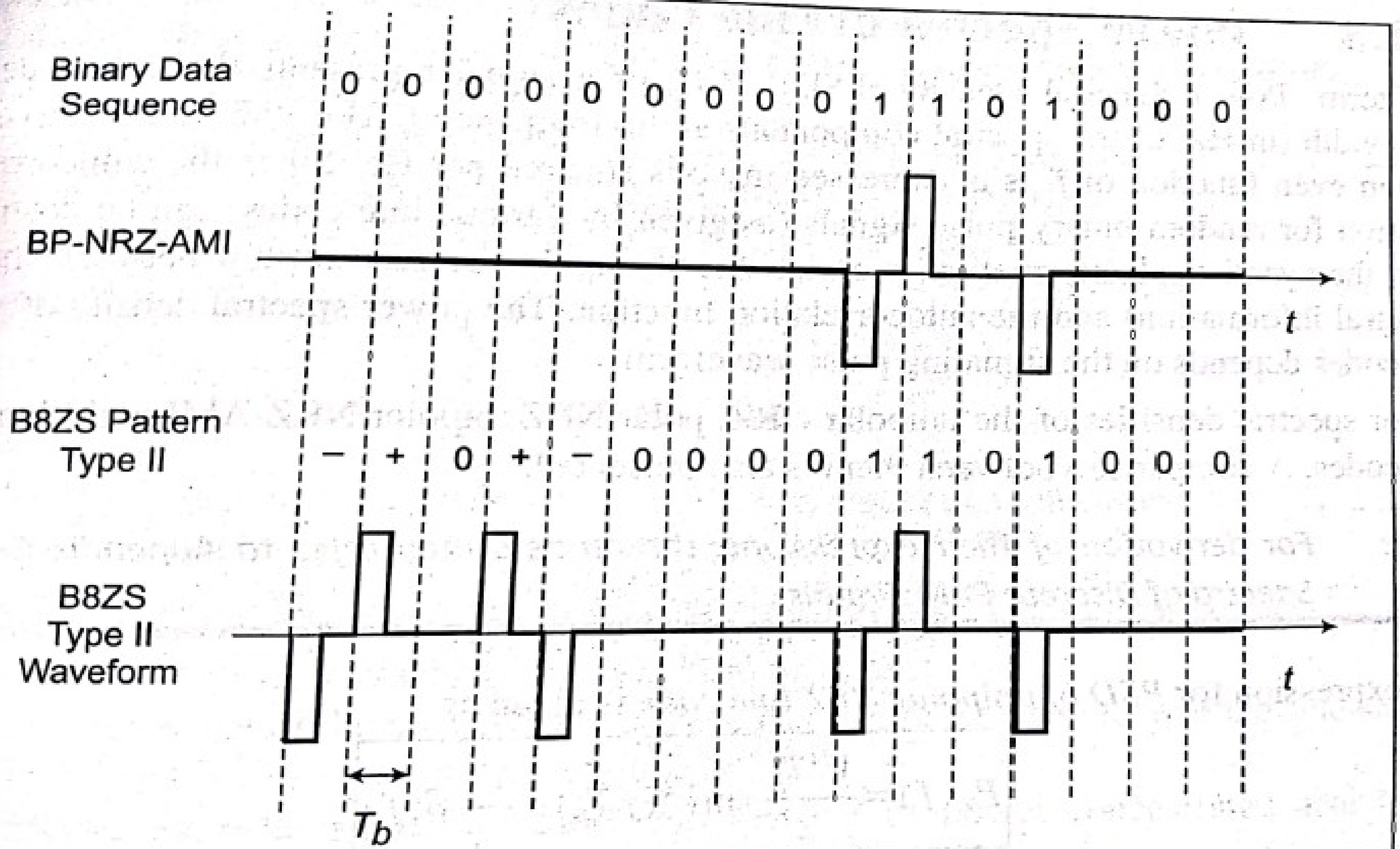


Figure 2.1.20 B8ZS Line-Coding Waveform for Example 2.1.26

The expression for *PSD of unipolar NRZ line code* is given as

$$\boxed{P_X(f) = \frac{V^2 T_b}{2}\,\text{sinc}^2(f T_b) + \frac{V^2}{4}\delta(f)}$$

It is observed that the presence of the Dirac delta function $\delta(f)$ in the second term accounts for one half of the power contained in the unipolar NRZ data format. Specifically, the power spectral density $P_X(f)$ is normalized with respect to $V^2 T_b$, and the frequency f is normalized with respect to the bit rate $1/T_b$.

Figure 2.1.21 shows the power spectral density of unipolar NRZ line-code waveform.

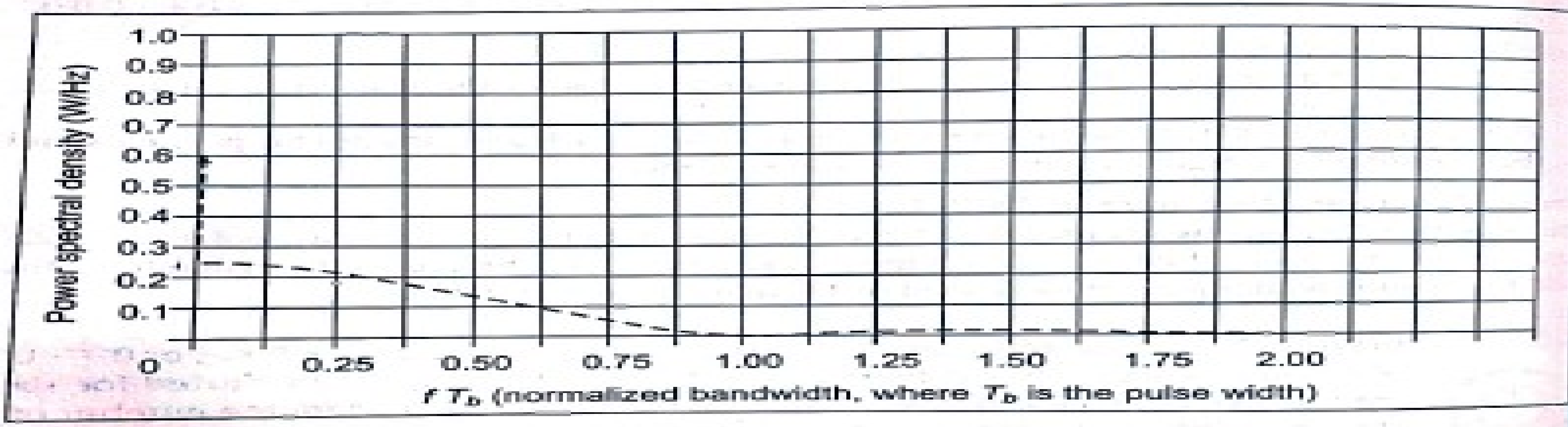


Figure 2.1.21 PSD of Unipolar NRZ Line Code

The PSD of unipolar NRZ line code has the following properties:

- The value of $P_X(f)$ is maximum at $f = 0$.
- The value of $P_X(f)$ is significant at all multiples of the bit rate $f_b = 1/T_b$.
- The peak value of $P_X(f)$ between f_b and $2f_b$ occurs at $f = 1.5 f_b$, and is 14 dB lower than the peak value of $P_X(f)$ at $f = 0$.

- The main lobe centered around $f = 0$ has 90% of the total power.
- When the NRZ signal is transmitted through an ideal low-pass filter with cut-off frequency at $f = f_b$, the total power is reduced by 10% only.
- The NRZ signal has no dc component.
- The power in the frequency range from $f = 0$ to $f = \pm \Delta f$ is $2G(f)\,\Delta f$.
- In the limit as Δf approaches to zero, the PSD becomes zero.

The expression for *PSD of Polar NRZ line code* is given as

PSD of Polar Line Cod

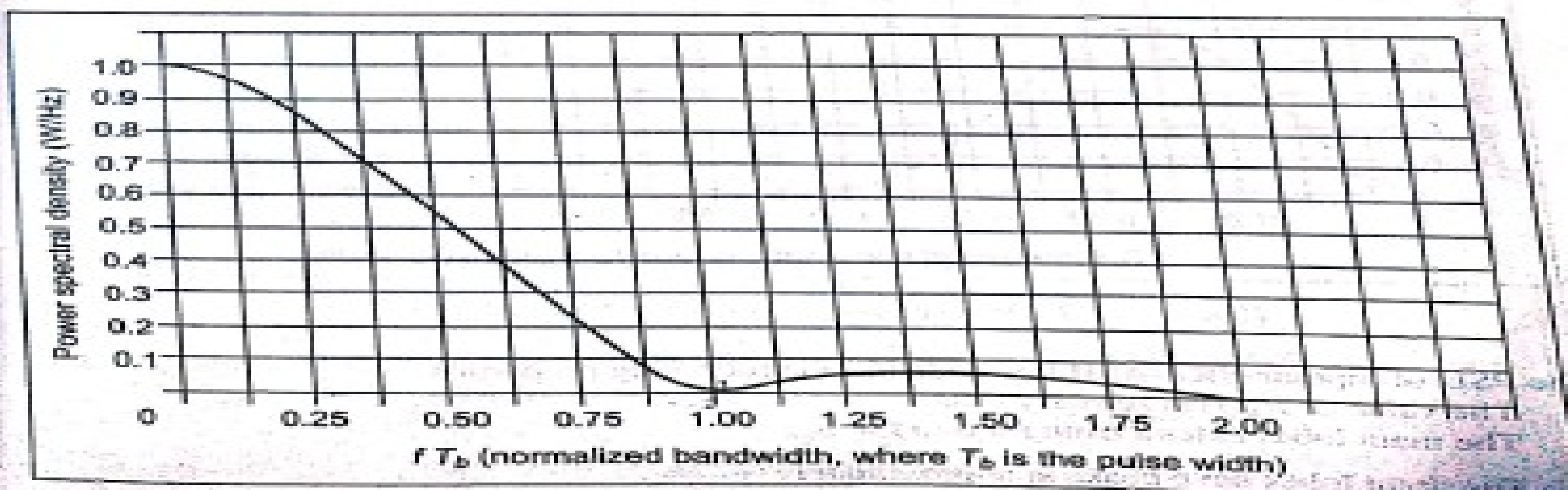
$$\boxed{P_X(f) = \frac{V^2 T_b}{2} \sin c^2 (f T_b)}$$

Specifically, the power spectral density $P_X(f)$ is normalized with respect to $V^2 T_b$, and the frequency f is normalized with respect to the bit rate $1/T_b$.

Figure 2.1.22 shows the power spectral density of polar NRZ line-code waveform.[8]

The PSD of Polar NRZ line code has the following properties:

- Most of the power of the polar NRZ line code lies within the main lobe of the sine-shaped curve, which extends up to the bit rate $1/T_b$.
- In general, the polar signaling format has most of its power concentrated near lower frequencies.
- Theoretically, the power spectrum becomes very small as frequency increases but never becomes zero above a certain frequency.
- The first non-dc null frequency, known as *essential bandwidth*, is R_b Hz where R_b is the clock frequency. The Nyquist bandwidth required to transmit R_b pulses per second is $R_b/2$.



This shows that the essential bandwidth of the bipolar NRZ-AMI as well as bipolar RZ signal is R_b, which is half that of the polar RZ signal and twice the Nyquist bandwidth ($R_b/2$). Although most of the power lies within a bandwidth equal to the bit rate $1/T_b$, the spectral content is relatively small around zero frequency or dc.

- Assume that the input binary data consisting of independent and equally likely symbols.
- The autocorrelation function of Manchester line code is same as that for the polar RZ line code.

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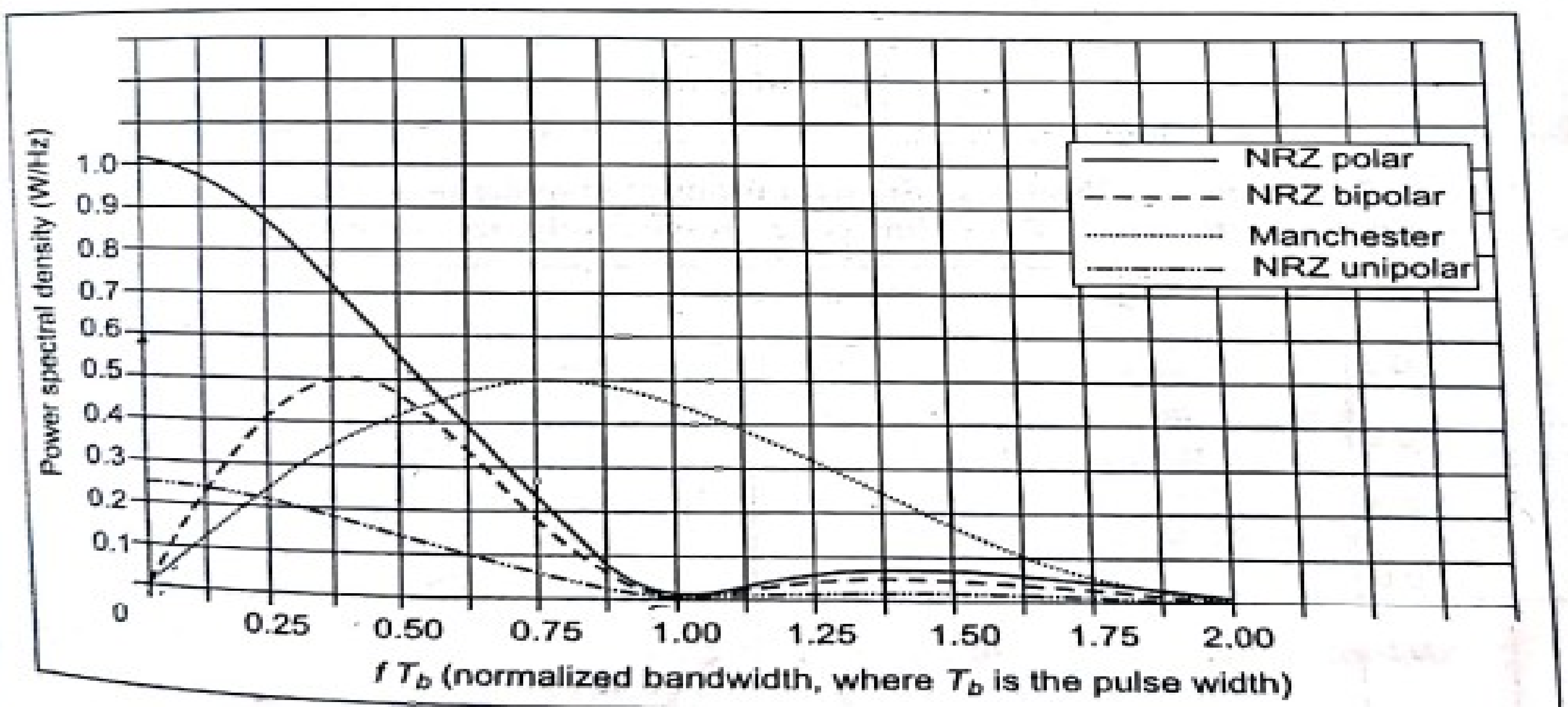
The expression for *PSD of Manchester line code* is given as

$$\boxed{P_X(f) = V^2 T_b \operatorname{sinc}^2\left(\frac{fT_b}{2}\right)\sin^2\left(\frac{\pi f T_b}{2}\right)}$$

Specifically, the power spectral density $P_X(f)$ is normalized with respect to V^2T_b, and the frequency f is normalized with respect to the bit rate $1/T_b$. It can be easily seen that the PSD of the Manchester line code has a dc null. The PSD of Manchester line code is shown in Figure 2.1.24 along with the power spectral density of unipolar NRZ, polar NRZ, and bipolar NRZ line codes for comparison purpose.



2.1.5 Application Areas of Line-Coding Techniques

Binary data can be transmitted using a number of different types of pulses. The choice o particular pair of pulses to represent the symbols 1 and 0 is called line coding and the choic generally made on the grounds of one or more of the following considerations:

- Presence or absence of a dc level.
- Power spectral density—particularly its value at 0 Hz.
- Bandwidth.
- Transparency (i.e., the property that any arbitrary symbol, or bit, pattern can be transmi and received).
- Ease of clock signal recovery for symbol synchronization.
- Presence or absence of inherent error detection properties.
- BER performance.

Table 2.1.6 Application Areas of Line-Coding Techniques

S. No.	Line-Coding Technique	Application Areas
1.	Unipolar NRZ	Internal computer waveforms
2.	Unipolar RZ	Baseband data transmission, Magnetic tape recording
3.	Polar NRZ-L	Digital logic circuits, Magnetic tape recording
4.	Bipolar RZ	Satellite telemetry communication links, Optical communicatio Magnetic tape-recording systems
5.	Manchester	IEEE 802.3 standard for Ethernet local area network (LAN)
6.	Bipolar RZ-AMI	Telephone systems as signaling scheme, T-digital carrier system
7.	B8ZS	T1 digital carrier system
8.	B6ZS	T2 digital carrier system
9.	B3ZS	T3 digital carrier system

2.2 INTERSYMBOL INTERFERENCE (ISI)

LO 2.2

When line-coded rectangular waveform is transmitted over a bandlimited channel, spectral distortion occurs due to suppression of a small part of the spectrum. This results into spreading of the pulse, known as *pulse dispersion*. Obviously, spreading of a pulse beyond its allotted time period T_b (pulse width) will tend to interfere with adjacent pulses. This is known as *Intersymbol Interference (ISI)*.

What We Discuss Here ....

In digital baseband signal transmission, ISI arises due to the dispersive nature of a communications channel. ISI is caused by non-ideal channels that are not distortionless over the entire signal bandwidth (since the transmitted signal bandwidth is more than the available channel bandwidth). ISI is not equivalent to channel noise but it is due to channel distortion. The ISI occurs due to the imperfections in the overall frequency response of the digital communication system. When a pulse of short width is transmitted through a band-limited communication system then the frequency components contained in the input pulse are differentially attenuated as well as delayed by the system. The pulse appearing at the output of the system will be dispersed over an interval which is longer than the transmitted pulse of short width.

What is InterSymbol Interference (ISI)?

2.2.1 Effects of ISI

Figure 2.2.1 shows a typical dispersed pulse due to ISI.

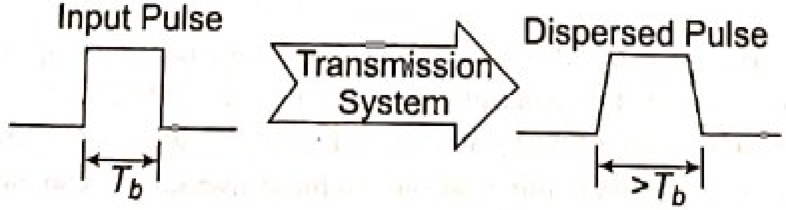


Figure 2.2.1 Dispersed Pulse due to ISI

It is seen that due to pulse dispersion, each symbol having short duration will interfere with each other when transmitted over the communication channel. This will result in ISI, as shown in Figure 2.2.2.

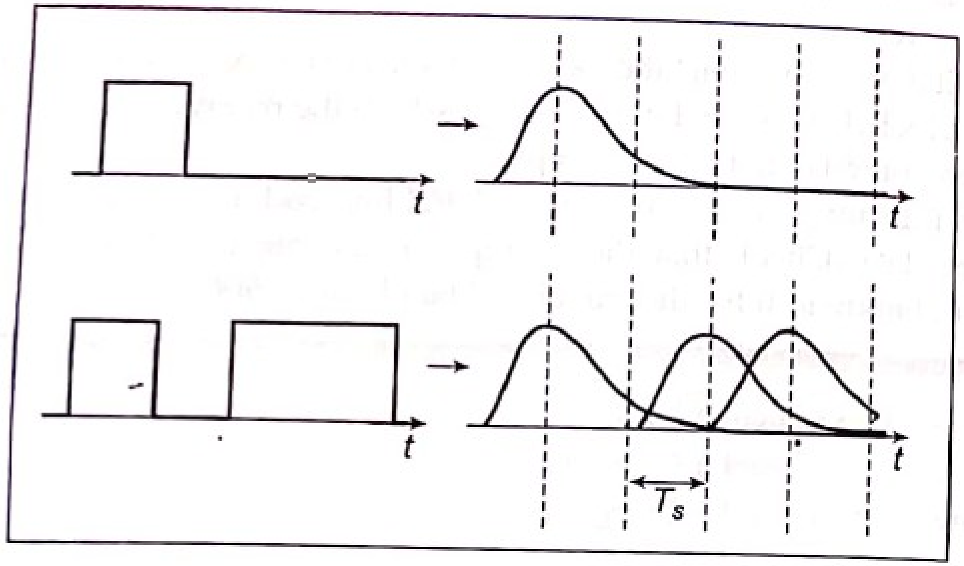


Figure 2.2.2 Effect of ISI on transmission of pulses

We know that in the absence of ISI and channel noise, the transmitted symbol can be decoded correctly at the receiver. However, due to occurrence of ISI, an error will be introduced in the decision-making device at the receiver output.[10]

In a bandlimited PCM channel, the received pulse waveforms may be distorted due to ISI and may extend to the next time slot (termed *interchannel interference*). If ISI is large enough then it can cause errors in detection of the received pulse at the regenerative repeater or the receiver of baseband digital communication system such as PCM.

Let us consider a mathematical model of a **baseband binary data transmission system**, as shown in Figure 2.2.3.

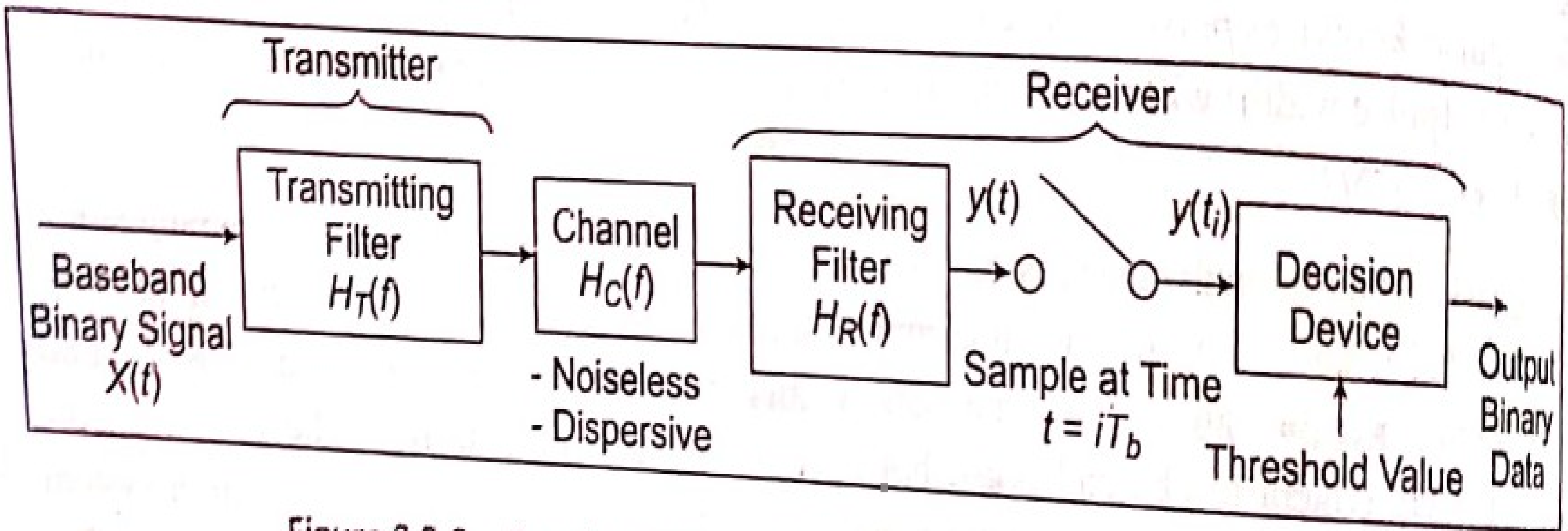


Figure 2.2.3 Baseband Binary Data Transmission System Model

Step I Let the waveform of the k^{th} pulse in the baseband-modulated binary signal be given as

$$X_k(t) = A_k v\,(t - kT_b)$$

where A_k is the coefficient, which depends on the input data and the type of line code used; $v(t)$ denotes the basic pulse waveform normalized such that $v(0) = 1$, and T_b is the pulse duration in seconds.

Step II This signal passes through a transmitting filter of transfer function, $H_T(f)$ and presented to the channel.

Assume The channel is noiseless, but dispersive in nature.

Step III The output is distorted as a result of transmission through the bandlimited channel (represented as a bandpass filter) of transfer function, $H_C(f)$.

Step IV The channel output is passed through a receiving filter of transfer function, $H_R(f)$.

Step V This filter output is sampled synchronously with that of transmitter.

The output of the receiving filter, $y(t)$ may be expressed as

$$y(t) = \mu \sum_{k=-\infty}^{\infty} A_k v(t - kT_b)$$

where μ is a scaling factor.

The receiving filter output, sampled at time $t_i = iT_b$ (with i can assume integer values), yields

$$y(t_i) = \mu \sum_{k=-\infty}^{\infty} A_k v(iT_b - kT_b)$$

$$\Rightarrow \qquad y(t_i) = \mu \sum_{k=i} A_k v(iT_b - kT_b) + \mu \sum_{\substack{k=-\infty \\ k \neq i}}^{\infty} A_k v(iT_b - kT_b)$$

$$\Rightarrow \qquad y(t_i) = \mu A_i + \mu \sum_{\substack{k=-\infty \\ k \neq i}}^{\infty} A_k v(iT_b - kT_b)$$

In this expression, we note that

- the first term is produced by the transmitted pulse itself; and
- the second term represents the residual effect of all other transmitted pulses on the decoding of the i^{th} bit.

This residual effect is known as ***Intersymbol Interference (ISI)***.

Step VI The sequence of samples is used to reconstruct the original data sequence by means of a decision device depending upon the preset threshold value.

If there is no ISI at the decision-making instants in the receiver despite pulse spreading or overlapping, pulse amplitudes can still be detected correctly. So it is utmost necessary either to eliminate or at least ...

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as the **intersymbol interference (ISI).**

5.26. Cause of Intersymol Interference (ISI)

The intersymbol interference ISI arises due to the imperfections in the overall frequency response of the system. When, a short pulse of duration T_b seconds is transmitted through a bandlimited system, then the frequency components contained in the input pulse are **differentially attenuated** and more importantly **differentially delayed** by the system. Due to this, the pulse appearing at the output of the system will be "dispersed" over an interval which is longer than T_b seconds. Due to this dispersion, the symbols each of duration T_b will interfere with each other when transmitted over the communication channel. This will result in the intersymbol interference (ISI).

The transmitted pulse of duration T_b seconds and the dispersed pulse of duration more than T_b seconds are shown in figure 5.31.

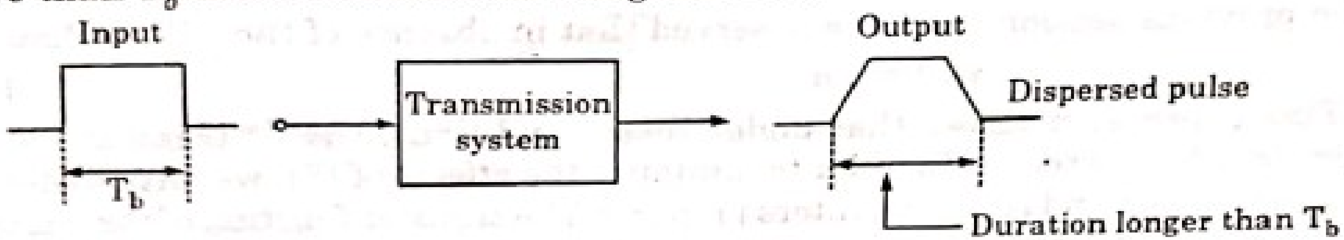


Fig. 5.31. *Cause of Intersymbol Interference (ISI).*

Effect of Intersymbol Interference (ISI)

Following are the effects of ISI:

(*i*) In the absence of ISI and noise, the transmitted bit can be decoded correctly at the receiver.

(*ii*) The presence of ISI will introduce errors in the decision at the receiver output.

(*iii*) Hence, the receiver can make an error in deciding whether it has received a logic 1 or a logic 0.

Remedy to Reduce ISI

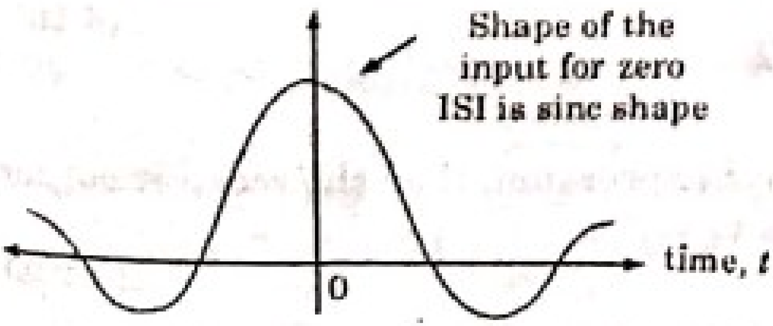
(*i*) It has been proved that the function which produces a zero intersymbol a logic 1 or a logic 0.

Remedy to Reduce ISI

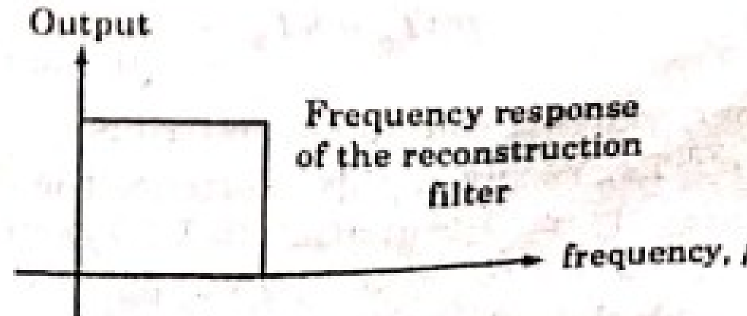
(i) It has been proved that the function which produces a zero intersymbol interference is a "sinc function". Hence, instead of a rectangular pulse if we transmit a sinc pulse then the ISI can be reduced to zero.

(ii) This is known as *Nyquist Pulse Shaping*. The sinc pulse transmitted to have a zero ISI has been shown in figure 5.32 (*a*).

(iii) Further, we know that Fourier transform of a sinc pulse is a rectangular function. Hence, to preserve all the frequency components, the frequency response of the filter must be exactly flat in the pass band and zero in the attenuation band as shown in figure 5.32 (*b*).



(a) Ideal pulse shape for zero ISI



(b) Frequency response of the filter

Fig. 5.32.

Advantages of Using the Sinc Pulse

(*i*) Bandwidth requirement (of the channel) is reduced.

(*ii*) ISI is reduced to zero.

Possible Difficulties

(*i*) It is necessary that the amplitude characteristics of $P(f)$ should be flat from $-B_0$ to B_0 and zero outside this band. But, abrupt transition at $\pm B_0$ is not physically realizable.

(*ii*) Due to discontinuity of $P(f)$ at $\pm B_0$, there is practically no margin of error in sampling times at the receiver end.

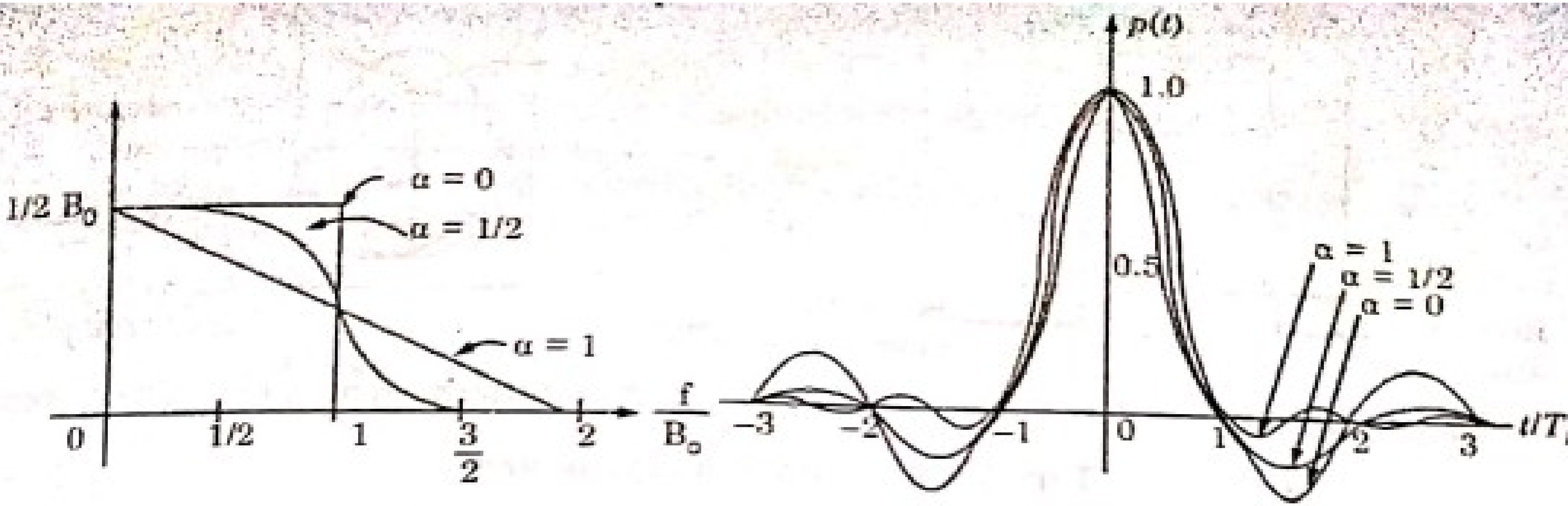
Remedy

(*i*) Due to these problems, it is practically not possible to use the sinc function as it is.

(*ii*) But, the practical difficulties can be overcome by increasing the bandwidth from $B_0 = R_b/2$ to an adjustable value between B_0 and $2B_0$.

(*iii*) The frequency response for different roll off factors (α) and the corresponding time response have been shown in figure 5.35(*a*) and (*b*) respectively.

- However, type of filter is practically not present or not possible. Hence, in practice, the frequency response of the filter is modified as shown in figure 5.33 with different roll off factors α to obtain the achievable filter response curves.



(a) *Frequency response* (b) *Time response*

Fig. 5.35. *Responses for different roll-off factors α.*

Observations

(i) For $\alpha = 0.5$ and 1, the characteristics of $P(f)$ changes gradually with respect to frequency. Hence it is easier to realize this characteristics practically.

(ii) The time response has a sinc shape and all the sinc functions pass through zero at $t = \pm T_b$, $\pm 2T_b$, ...

(iii) The amplitude of side lobes increases with reduction in the value of α.

(iv) With $\alpha = 0$, the bandwidth requirement is maximum equal to $2B_0$.

5.27. Nyquist's Criterion for Distortionless Basband Binary Transmission

In the previous section, we have observed that in absence of the ISI, we have

$$y(t_i) = \mu a_i \quad ...(5.105)$$

This expression shows that under these conditions, the i^{th} transmitted bit can be decoded correctly. In order to minimize the effects of ISI, we have to design the transmitting and receiving filters properly. The transfer function of the channel and the shape of transmitted pulse are generally specified. Therefore, it becomes the first step towards design of filters. From this information we have to determine the transfer functions of the transmitting and receiving filters, to reconstruct the transmitted data sequence $\{b_k\}$. This is achieved by first extracting and then decoding the corresponding sequence of weights from the output $y(t)$.

As expressed by following equation

$$y(t) = \mu \sum_{k=-\infty}^{\infty} a_k\, p(t - kT_b) \quad ...(5.106)$$

This shows that output $y(t)$ is dependent on a_k, the received pulse $p(t)$ and the scaling factor μ.

Extraction

Extraction is basically the process of sampling. The signal $y(t)$ is sampled at $t = iT_b$.

Decoding

The decoding should be such that the contribution of the weighted pulse i.e., $a_k\, p(iT_b - kT_b)$ for $i = k$ be free from ISI. This can be stated mathematically as under:

$$p(iT_b - kT_b) = \begin{cases} 1 & \text{for } i = k \\ 0 & \text{for } i \neq k \end{cases} \quad ...(5.107)$$

where $p(0) = 1$ due to normalizing.

If $p(t)$ i.e., received pulse satisfies the above expression, then the receiver output given by equation (5.107) reduces to

$$y(t_i) = \mu a_i \quad ...(5.108)$$

... indicates zero ISI in the absence of noise.

5.29. Concept of Eye Pattern

Eye pattern is a pattern displayed on the screen of a cathode ray oscilloscope (C.R.O.). The shape of this pattern resembles with the shape of human eye. Therefore, it is called as *eye pattern*. Eye pattern is a practical way to study the intersymbol interference (ISI) and its effects on a PCM or data communication system. The eye pattern is obtained on the C.R.O. by applying the received signal to vertical deflection plates (*Y*-plates) of the C.R.O. and a sawtooth wave at the transmission symbol rate *i.e.*, ($1/T_b$) to the horizontal deflection plates (*X*-plates) as shown in figure 5.36(*c*). The received digital signal and the corresponding oscilloscope display are as shown in figure 5.36(*a*) and (*c*) respectively. The resulting oscilloscope display shown in figure 5.36(*c*) is called as the *eye pattern*. This is due to its resemblance to the human eye.

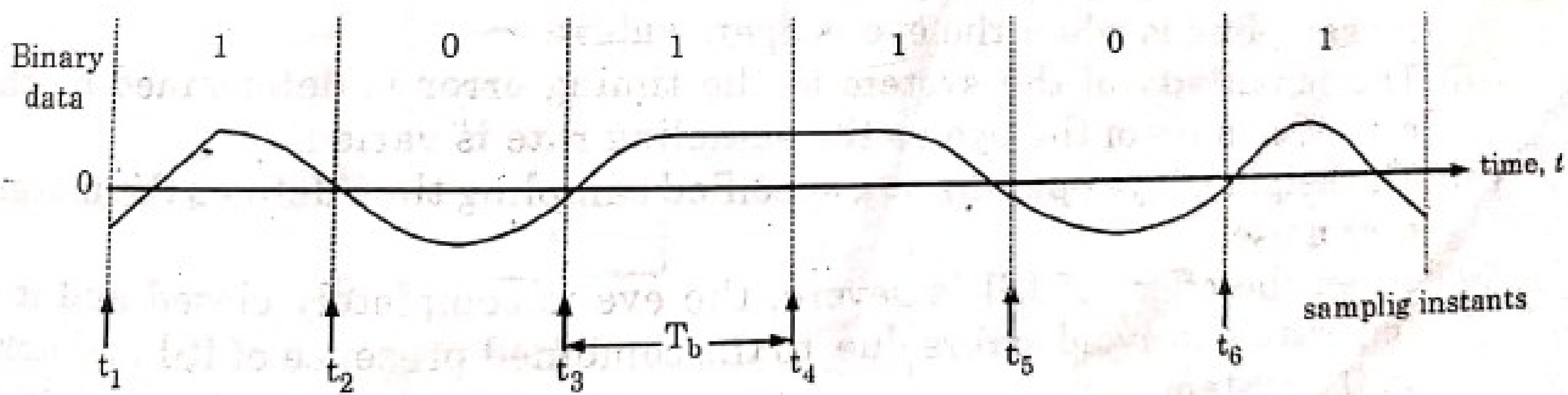
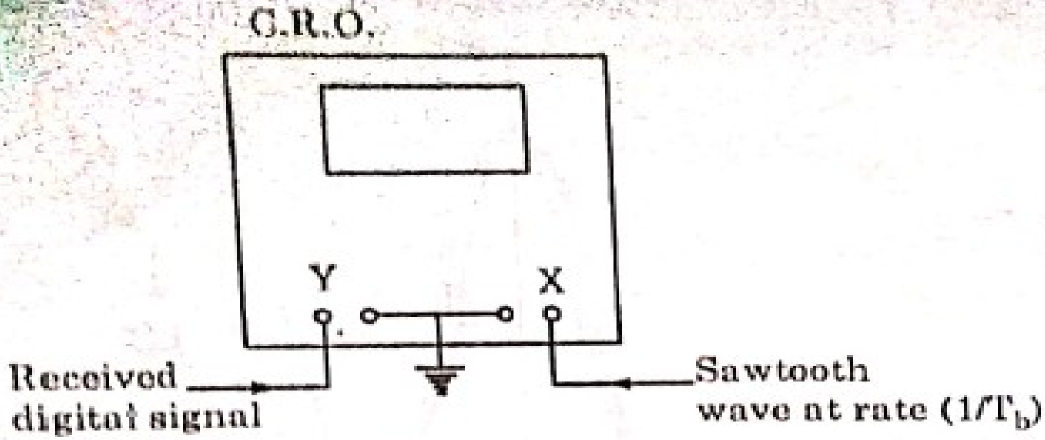
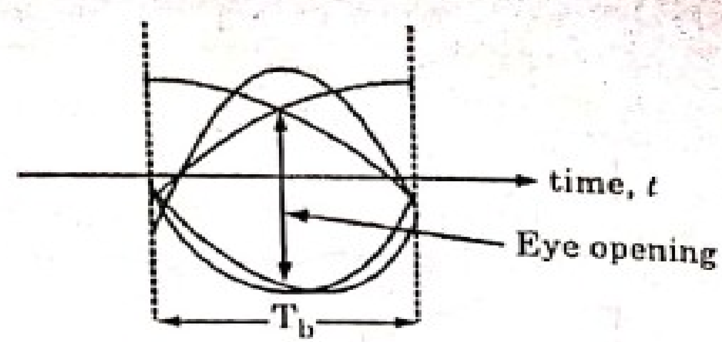


Fig. 5.36. *(a) Distorted binary wave*



(b) Oscilloscope connections.



(c) Eye pattern seen on the CRO screen.

Fig. 5.36. *Obtaining eye pattern.*

The interior region of the eye pattern is called as the eye opening. The eye pattern provides a great deal of information about the performance of the system. The information obtainable is as follows (Figure 5.37)

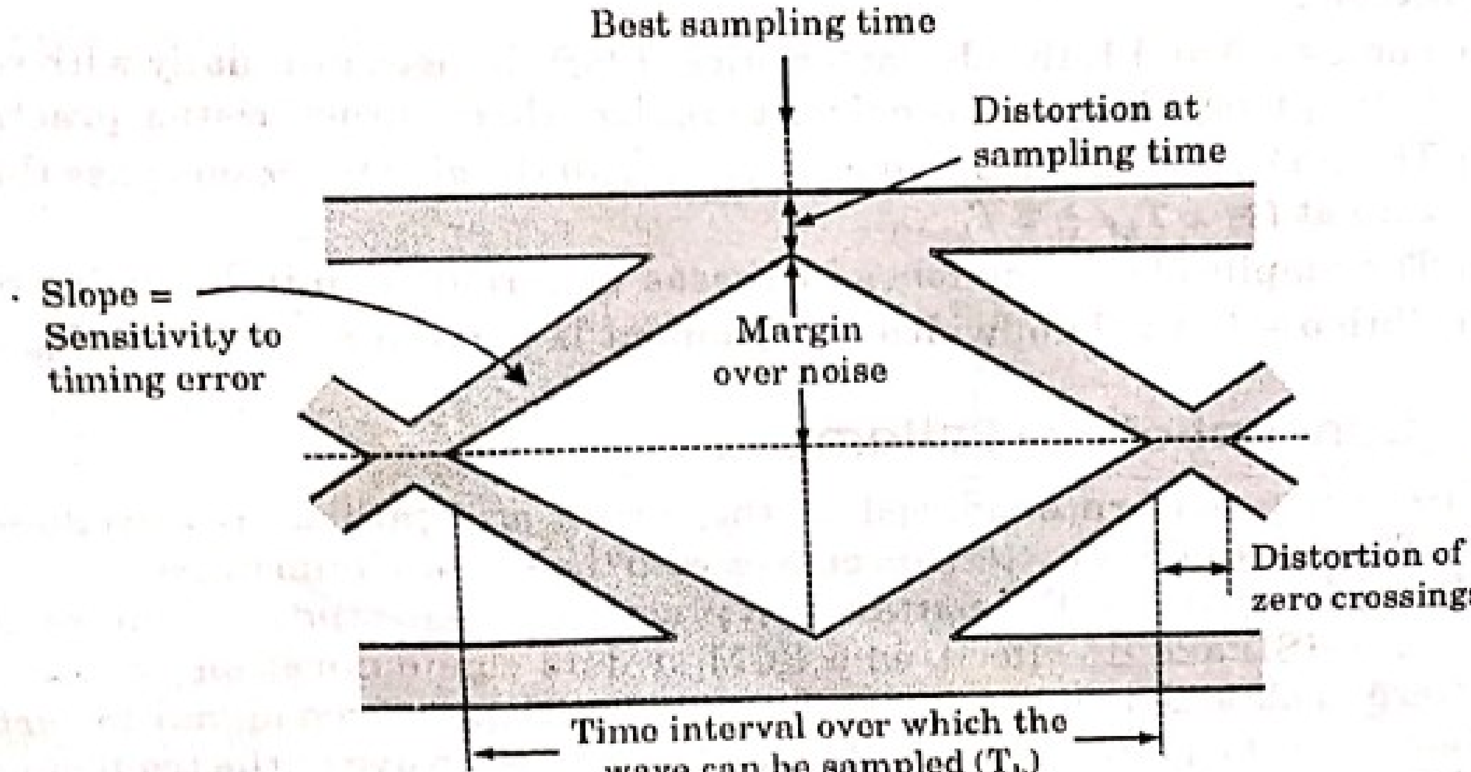


Fig. 5.37. *Interpretation of eye pattern.*

Information Obtained from Eye Pattern

(*i*) The width of the eye opening defines the time interval over which the received wave can be sampled, without an error due to ISI. The best time for sampling is when the eye js open widest.

(*ii*) The sensitivity of the system to the timing error is determined by the rate of closure of the eye as the sampling rate is varied.

(*iii*) The height of eye opening at a specified sampling time defines the margin over noise.

(*iv*) When the effect of ISI is severe, the eye is completely closed and it is impossible to avoid errors due to the combined presence of ISI and noise in the system.

REVIEW QUESTIONS

1. Explain the difference between bandpass transmission and passband transmission.
2. Explain various PAM digital formats.
3. Explain the followings:
   (*i*) Intersymbol Interference, (*ii*) Eye pattern
4. What is Line coding ? Explain.
5. What are the properties of Line coding?
6. Draw the block diagram of a baseband digital communications of various blocks.
7. What is Nyquist criterion of zero ISI ? Explain.
8. Why do we need to use the dicrete PAM formats?
9. State the important properties of line codes.
10. What is the meaning of word "RZ".
11. State true or false
    (*a*) The dc value of unipolar RZ format is zero.
    (*b*) The unipolar NRZ signal used two different amplitudes
    (*c*) Polar NRZ format has a zero dc value
12. In the bipolar NRZ or AMI line codes, the binary 0 is represented by
    (*a*) alternate zeros and ones.
    (*b*) alternate + A and – A amplituds.
    (*c*) zero amplitudes
13. The synchronization at the receiver is better for:
    (*a*) Bipolar RZ (*b*) Bipolar NRZ (*c*) Manchester
14. Explain the various techniques to detect the baseband digital signals.

Digital Modulation Technique

As discussed earlier, Modulation is defined as the process by which some characteristics of a carrier is varied in accordance with a modulating signal. In digital communications, the modulating signal consists of binary data or an M-ary encoded version of it. This data is used to modulate a carrier wave (usually sinusoidal) with fixed frequency. In fact, the input data may represent the digital computer outputs or PCM waves generated by digitizing voice or video signals. The channel may be a telephone channel, microwave radio link, satellite channel or an optical fiber. In digital communication, the modulation process involves switching or keying the amplitude, frequency or phase of the carrier in accordance with the input data.

When we have to transmit a digital signal over a long distance, we need continuous-wave (CW) modulation. For this purpose, the transmission medium can be in form of radio, cable or other type of channel. Also, a carrier signal having some frequency f_c is used for modulation. Then the modulating digital signal modulate some prameter like frequency, phase or amplitude of the carrier. Due to thi process, there is some deviation in carrier frequency f_c. This deviation is know as the bandwidth of the channel. This means that the channel has to transmi some range or band of frequencies. Such type of transmission is known as **bandpas transmission** and the communication channel is known as **bandpass channel**

Here, the word bandpass is used since the range of frequencies does not star from zero Hz to f_m Hz. In fact, the range of frequencies from zero Hz to f_m Hz i known as **low-pass signal** and such channel is known as **low-pass channel.**

Now, when it is required to transmit digital signals on a bandpass channel the amplitude, frequency or phase of the sinusoidal carrier is varied in accordance with the incoming digital data. Since the digital data is in discrete steps, the modulation of the bandpass sinusoidal carrier is also done in discrete steps. Due to this reason, this type of modulation (*i.e.*, Digital modulation) is also known as switching or signaling. Now, if an amplitude of the carrier is switched depending on the input digital signal, then it is called Amplitude shift keying (ASK).

This process is quite similar to analog amplitude modulation. If the frequency of the sinusoidal carrier is switched depending upon the input digital signal, then it is known as the frequency shift keying (FSK). This is very much similar to the analog frequency modulation. If the phase of the carrier is switched depending upon the input digital signal, then it is called phase shift keying (PSK). This is similar to phase modulation. Since the phase and frequency modulation has constant amplitude envelope, therefore FSK and PSK also has a constant amplitude envelope. Because of constant amplitude of FSK and PSK, the effect of non-linearities, noise interference is minimum on signal detection. However, these effects are more pronounced on ASK. Therefore, FSK and PSK are preferred over ASK.

Figure 7.1 shows the waveforms for amplitude-shift keying, phase-shift keying and frequency shift keying. In these waveforms, a single feature of the carrier (*i.e.*, amplitude, phase or frequency) undergoes modulation.

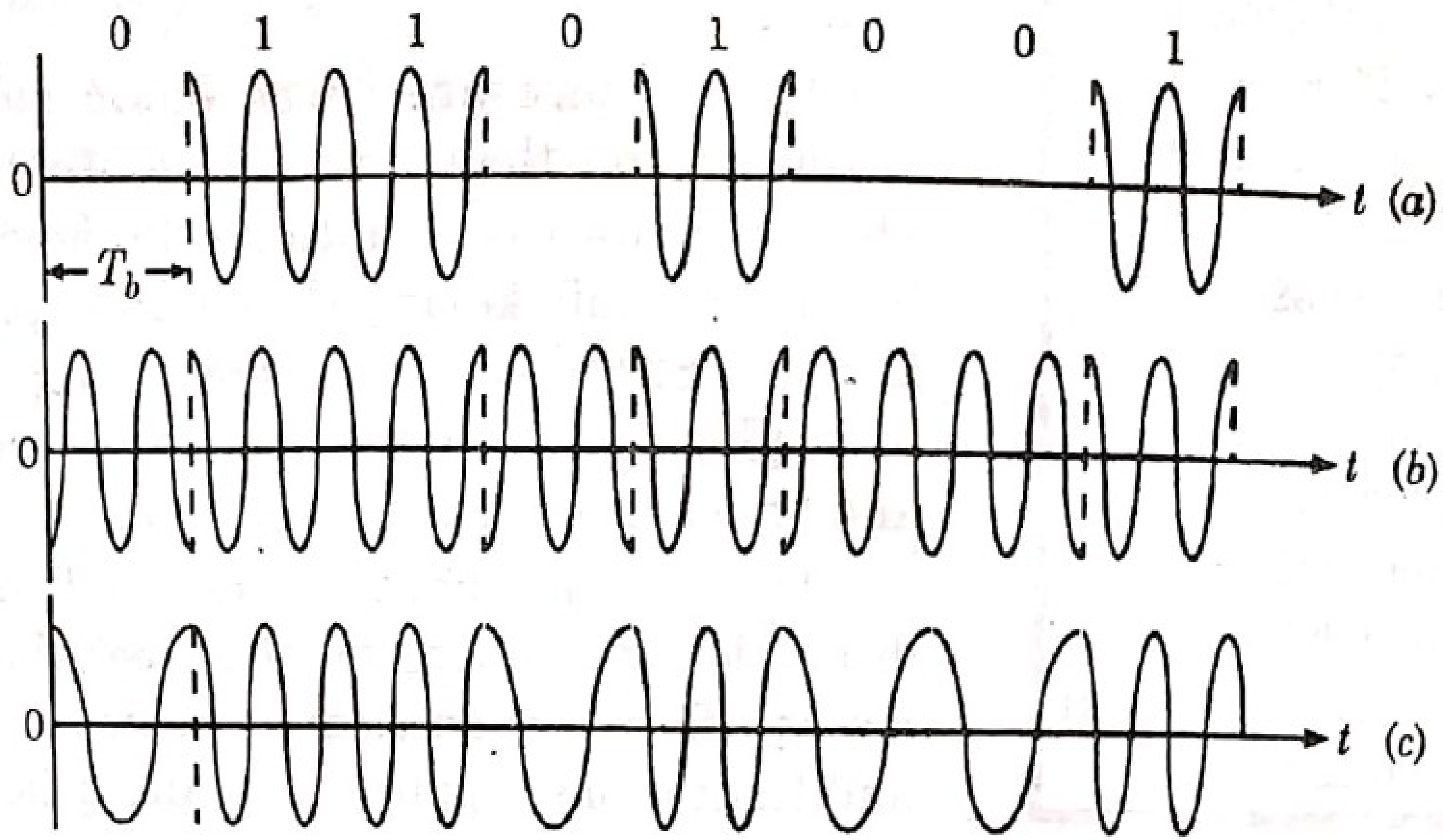


Fig. 7.1. *The three basic forms of signaling binary information, (a) Amplitude-shift keying, (b) Phase-shift keying, (c) Frequency shift keying with continuous phase*

In digital modulations, instead of transmitting one bit at a time, we transmit two or more bits simultaneously. This is known as M-ary transmission. This type of transmission results in reduced channel bandwidth. However, sometimes, we use two quadrature carriers for modulation. This process is known as Quadrature modulation.

Thus, we see that there are a number of modulation schemes available t the designer of a digital communication system required for data transmissio over a bandpass channel.

Every scheme offers system trade-offs of its own. However, the final choice made b the designer is determined by the way in which the available primary communicatio resources such as transmitted power and channel bandwidth are best exploited. I particular, the choice is made in favour of a scheme which possesses as many of th following design characteristics as possible:

(*i*) Maximum data rate,

(*ii*) Minimum probability of symbol error,

(*iii*) Minimum transmitted power,

(*iv*) Maximum channel bandwidth,

(*v*) Maximum resistance to interfering signals,

(*vi*) Minimum circuit complexity.

7.3. Types of Digital Modulation Techniques

(U.P. Tech., Sem. Examination, 2003-2004)

Basically, digital modulation techniques may be classified into coherent or non-coherent techniques, depending on whether the receiver is equipped with a phase-recovery circuit or not. The phase-recovery circuit ensures that the oscillator supplying the locally generated carrier wave receiver is synchronized* to the oscillator supplying the carrier wave used to originally modulate the incoming data stream in the transmitter.

(i) Coherent Digital Modulation Techniques

Coherent digital modulation techniques are those techniques which employ coherent detection. In coherent detection, the local carrier generated at the receiver is phase locked with the carrier at the transmitter. Thus, the detection is done by correlating received noisy signal and locally generated carrier. The coherent detection is a synchronous detection.

(ii) Non-coherent Digital Modulation Techniques

Non-coherent digital modulation techniques are those techniques in which the detection process does not need receiver carrier to be phase locked with transmitter carrier.

The advantage of such type of system is that the system becomes simple. But the drawback of such a system is that the error probability increases.

In fact, the different digital modulation techniques are used for various specific application areas.

7.4. Coherent Binary Modulation Techniques

As mentioned earlier, the binary (*i.e.*, digital) modulation has three basic forms amplitude-shift keying (ASK), phase-shift keying (PSK) and frequency-shift keying (FSK). In this section, let us discuss different coherent binary modulation techniques.

7.5. Coherent Binary Amplitude Shift Keying or On-Off Keying

Amplitude shift keying (ASK) or ON-OFF keying (OOK) is the simplest digital modulation technique. In this method, there is only one unit energy carrier and

is switched on or off depending upon the input binary sequence. The AS
aveform may be represented as,

$$s(t) = \sqrt{2P_s} \cos(2\pi f_c t) \text{ (To transmit '1')} \qquad ...(7.$$

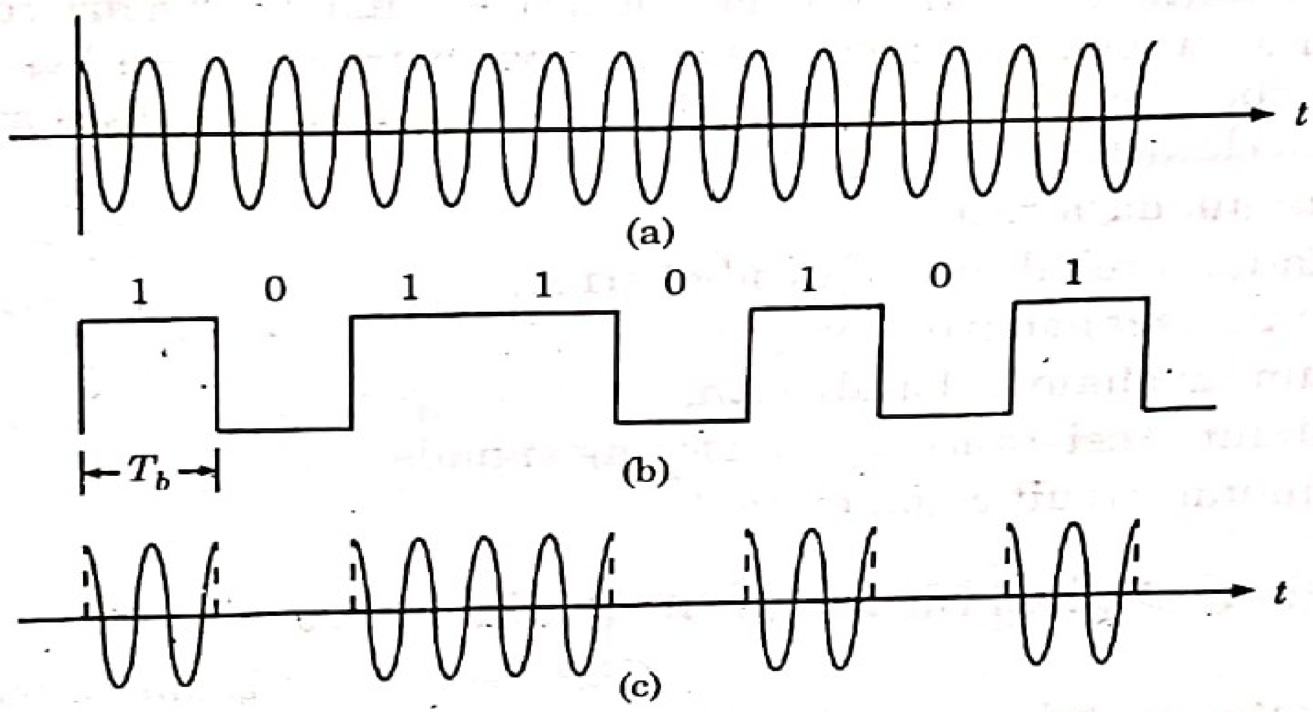


Fig. 7.2: *Amplitude-shift keying waveforms, (a) Unmodulated carrier, (b) Unipolar bit sequence, (c) ASK waveform.*

To transmit symbol '0', the signal $s(t) = 0$ *i.e.*, no signal is transmitted. Signal $s(t)$ contains some complete cycles of carrier frequency 'f_c'.

Hence, the ASK waveform looks like an ON-OFF of the signal. Therefore, it is also known as the **ON-OFF keying** (OOK). Figure 7.2 shows the ASK waveform.

7.5.2. Generation of ASK Signal

ASK signal may be generated by simply applying the incoming binary data (represented in unipolar form) and the sinusoidal carrier to the two inputs of a product modulator (*i.e.*, balanced modulator). The resulting output will be the ASK waveform. This is shown in figure 7.4. Modulation causes a shift of the baseband signal spectrum. The ASK signal, which is basically the product of the binary sequence and the carrier signal, has a power spectral density (PSD) same as that of the baseband on-off signal but shifted in the frequency domain by $\pm f_c$.

This is shown in figure 7.5. It may be noted that two impulses occur at $\pm f_c$. The spectrum of the ASK signal shows that it has an infinite bandwidth. However for practical purpose, the bandwidth is often defined as the bandwidth of an ideal bandpass filter centered at f_c whose output contains about 95% of the total average power content of the ASK signal. It may be proved that according to this criterion the bandwidth of the ASK signal is approximately $3/T_b$ Hz. The bandwidth of the ASK signal can however, be reduced by using smoothed versions of the pulse waveform instead of rectangular pulse waveforms.

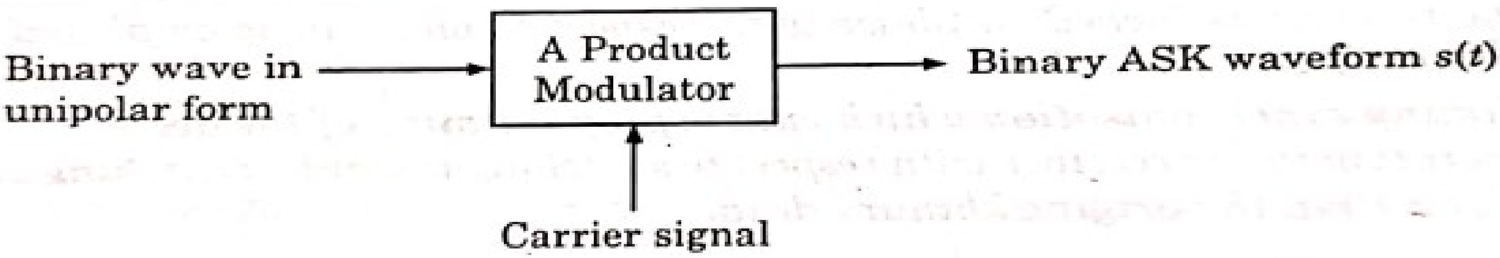


Fig. 7.4. *Generation of binary ASK waveform.*

$B_{ASK} = (1+r) f$, $r = 0$

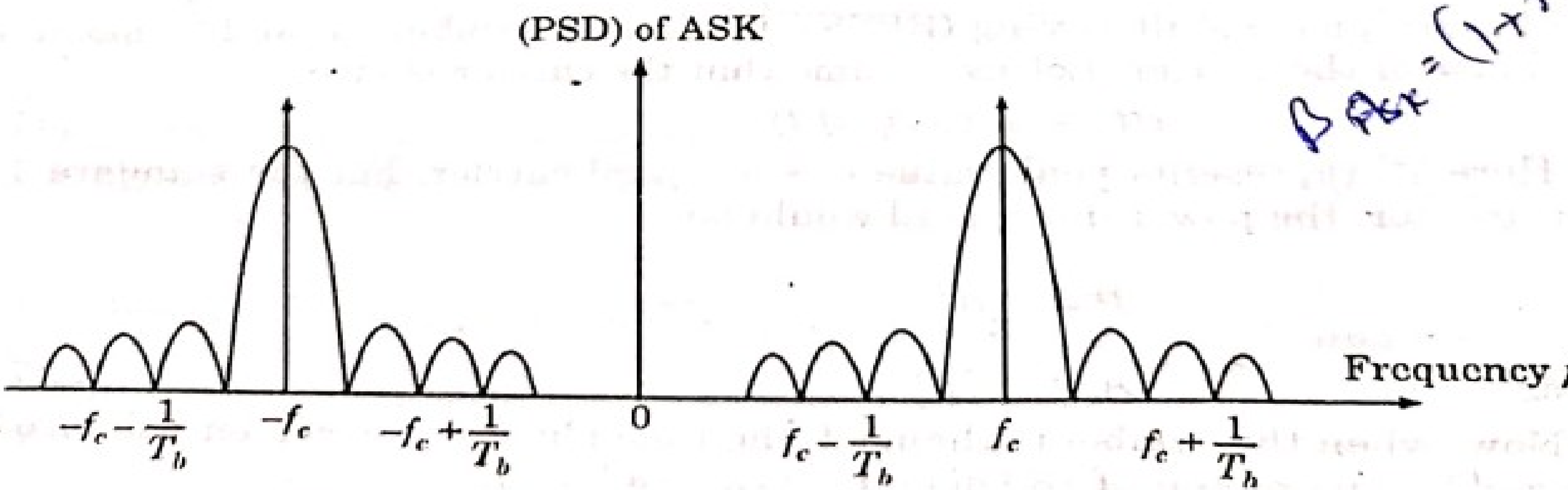


Fig. 7.5. *Power spectral density of ASK signal.*

7.5.3. Coherent Demodulation of Binary ASK

The demodulation of binary ASK waveform can be achieved with the help of *coherent detector* as shown in figure 7.6. It consists of a product modulator which is followed by an integrator and a decision-making device. The incoming ASK signal is applied to one input of the product modulator. The other input of the product modulator is supplied with a sinusoidal carrier which is generated with the help of a local oscillator. The output of the product modulator goes to input of the integrator. The integrator operates on the output of the multiplier for successive bit intervals and essentially performs a low-pass filtering action. The output of the integrator goes to the input of a decision-making device.

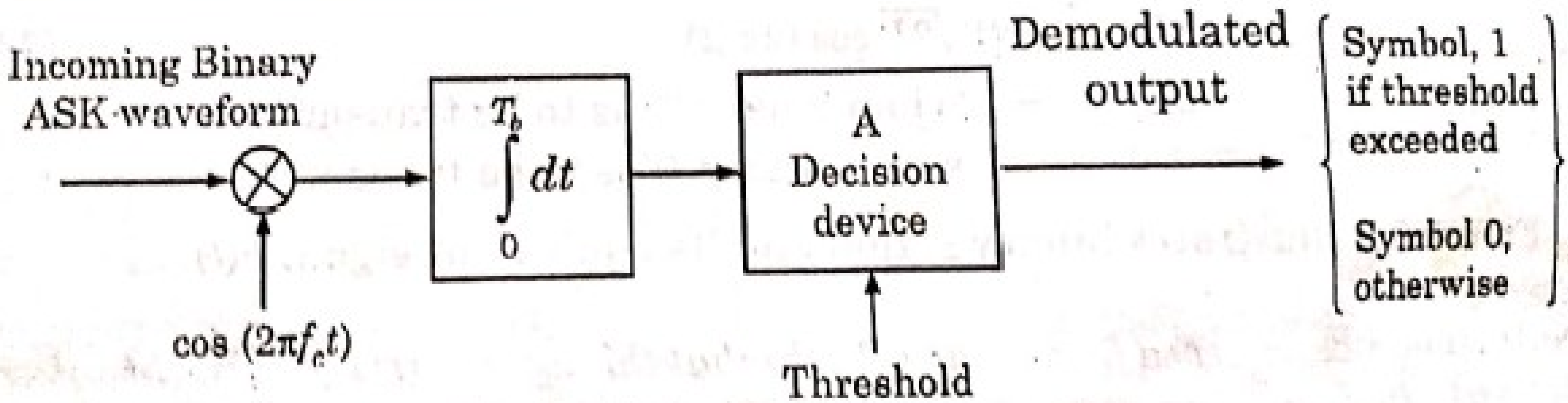


Fig. 7.6. *Coherent detection of binary ASK signals.*

Now, the decision-making device compares the output of the integrator with a preset threshold. It makes a decision in favour of symbol 1 when the threshold is exceeded and in favour of symbol 0 otherwise. The *coherent detection* makes the use of linear operation. In this method we have assumed that the local carrier is in perfect synchronisation with the carriers used in the transmitter. This means that the frequency and phase of the locally generated carrier is same as those of the carriers used in the transmitter.

The following two forms of synchronisation are required for the operation of coherent (or synchronous detector):

(i) Phase synchronisation which ensures that carrier wave generated locally in the receiver is locked in phase with respect to one that is employed in the transmitter.

(ii) Timing synchronisation which enable proper timing of the decision making operation in the receiver with respect to switching instants (switching between 1 and 0) in the original binary data.

7.6. Binary Phase Shift Keying (BPSK)

In a binary phase shift keying (BPSK), the binary symbols '1' and '0' modulate the phase of the carrier. Let us assume that the carrier is given as,

$$s(t) = A \cos(2\pi f_c t) \quad \text{...(7.4)}$$

Here 'A' represents peak value of sinusoidal carrier. For the standard 1 Ω load resistor, the power dissipated would be,

$$P = \frac{1}{2}A^2$$

or

$$A = \sqrt{2P} \quad \text{...(7.5)}$$

Now, when the symbol is changed, then the phase of the carrier will also be changed by an amount of 180 degrees (*i.e.*, π radians).

Let us consider, for example,

For symbol '1', we have

$$s_1(t) = \sqrt{2P}\cos(2\pi f_c t) \quad \text{...(7.6)}$$

If next symbol is '0', then we have

For symbol '0', we have

$$s_2(t) = \sqrt{2P}\cos(2\pi f_c t + \pi) \quad \text{...(7.7)}$$

Now, because $\cos(\theta + \pi) = -\cos\theta$, therefore, the last equation can be written as

$$s_2(t) = -\sqrt{2P}\cos(2\pi f_c t) \quad \text{...(7.8)}$$

With the above equation, we can define BPSK signal combinely as,

$$s(t) = b(t)\sqrt{2P}\cos(2\pi f_c t) \quad \text{...(7.9)}$$

where $b(t) = +1$ when binary '1' is to be transmitted.

$\qquad\qquad -1$ when binary '0' is to be transmitted

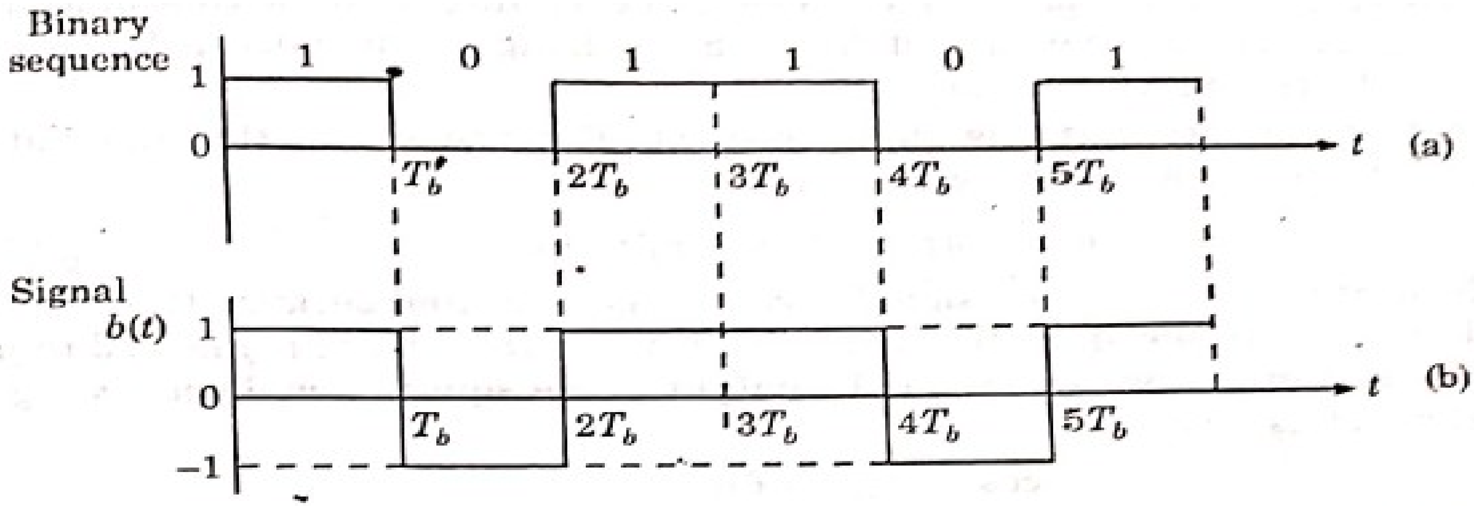


Fig. 7.7. *(a) Binary sequence, (b) The corresponding bipolar signal b(t).*

7.6.1. Generation of BPSK Signal

The BPSK signal may be generated by applying carrier signal to a balanced modulator. Here, the bipolar signal $b(t)$ is applied as a modulating signal to the balanced modulator.

Figure 7.8 shows the block diagram of a BPSK signal generator.

A NRZ level encoder converts the binary data sequence into bipolar NRZ signal.

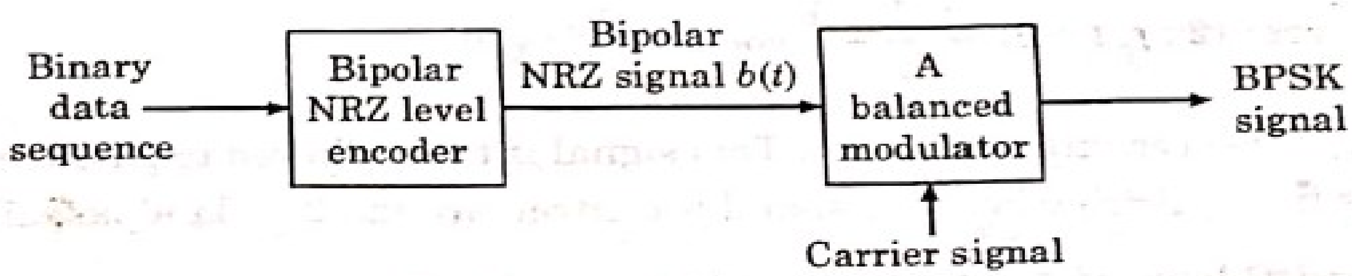


Fig. 7.8. *Generation of BPSK.*

7.6.2. Reception (*i.e.* Detection) of BPSK Signal

Figure 7.9 shows the block diagram of the scheme to recover baseband signal from BPSK signal. The transmitted BPSK signal is given as

$$s(t) = b(t)\sqrt{2P}\cos(2\pi f_c t)$$

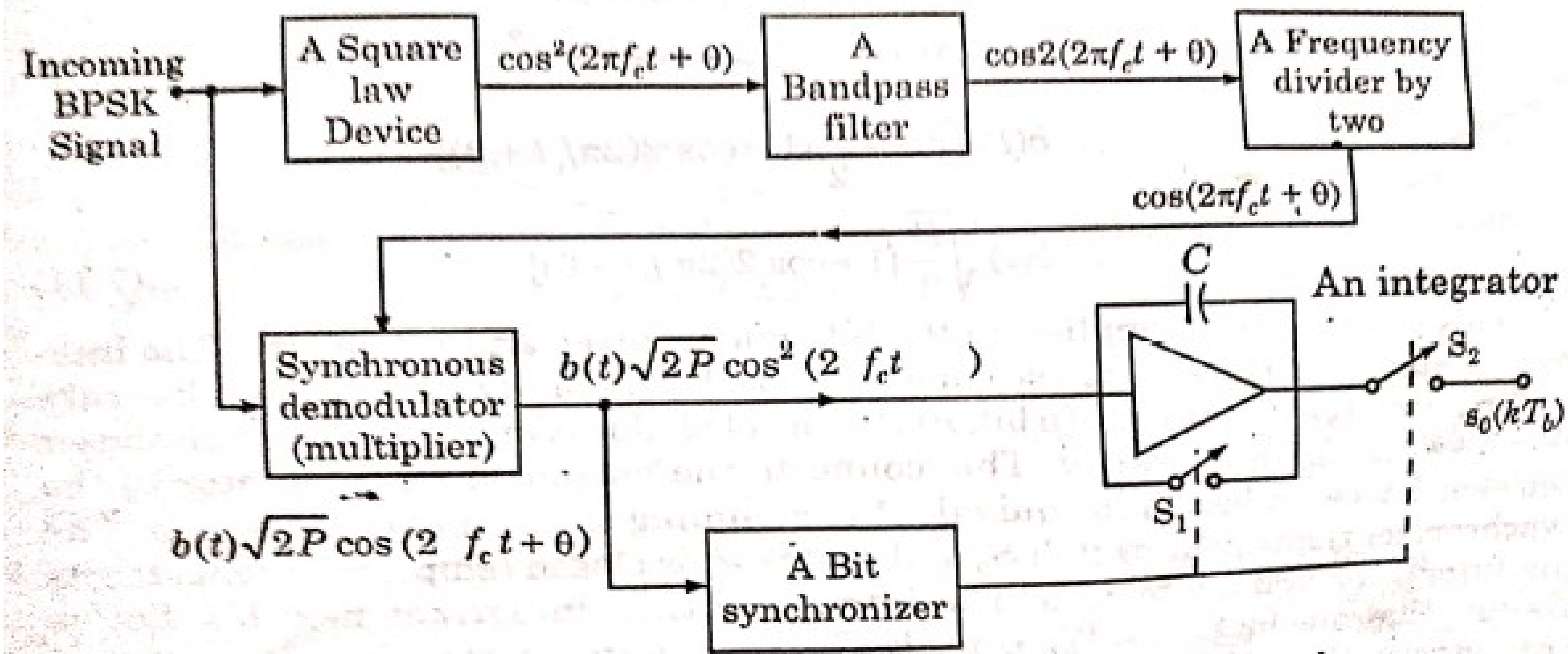


Fig. 7.9. *Reception of baseband signal in BPSK signal.*

This signal undergoes the phase change depending upon the time delay from transmitter end to receiver end. This phase change is, usually, a fixed phase shift in the transmitted signal.

Let us consider that this phase shift is θ. Because of this, the signal at the input of the receiver can be written as

$$s(t) = b(t)\sqrt{2P}\cos(2\pi f_c t + \theta) \qquad ...(7.10)$$

Now, from this received signal, a carrier is separated because this is coherent detection. As shown in the figure 7.9, the received signal is allowed to pass through a square law device. At the output of the square law device, we get a signal which is given as

$$\cos^2(2\pi f_c t + \theta)$$

Here, it may be noted that we have neglected the amplitude, since we are only interested in the carrier of the signal.

Again, we know that

$$\cos^2\theta = \frac{1+\cos 2\theta}{2}$$

Therefore, we have

$$\cos^2(2\pi f_c t + \theta) = \frac{1+\cos 2(2\pi f_c t+\theta)}{2}$$

$$\cos^2(2\pi f_c t + \theta) = \frac{1}{2} + \frac{1}{2}\cos 2(2\pi f_c t + \theta)$$

Here, $\frac{1}{2}$ represents a DC level. This signal is then allowed to pass through bandpass filter (BPF) whose passband is centred around $2f_c$. Bandpass filter removes the DC level of $\frac{1}{2}$ and at the output, we obtain,

$$\cos 2(2\pi f_c t + \theta)$$

... it is passed through

$\cos 2(2\pi f_c t + \theta)$

This signal is having frequency equal to $2 f_c$. Hence, it is passed through a frequency divider by two. Thus, at the output of frequency divider, we get a carrier signal whose frequency is f_c i.e., $\cos(2\pi f_c t + \theta)$.

The synchronous (i.e., coherent) demodulator multiplies the input signal and the recovered carrier. Hence, at the output of multiplier, we get

$$b(t)\sqrt{2P}\cos(2\pi f_c t+\theta)\times\cos(2\pi f_c t+\theta) = b(t)\sqrt{2P}\cos^2(2\pi f_c t+\theta)$$

$$= b(t)\sqrt{2P}\times\frac{1}{2}[1+\cos 2(2\pi f_c t+\theta)]$$

$$= b(t)\sqrt{\frac{P}{2}}[1+\cos 2(2\pi f_c t+\theta)] \quad ...(7.11)$$

This signal is then applied to the bit synchronizer and integrator. The integrator integrates the signal over one bit period. The bit synchronized takes care of starting and ending times of a bit. At the end of bit duration T_b, the bit synchronizer closes switch S_2 temporarily. This connects the output of an integrator to the decision device. In fact, it is equivalent to sampling the output of integrator. The synchronizer then opens switch S_2 and switch S_1 is closed temperorily. This resets the integrator voltage to zero. The integrator then integrates next bit. Let us assume that one bit period 'T_b' contains integral number of cycles of the carrier. This means that the phase change occurs in the carrier only at zero crossing.

This has been shown in figure 7.10. This BPSK waveform has full cycles of sinusoidal carrier.

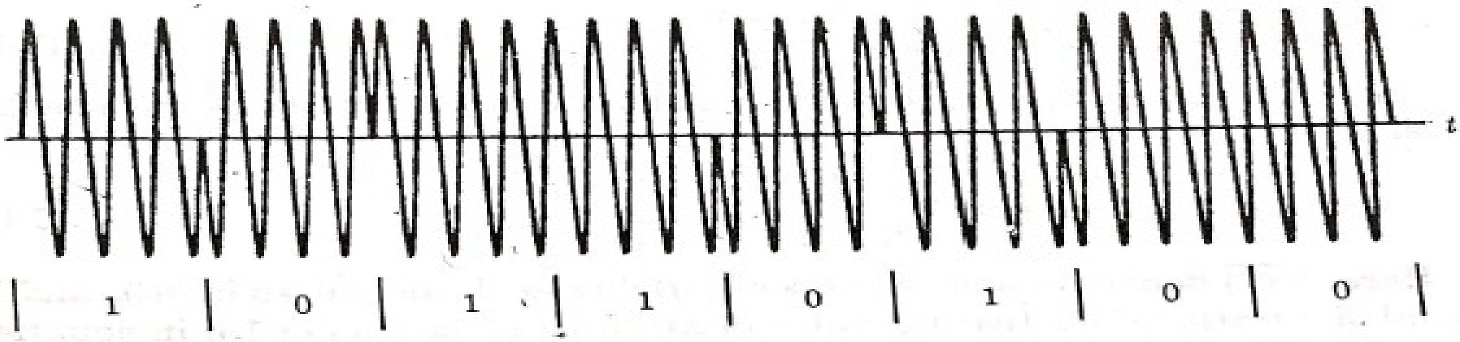


Fig. 7.10. *The BPSK waveform.*

Figure 7.12 shows the plot of this equation. This figure, thus, clearly shows that there are two lobes, one at f_c and other at $-f_c$. The same spectrum of figure 7.11 has been placed at $+f_c$ and $-f_c$. However, the amplitudes of main lobes are $\frac{PT_b}{2}$ in figure 7.12.

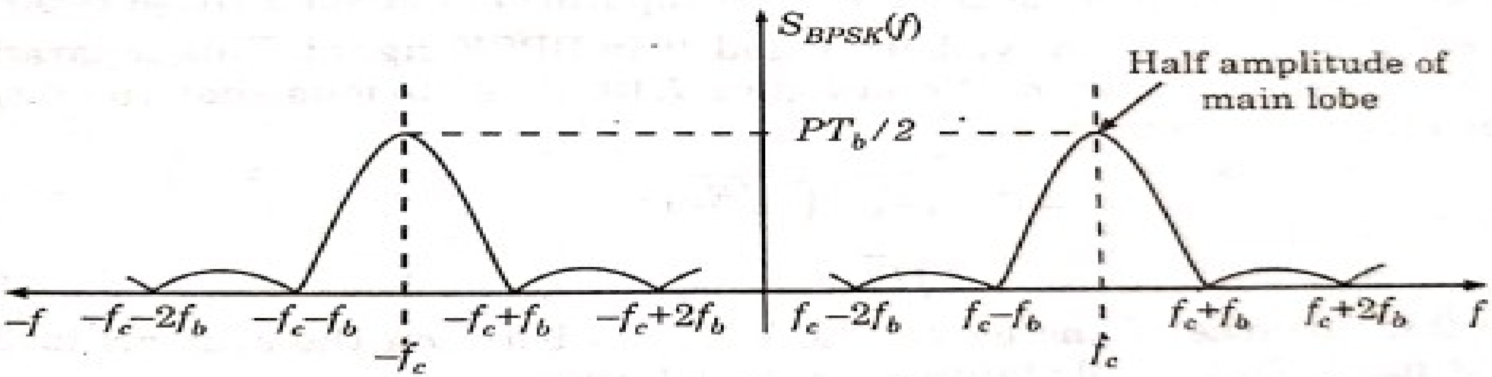


Fig. 7.12. *Plot of power spectral density of BPSK signal.*

Hence, they are reduced to half. The spectrum of $S(f)$ as well as $S_{BPSK}(f)$ extends overall the frequencies.

7.7. Coherent Binary Frequency Shift Keying (BFSK)

In binary frequency shift keying (BFSK), the frequency of the carrier is shifted according to the binary symbol. However, the phase of the carrier is unaffected. This means that we have two different frequency signals according to binary symbols. Let there be a frequency shift by Ω. Then we can write following equations.

If $\quad b(t) = \text{'1'}, \quad$ then $s_H(t) = \sqrt{2P_s} \cos (2\pi f_c + \Omega)\, t$ ...(7.25)

If $\quad b(t) = \text{'0'}, \quad$ then $s_L(t) = \sqrt{2P_s} \cos (2\pi f_c - \Omega)\, t$ ...(7.26)

Hence, there is increase or decrease in frequency by Ω. Let us use the following conversion table to combine above two FSK equations:

Table 7.1. Conversion table for BPSK representation

$b(t)$ *Input*	$d(t)$	$P_H(t)$	$P_L(t)$
1	+ 1V	+ 1V	0V
0	– 1V	0V	+ 1V

The equations (7.25) and (7.26) combinely may be written as

$$s(t) = \sqrt{2P_s} \cos \left[(2\pi f_c + d(t)\Omega)\, t \right] \quad \text{...(7.27)}$$

Hence, if symbol '1' is to be transmitted, the carrier frequency will be $f_c + \left(\frac{\Omega}{2\pi}\right)$.

If symbol '0' is to be transmitted, then the carrier frequency will be $f_c - \left(\frac{\Omega}{2\pi}\right)$.

Therefore, we have

Thus, $\quad f_H = f_c + \frac{\Omega}{2\pi} \quad$ for symbol '1' ...(7.28)

$\quad f_L = f_c - \frac{\Omega}{2\pi} \quad$ for symbol '0' ...(7.29)

7.7.1. Generation of BFSK

It may be observed from Table 7.1 that $P_H(t)$ is same as $b(t)$ and also $P_L(t)$ is inverted version of $b(t)$. The block diagram for BFSK generation is shown in figure 7.14.

The binary FSK can be viewed as two interleaved ASK signals with carrier frequencies f_{c0} and f_{c1}, respectively. Therefore, binary FSK modulator basically comprises of two balanced modulators.

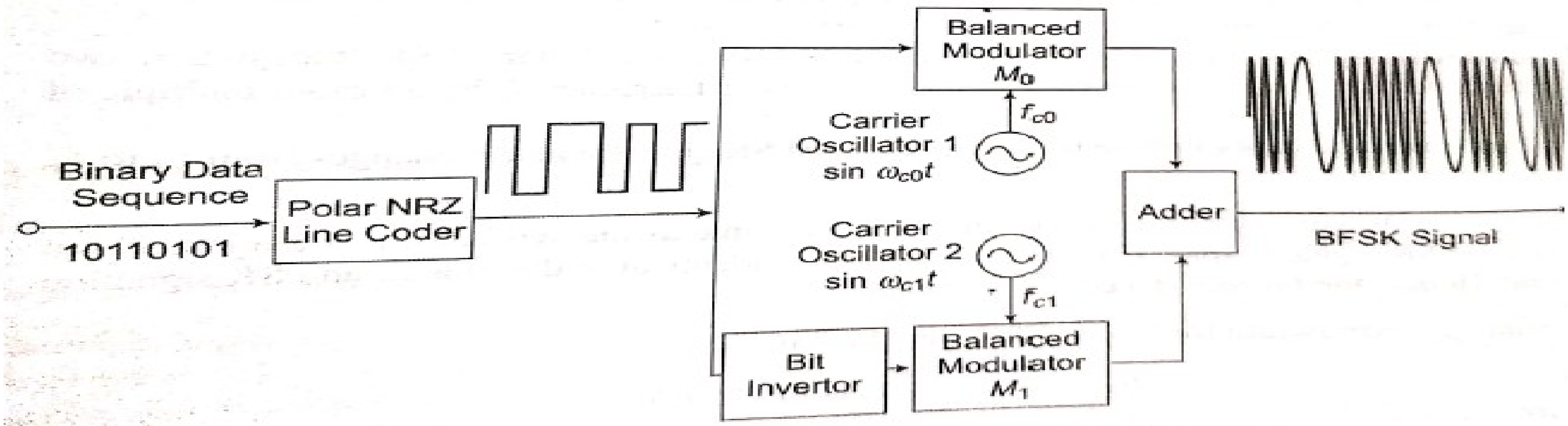


Figure 3.1.9 Functional Block Schematic of FSK Modulator

We know that input sequence $b(t)$ is same as $P_H(t)$. An inverter is added after $b(t)$ to get $P_L(t)$. The level shifter $P_H(t)$ and $P_L(t)$ are unipolar signals. The level shifter converts the '+1' level to $\sqrt{P_sT_b}$. Zero level is unaffected. Thus, the output of the level shifters will be either $\sqrt{P_sT_b}$ (if '+1') or zero (if input is zero). Further, there are product modulators after level shifter. The two carrier signals $\phi_1(t)$ and $\phi_2(t)$ are used. $\phi_1(t)$ and $\phi_2(t)$ are orthogonal to each other. In one bit period of input signal (*i.e.*, T_b), $\phi_1(t)$ or $\phi_2(t)$ have integral number of cycles.

Thus, the modulated signal is having continuous phase. Figure 7.15 shows such type of BFSK signal. The adder then adds the two signals.

Thus, the modulated signal is having continuous phase. Figure ... such type of BFSK signal. The adder then adds the two signals.

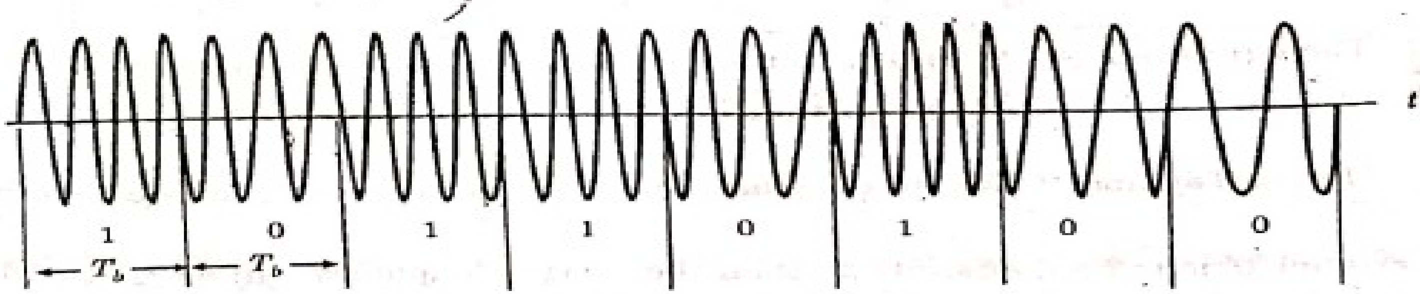


Fig. 7.15. *The BFSK signal.*

7.7.4. Detection of BFSK

Figure 7.17 shows the block diagram of a scheme for demodulation of BFSK wave using coherent detection technique. The detector consists of two correlators that are individually tuned to two different carrier frequencies to represent symbols '1' and '0'. A correlator consists of a multiplier followed by an integrator. Then, the received binary FSK signal is applied to the multipliers of both the correlators. To the other input of the multipliers, carriers with frequency f_{c1} and f_{c2} are applied as shown in figure 7.17. The multiplied output of each multiplier is subsequently passed through integrators generating output l_1 and l_2 in the two paths. The output of the two integrators are then fed to the decision making device. The decision making device is essentially a comparator which compares the output l_1 (in the upper path) and output l_2 (in the lower path). If the output l_1 produced in the upper path (associated with frequency f_{c1}) is greater than the output l_2 produced in the lower path (associated with frequency f_{c2}), the detector makes a decision in favour of symbol 1. If the output l_1 is less than l_2, then the decision making device decides in favour of symbol 0 (say). This type of digital communication receivers are also called *correlation receivers*. As discussed earlier, the detector based upon coherent detection requires phase and timing synchronisation.

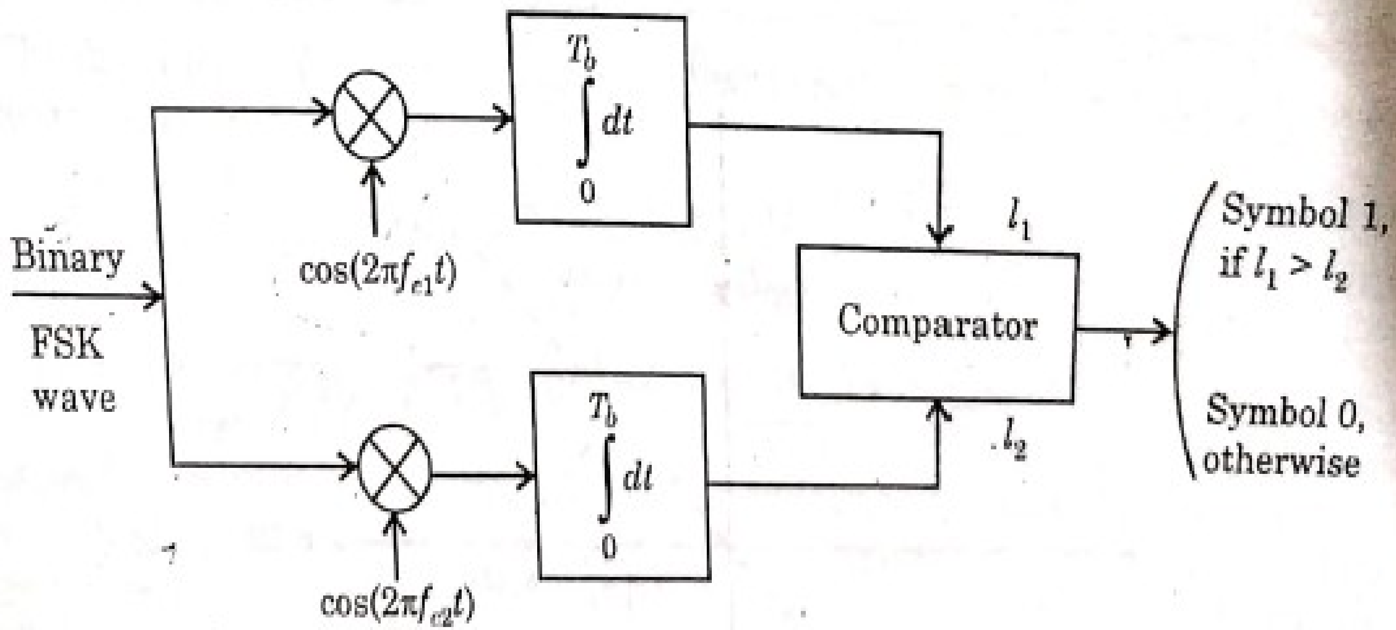


Fig. 7.17. *Block diagram of BFSK receiver (detection of BFSK).*

7.8. Non-Coherent Binary Modulation Techniques

As discussed earlier, coherent detection exploits knowledge of the carrier wave's phase reference, and thus providing the optimum error performance attainable with a digital modulation format of interest. However, when it is impractical to have knowledge of the carrier phase at the receiver, we make use of **non-coherent detection**. Thus, in this section, we shall study non-coherent binary modulation techniques *i.e.*, we shall study non-coherent detection of ASK and FSK. In the case of phase-shift keying (PSK), we cannot have "non-coherent PSK" since non-coherent means doing without phase information. However, there is a 'pseudo PSK' technique known as differential phase-shift keying (DPSK) which can be viewed as the non-coherent form of PSK.

7.9. Non-Coherent Binary Amplitude Shift Keying (ASK)

In the binary ASK case, the transmitted signal is defined as

$$s(t) = \sqrt{2P_s}\cos(2\pi f_c t)$$

Binary ASK signal can also be demodulated non-coherently using envelope detector. This greatly simplifies the design consideration required in synchronous detection. Non-coherent detection schemes do not require a phase-coherent local oscillator. This method involves some form of rectification and low pass filtering at the receiver. The block diagram of a non-coherent receiver for ASK signal has been shown in figure 7.20.

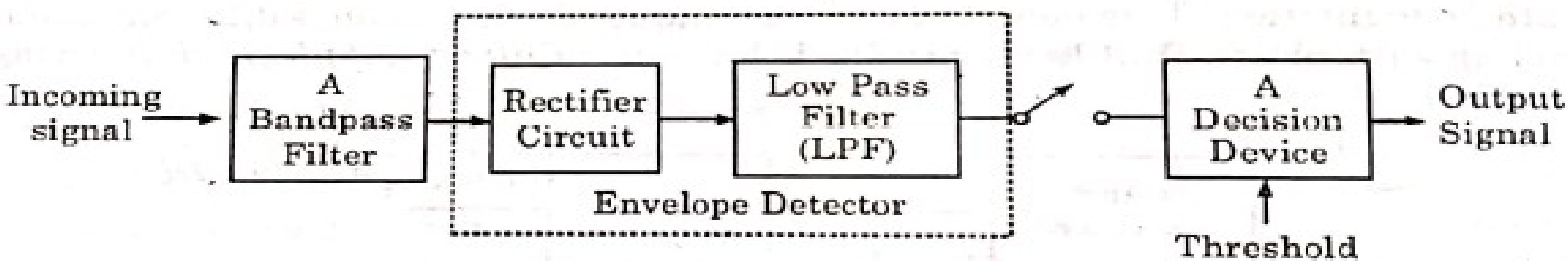


Fig. 7.20. *Non-coherent ASK detector.*

7.10. Non-Coherent Detection of FSK

Binary FSK waves may be demodulated non-coherently using envelope detector. The received FSK signal is applied to a bank of two bandpass filters, one tuned to frequency f_{c1} and the other tuned to f_{c2}. Each filter is followed by an envelope detector. The resulting outputs of the two envelope detectors are sampled and then compared with each other. The arrangement for non-coherent detection of FSK signal has been shown in figure 7.21.

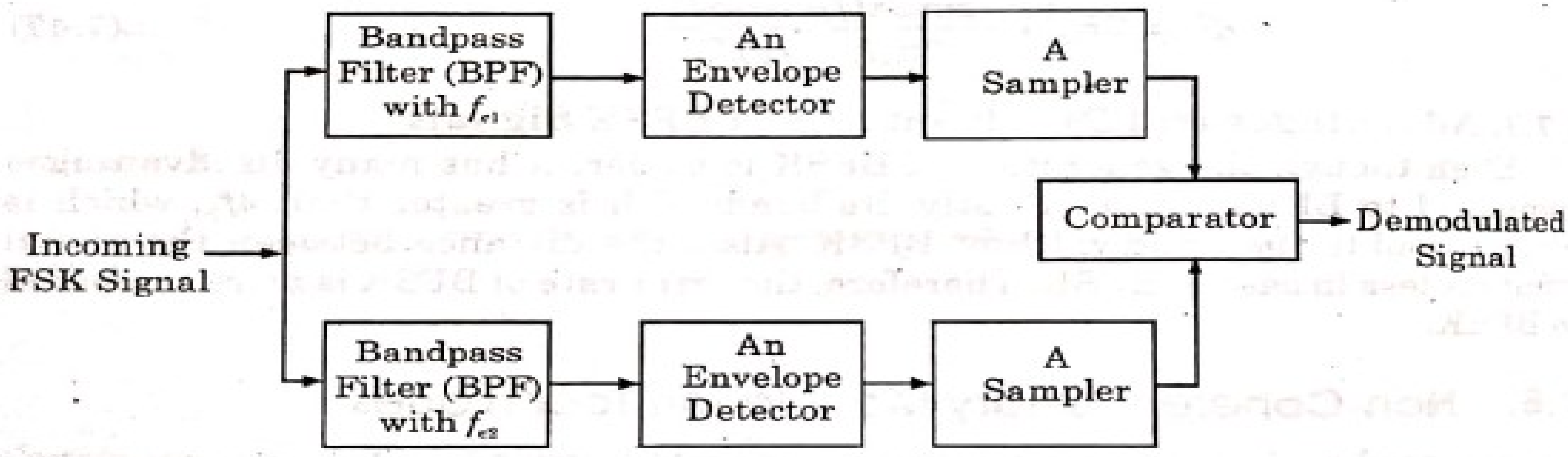


Fig. 7.21. *Non-coherent detection of FSK binary signals.*

A decision is made in favour of symbol '1' if the envelope detector output derived from the filter tuned to frequency f_{c1} is larger than that derived from the second filter. Otherwise, a decision is made in favour of the symbol 0.

7.11. Differential Phase Shift Keying (DPSK)

(U.P. Tech., Semester, Examination, 2003-2004)

We can view differential phase-shift keying as the non-coherent version of the PSK. Differential phase shift keying (DPSK) is differentially coherent modulation method. DPSK does not need a synchronous (coherent) carrier at the demodulator. The input sequence of binary bits is modified such that the next bit depends upon the previous bit. Therefore, in the receiver, the previous received bits are used to detect the present bit.

7.11.1. Generation of DPSK

Thus, in order to eliminate the need for phase synchronisation of coherent receiver with PSK, a differential encoding system can be used with PSK. The digital information content of the binary data is encoded in terms of signal transitions. As an example, the symbol 0 may be used to represent transition in a given binary sequence (with respect to the previous encoded bit) and symbol '1' to indicate no transition. This new signaling technique which combines differential encoding with phase-shift keying (PSK) is known as *differential phase-shift keying (DPSK)*.

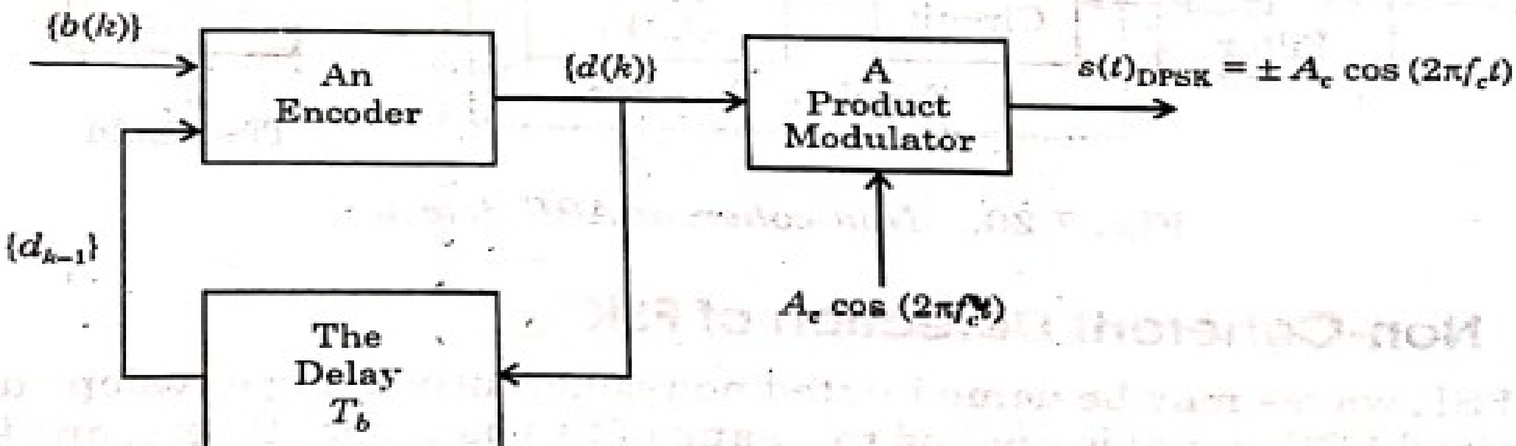


Fig. 7.22. *Illustration of the scheme to generate DPSK signals.*

A schematic arrangement for generating DPSK signal has been shown in figure 7.23. The data stream $b(t)$ is applied to the input of the encoder. The output of the encoder is applied to one input of the product modulator. To the other input of this product modulator, a sinusoidal carrier of fixed amplitude and frequency is applied. The relationship between the binary sequence and its differentially encoded version is illustrated in Table 7.2 for a assumed data sequence 0 0 1 0 0 1 0 0 1 1 1. In this illustration it has been assumed that the encoding has been done in such a way that *transition* in the given bianry sequence with respect to the previous encoded bit is represented by a symbol 0 and no *transition* by symbol '1'. It may be noted that an extra bit (symbol 1) has been arbitrarily added as an initial bit. This is essential to determine the encoded sequence. The phase of the generated DPSK signal has been shown in the third row of Table 7.2.

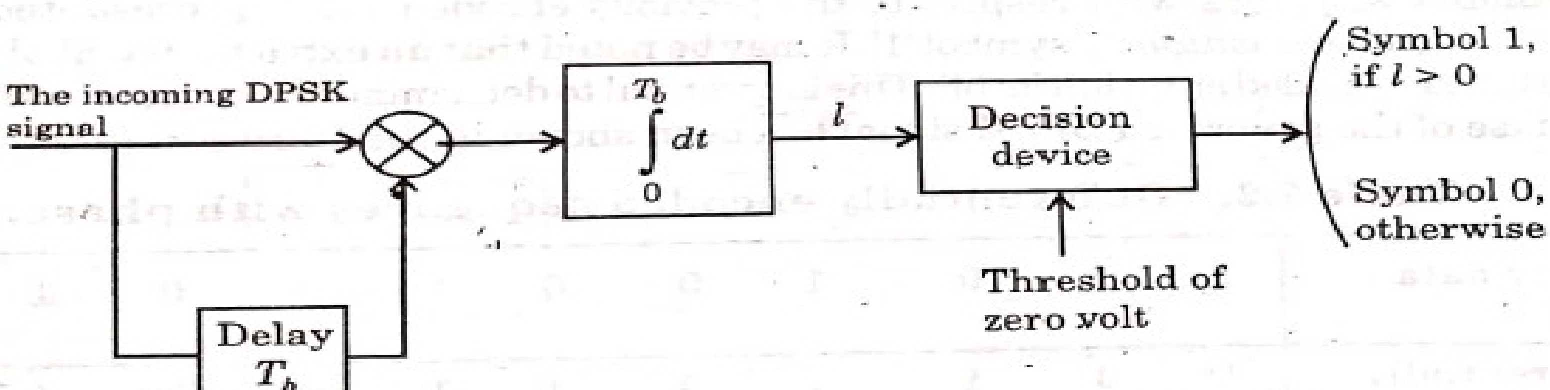
Table 7.2. Differentially encoded sequences with phase.

Binary data $\{b(k)\}$		0	0	1	0	0	1	0	0	1	1
Differentially encoded data $\{d(k)\}$	1*	0	1	1	0	1	1	0	1	1	1
Phase of DPSK	0	π	0	0	π	0	0	π	0	0	0
Shifted differentially encoded data $\{d_{k-1}\}$		1	0	1	1	0	1	1	0	1	1
Phase of shifted DPSK		0	π	0	0	π	0	0	π	0	0
Phase comparison output		−	−	+	−	−	+	−	−	+	+
Detected binary sequence		0	0	1	0	0	1	0	0	1	1

* Arbitrary starting reference bit.

7.11.2. Detection of DPSK

For detection of the differentially encoded PSK (*i.e.*, DPSK), we can use the receiver arrangement as shown in figure 7.23. The received DPSK signal is applied to one input of the multiplier. To the other of the multiplier, a delayed version of the received DPSK signal by the time interval T_b is applied. The delayed version of the received DPSK signal (in the absence of channel noise) has been shown in the 4th row of the table. The output of the difference is proportional to cos (ϕ), here ϕ is the difference between the carrier phase angle of the received DPSK signal and its delayed version, measured in the same bit interval. The phase angle of the DPSK signal and its delayed version have been shown in 3rd and 5th rows respectively. The phase difference between the two sequences for each bit interval is used to determine the sign of the phase comparator output. When $\phi = 0$, the integrator output is positive whereas when $\phi = \pi$, the integrator output is negative. By comparing the integrator output with a decision level of zero volt, the decision device can reconstruct the binary sequence by assigning a symbol '0' for negative output and a symbol '1' for positive output. The reconstructed binary data is shown in the last row of the table. It is thus seen that in the absence of noise, the receiver can reconstruct the transmitted bianry data exactly. *DPSK may be viewed as a non-coherent version of PSK. It may also be noted that the reconstruction is invariant with the choice of the initial bit in the encoded data. This has been illustrated in the example 7.1 given below.*



3.2 M-ARY DIGITAL MODULATION TECHNIQUES

In binary digital modulation techniques, we have seen that the required transmission bandwidth for a data stream whose bit duration is T_b must be nominally $2f_b$, where the input bit rate $f_b = 1/T_b$.

For example, in binary PSK, we transmit each data bit individually. Depending on whether the bit is logic 1 or logic 0, we transmit one or another of a sinusoidal carrier signal for the bit interval T_b, the sinusoids differing in phase by 180°.

In multi-level or *M*-ary digital modulation techniques, one of the *M* possible signal elements are transmitted during each bit interval of T_b seconds. We begin our discussions on Quadrature Phase Shift Keying (*QPSK*) and its variants such as *offset QPSK* and *π/4-QPSK*. Then, we move on to understand constant envelope digital modulation techniques such as Minimum Shift Keying (*MSK*) and *Gaussian MSK*. Finally, we describe hybrid digital modulation technique known as Quadrature Amplitude Modulation (*QAM*).

3.2.1 Quadrature Phase Shift Keying (QPSK)

Quadrature Phase-Shift Keying, or Quadriphase-Shift Keying (QPSK) is one of bandwidth efficient bandpass digital modulation technique that makes use of quadrature multiplexing. The QPSK system can be considered equivalent to two BPSK systems operating in parallel and having carrier signals which are of the same frequency but in phase quadrature. Just like in BPSK, in QPSK too, the information is carried in the phase of the transmitted signal. However, unlike in BPSK, in QPSK, we combine two successive bits in a bit stream to form a symbol.

With two bits, there are four possible conditions: 11, 10, 01, and 00. Depending on which of the four two-bit symbols (called *dibits*) develops in the input data stream, we transmit one or another of four sinusoids of duration $2T_b$, the sinusoids differing in phase by 90° or 180° with another. Thus, in a QPSK modulator, each dibit code generates one of the four possible output phases (+45°, +135°, –45°, –135°), hence, the name "quadrature" meaning "four" phases.

Mathematically, the QPSK signal for one symbol duration, consisting of two bits each, can be expressed as

$$
\boxed{S_{\text{QPSK}}(t) = \begin{cases} \sin(2\pi f_c t - 3\pi/4) & \text{for } 0\,0 \\ \sin(2\pi f_c t - \pi/4) & \text{for } 0\,1 \\ \sin(2\pi f_c t + 3\pi/4) & \text{for } 1\,0 \\ \sin(2\pi f_c t + \pi/4) & \text{for } 1\,1 \end{cases}}
$$

In general, we can say that QPSK is an M-ary constant-amplitude quadrature PSK digital modulation scheme in which number of bits is two ($n = 2$) and the number of signaling elements (symbols) are four, i.e., $M = 4$ (hence, sometimes known as "quaternary" meaning "4", 4-ary PSK).

Table 3.2.1 Symbols, Bits, and Phase Shift in QPSK Signal

Symbol	Binary Input	Phase Shift in QPSK Signal
s_1	0 0	$-135°$ or $-3\pi/4$ radians
s_2	0 1	$-45°$ or $-\pi/4$ radians
s_3	1 0	$+135°$ or $3\pi/4$ radians
s_4	1 1	$+45°$ or $\pi/4$ radians

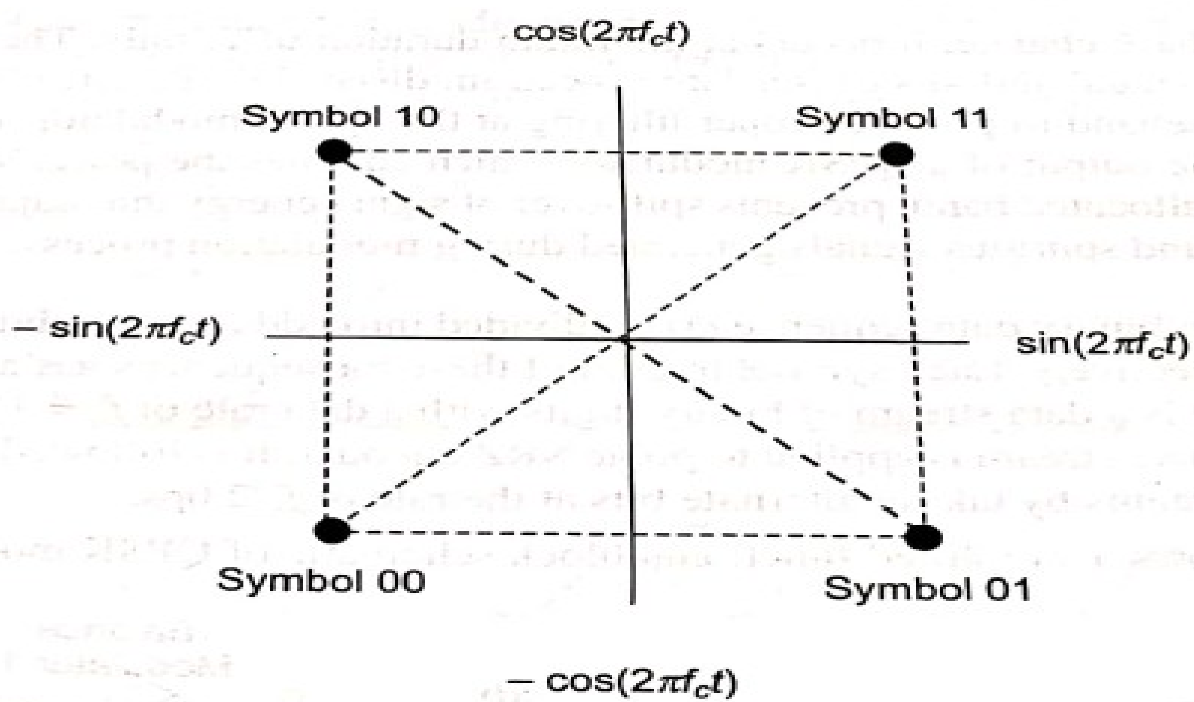


Figure 3.2.1 Constellation Diagram of QPSK Signal

It is seen that the transmitted QPSK signal has to go to zero amplitude momentarily as it makes 180° transition (from 11 to 00 or from 10 to 01 or vice versa). Whereas it makes only 90° transition (from 11 to 10, 10 to 00, or 00 to 01, or 01 to 11, or vice versa).

The rate of change at the output (baud) is equal to one-half the input bit rate $f_b/2$ (i.e., two input bits produce one output phase change). As a result, each bit in QPSK can be transmitted using half the bandwidth that of required to transmit BPSK signal, i.e., QPSK signal requires minimum Nyquist transmission bandwidth equal to input bit rate f_b only, a significant improvement as compared to that of BPSK.[12]

In QPSK, the input binary data sequence $d(t)$ is divided into odd and even bit sequences, that is, $d_1(t)$ and $d_2(t)$ respectively. Each symbol in both of these bit sequences has a period of $T_s = 2T_b$ seconds. The input is a data stream of binary digits with a data rate of $f_b = 1/T_b$, where T_b is the bit duration. This data stream is applied to polar NRZ encoder. It is followed by conversion into two separate bit streams by taking alternate bits at the rate of $f_b/2$ bps.

Figure 3.2.3 shows a simplified functional block schematic of QPSK modulator.

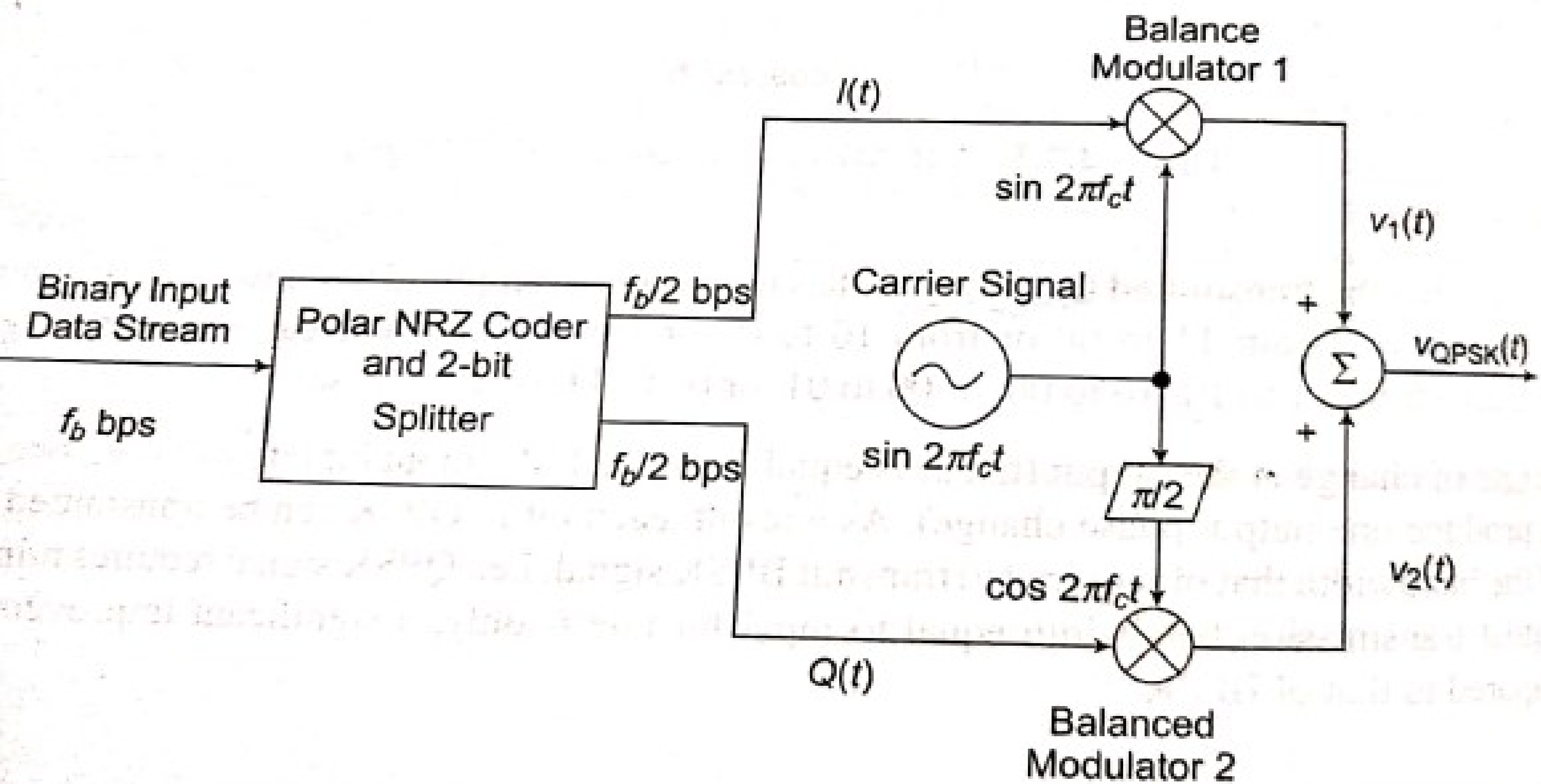


Figure 3.2.3 Functional Block Schematic of QPSK Modulator

QPSK is characterized by two parts of the baseband data signal: the in-phase signal $I(t)$ and the Quadrature signal $Q(t)$. QPSK modulator is a combination of two BPSK modulators. Thus, at the output of the I-balanced modulator 1, there are two types of phases produced, which are $+\cos(2\pi f_c t)$ and $-\cos(2\pi f_c t)$. Similarly, at the output of the Q-balanced modulator 2, again two phases are produced as $+\sin(2\pi f_c t)$ and $-\sin(2\pi f_c t)$. When the linear adder combines these two groups of orthogonal BPSK modulated signals, there will be four possible phases which are

$+\cos(2\pi f_c t) + \sin(2\pi f_c t)$, $-\cos(2\pi f_c t) + \sin(2\pi f_c t)$,

$+\cos(2\pi f_c t) - \sin(2\pi f_c t)$, and $-\cos(2\pi f_c t) - \sin(2\pi f_c t)$.

- The binary input data stream of 1s and 0s is encoded into a polar non-return-to-zero (NRZ) stream of 1s and −1s.
- The *bit splitter* (also known as 2-bit demultiplexer) segregates the odd and even indexed polar binary digits.
- All the odd-indexed bits are directed to the upper in-phase channel (I-channel) where they phase-modulate the in-phase carrier signal $\sin(2\pi f_c t)$ using a balanced modulator 1.
- All the even-indexed bits are directed to the lower quadrature-phase channel (Q-channel) where they phase-modulate the quadrature-phase carrier signal $\cos(2\pi f_c t)$ which is $\pi/2$-shift version of the in-phase carrier signal $\sin(2\pi f_c t)$ using a balanced modulator 2.
- The outputs of balanced modulators, $v_1(t)$ and $v_2(t)$ are basically two BPSK signals which are linearly added in summer.
- The output QPSK signal can be expressed as:

$$v_{QPSK}(t) = \frac{1}{\sqrt{2}}\left[I(t)\sin(2\pi f_c t) - Q(t)\cos(2\pi f_c t)\right]$$

Note... *When $d_1(t)$ and $d_2(t)$ signals modulate the sinusoidal carrier signals of frequency f_c, the PSD is shifted to $\pm f_c$*

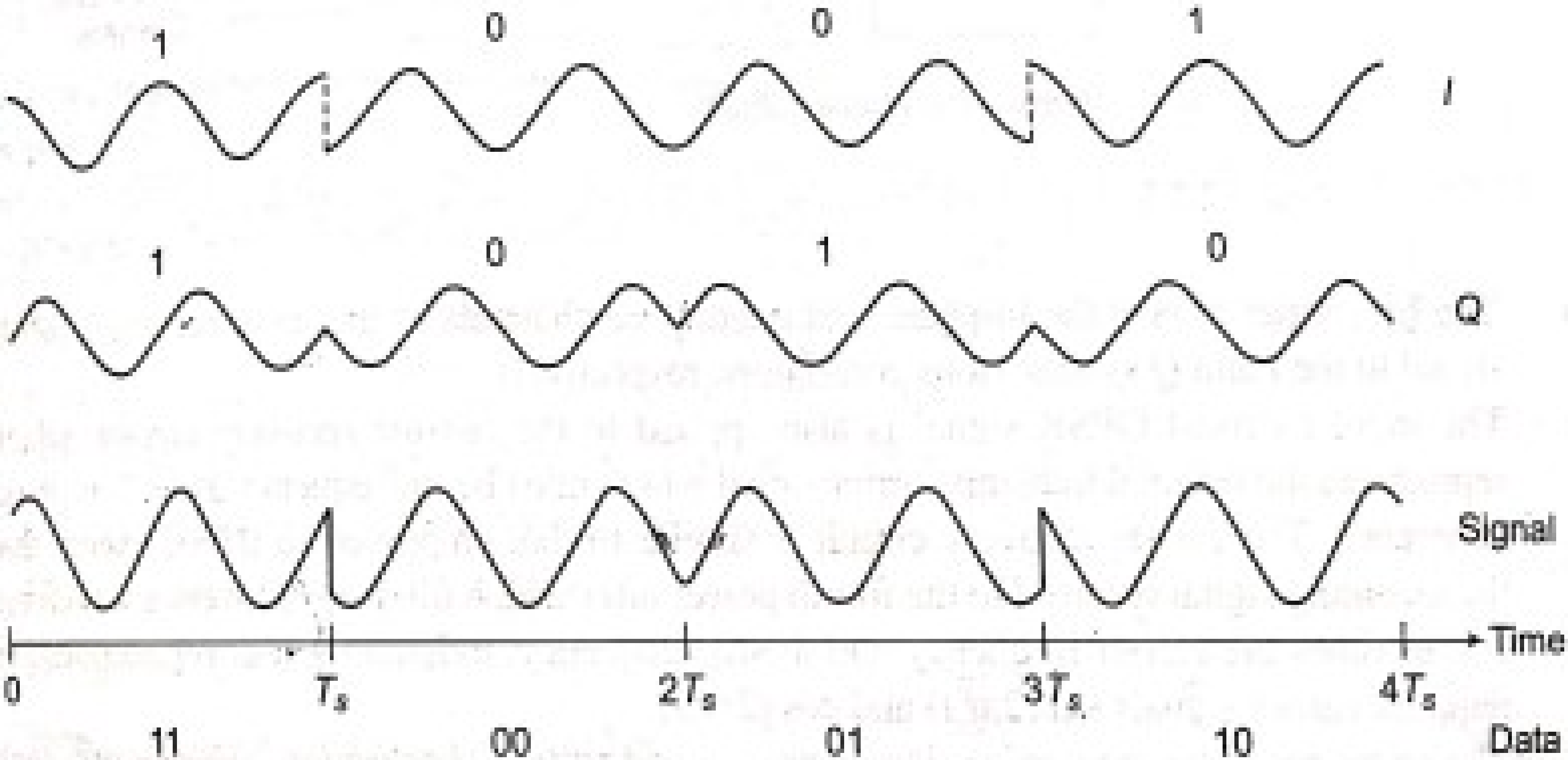


Figure 3.2.4 QPSK Signal in Time Domain *(Ts: Symbol duration)*

The binary data shown here is 1 1 0 0 0 1 1 0. The odd bits contribute to the in-phase (I-channel) component as **1** 1 **0** 0 **0** 1 **1** 0, and the even bits contribute to the quadrature-phase (Q-channel) component as 1 **1** 0 **0** 0 **1** 1 **0**. Note that there are abrupt changes in phase at some of the symbol-period boundaries.

The in-phase and quadrature channels of the coherent QPSK detector are typical BPSK coherent detectors.[13]

Figure 3.2.5 shows a simplified functional block schematic of coherent QPSK detector.

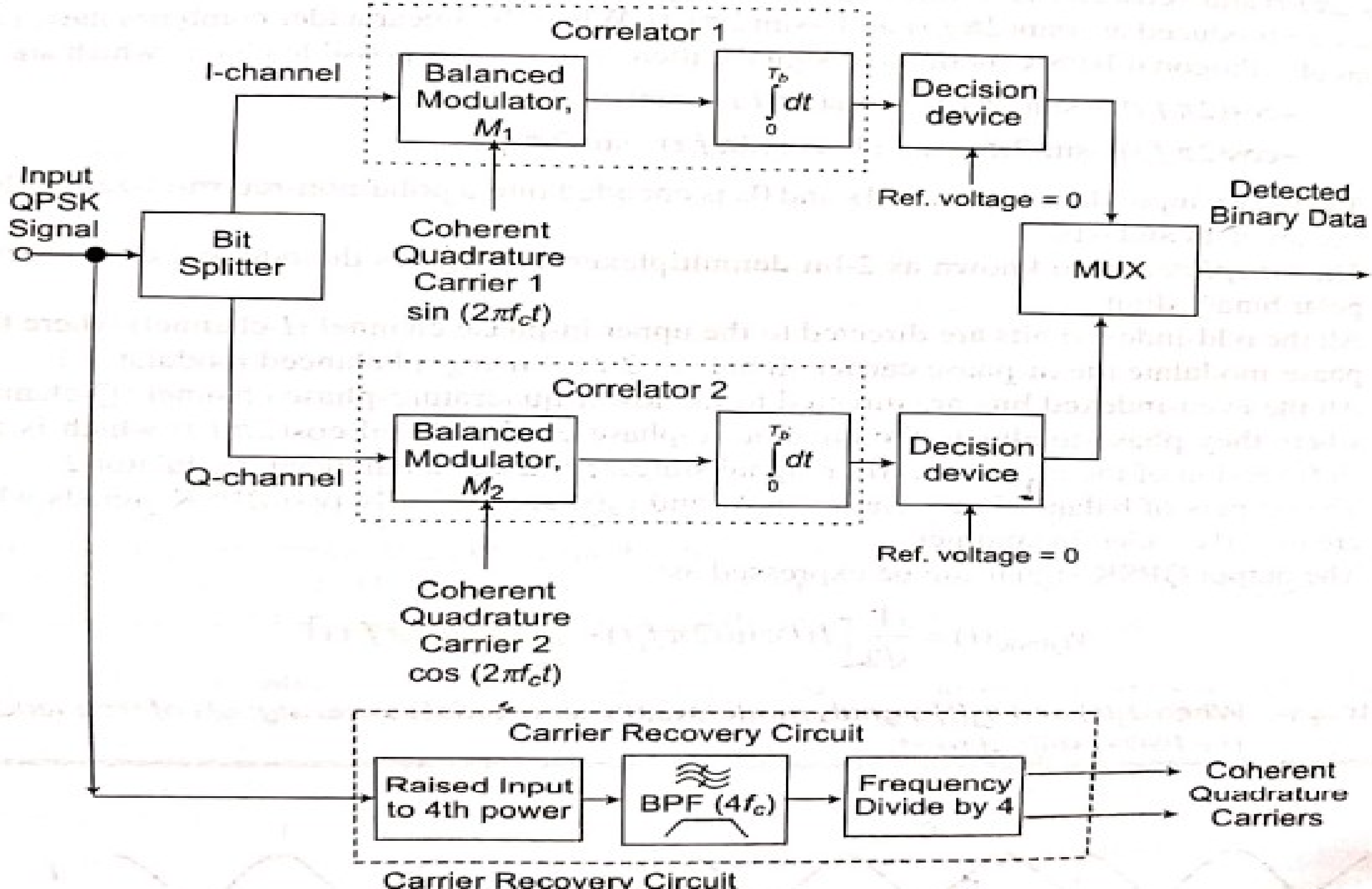


Figure 3.2.5 Functional Block Schematic of Coherent QPSK Detector

- The bit splitter directs the in-phase and quadrature channels of the input received QPSK signal to the *I* and *Q* synchronous correlators, respectively.
- The input received QPSK signal is also applied to the *carrier recovery circuit*, which reproduces the original transmit carrier signal which must be in frequency as well as phase coherence. The carrier recovery circuit is similar to that employed in BPSK except that the incoming signal is raised to the fourth power after which filtering recovers a waveform at four times the carrier frequency and finally frequency division by four regenerates the required carrier signals sin $(2\pi f_c t)$ and cos $(2\pi f_c t)$.
- The recovered coherent carrier signals are applied to two synchronous *correlators*, each comprising of *balanced modulator* and an *integrator*.
- The *integrator* circuits integrate the signals over two bit intervals, that is $T_s = 2T_b$.

[13] The bandwidth required by QPSK is one-half that of BPSK for the same BER. The transmission data rate in QPSK is higher because of reduced bandwidth. The variation in QPSK signal amplitude is not significant; hence, carrier power almost remains constant. A typical differential QPSK (DQPSK) uses phase shifts with respect to the phase of the previous symbol

If the output of the *correlator I* is greater than reference voltage of 0 in the decision device, then the in-phase channel decision device decides that the detected bit is a binary 1, and if it is less than 0, then it decides that the detected bit is a binary 0.
Similarly, if the output of the *correlator* 2 is greater than reference voltage of 0 in the decision device, then the quadrature channel's decision device decides that the detected bit is a binary 1, and if it is less than 0, then it decides that the detected bit is a binary 0.
These binary digits from the outputs of the decision devices in the two channels are then multiplexed in the bit combining circuit (MUX).
The final output is detected binary data of coherent QPSK detector which is an estimate of the transmitted binary data sequence with minimum possible probability of error for additive white Gaussian noise (AWGN) since it is a correlation reception.

In a QPSK digital communication system, the bit rate of a bipolar NRZ data sequence is 1 Mbps and carrier frequency of transmission is 100 MHz. Determine the symbol rate of transmission and the bandwidth requirement of the communications channel.

Solution Given bit rate of a bipolar NRZ data sequence, $f_b = 1$ Mbps or 1×10^6 bps

Therefore, bit period, $T_b = \frac{1}{f_b} = \frac{1}{1 \text{ Mbps}} = 1\ \mu\text{s}$

In QPSK digital communication system, two successive binary bits form one symbol.

That is, the symbol duration, $T_s = 2T_b = 2\ \mu\text{s}$

Therefore, symbol rate of transmission $= \frac{1}{T_s} = \frac{1}{2\ \mu\text{s}} = 500$ k symbols/second **Ans.**

Minimum bandwidth, $B_{QPSK} = f_b = 1$ MHz **Ans.**

SOLVED EXAMPLE 3.2.2 **QPSK Bandwidth Requirement**

Consider a QPSK system having a bit rate of 9600 bps. Determine the bandwidth required by the QPSK signal using raised-cosine filter with roll-off factor of 0.35 and 0.5.

Solution We know that transmitted signal bandwidth = Symbol rate × (1 + roll-off factor)

Given bit rate of a bipolar NRZ data sequence, $f_b = 9600$ bps

In QPSK, symbol rate = ½ × bit rate = ½ × 9600 bps = 4800 symbols per second

(a) For given roll-off factor of 0.35, required bandwidth = 4800 × (1 + 0.35)
 Hence, required bandwidth = 6480 Hz **Ans.**

(b) For given roll-off factor of 0.5, required bandwidth = 4800 × (1 + 0.5)
 Hence, required bandwidth = 7200 Hz **Ans.**

3.2.2 Offset QPSK (OQPSK)

In the QPSK, the phase changes of the carrier signal may be by ±90°, or sometimes even b ±180°, depending upon whether the sign change of the in-phase and quadrature componen occurs simultaneously or not. From the signal constellation diagram of QPSK, it can be see that

- Whenever adjacent dibits in the binary data sequence differ only in one of the digits (e.g 1 1 and 0 1, or 1 1 and 1 0, or 0 0 and 1 0, or 0 0 and 0 1), the transition involves a chang of carrier phase by ±90° only.
- On the other hand, whenever adjacent dibits in the binary data sequence differ in both th binary digits (e.g., 1 1 and 0 0, or 0 1 and 1 0), the transition involves a change of carrie phase by ±180°.

Such sudden changes in the phase of the carrier signal can result in

- Reduction of the amplitude of the QPSK signal when it is passed through a low-pass filter during transmission (before detection).
- The resulting amplitude reduction of the QPSK signal can lead to errors during detection process.
- The changes in carrier phase by ±180° in particular, cause considerable reduction in the envelope amplitude and need to be avoided.
- Large amplitude fluctuation causes significant reduction in the desired quality in digital communication systems.

Offset Quadrature Phase Shift Keying (OQPSK) is a variant of quadrature phase shift keying modulation using four different values of the phase to transmit. It is sometimes called Staggered Quadrature Phase Shift Keying (SQPSK). By offsetting the timing of the odd and even bits by one bit-period, or half a symbol-period, the in-phase and quadrature-phase components will never change at the same time. So, the carrier-phase changes are confined to ±90° only. The signal doesn't cross zero, because only one bit of the symbol is changed at a time. So the extent of amplitude fluctuations is reduced, thereby reducing the probability of occurrence of symbol errors in the detection process. Theoretically, the average probability of symbol error is exactly the same for QPSK and OQPSK for coherent detection.

Offset QPSK (OQPSK) is a modified form of QPSK where the bit waveforms on the I and Q channels are offset or shifted in phase from each other by one-half of a bit interval. This means that OQPSK is obtained by introducing a shift or offset equal to one bit delay (T_b) in the bit stream for the quadrature component with respect to the in-phase component bit stream.

Mathematically, the OQPSK signal can be expressed as

$$S_{OQPSK}(t) = \frac{1}{\sqrt{2}}\left[I(t)\cos(2\pi f_c t) - Q(t)\sin(2\pi f_c t)\right]$$

The changes in the I-channel occur at the midpoints of the Q-channel bits and vice versa. There is never more than a single bit change in the dibit code. There is never more than a 90° phase shift in the output phase. In other words, the modulated OQPSK signal transitions are ±90 degree maximum. It has no 180 degree phase shift and this result in reduction of out-of-band radiations. However, the abrupt phase transitions still remain.

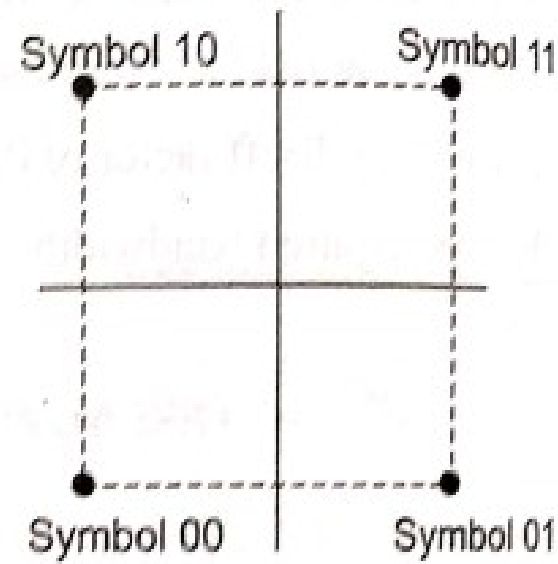


Figure 3.2.6 Constellation Diagram of OQPSK Signal

The minimum bandwidth is twice that of a QPSK for a given transmission data rate.

Figure 3.2.7 shows a simplified functional block schematic of a OQPSK modulator.

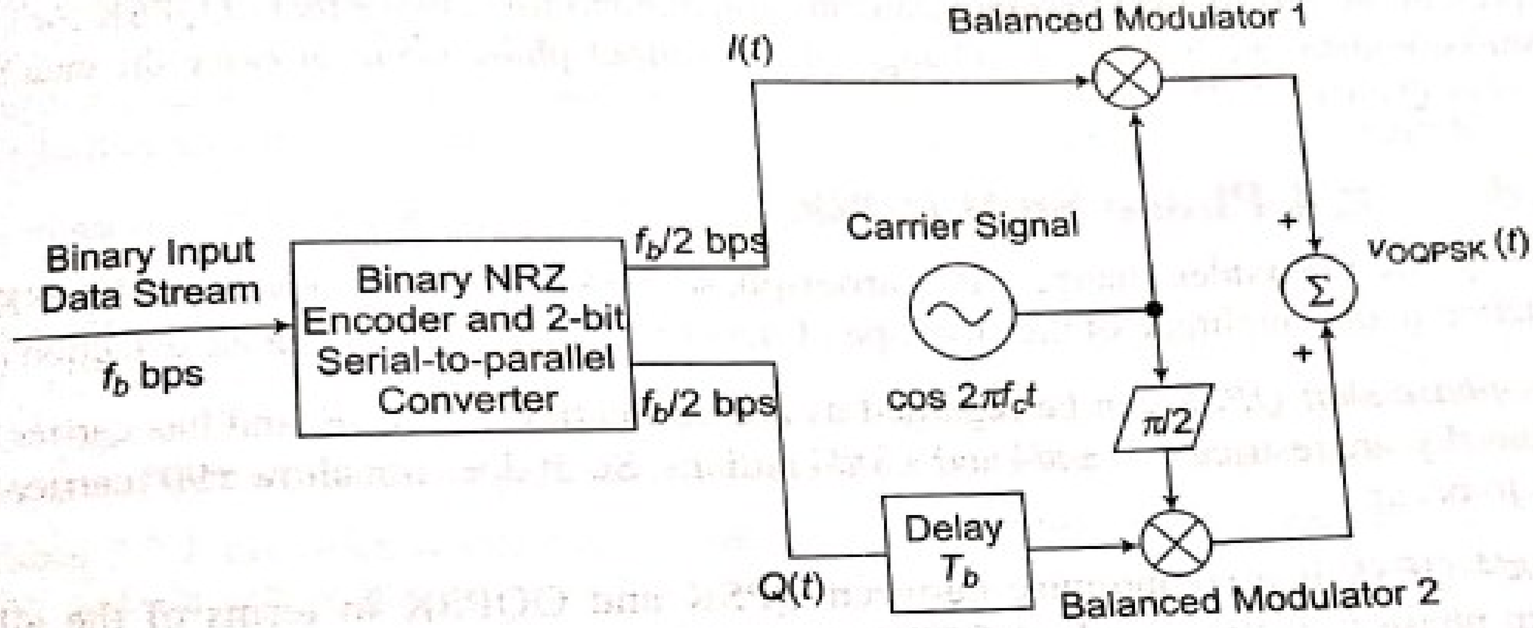


Figure 3.2.7 Functional Block Schematic of OQPSK Modulator

- The binary input data stream is applied to a binary NRZ encoder for line encoding.
- Then it is separated into odd bit and even bit.
- The quadrature data sequence will start with a delay of one bit period after the first odd bit is available in the in-phase data sequence.
- This delay is called *offset*, which leads to *offset QPSK*.
- The in-phase data sequence and carrier signal are applied to *I*-path balanced modulator.
- The quadrature data sequence and $\pi/2$-phase shifted version of carrier signal are applied to *Q*-path balanced modulator.
- Their outputs are then combined together to generate OQPSK signal.

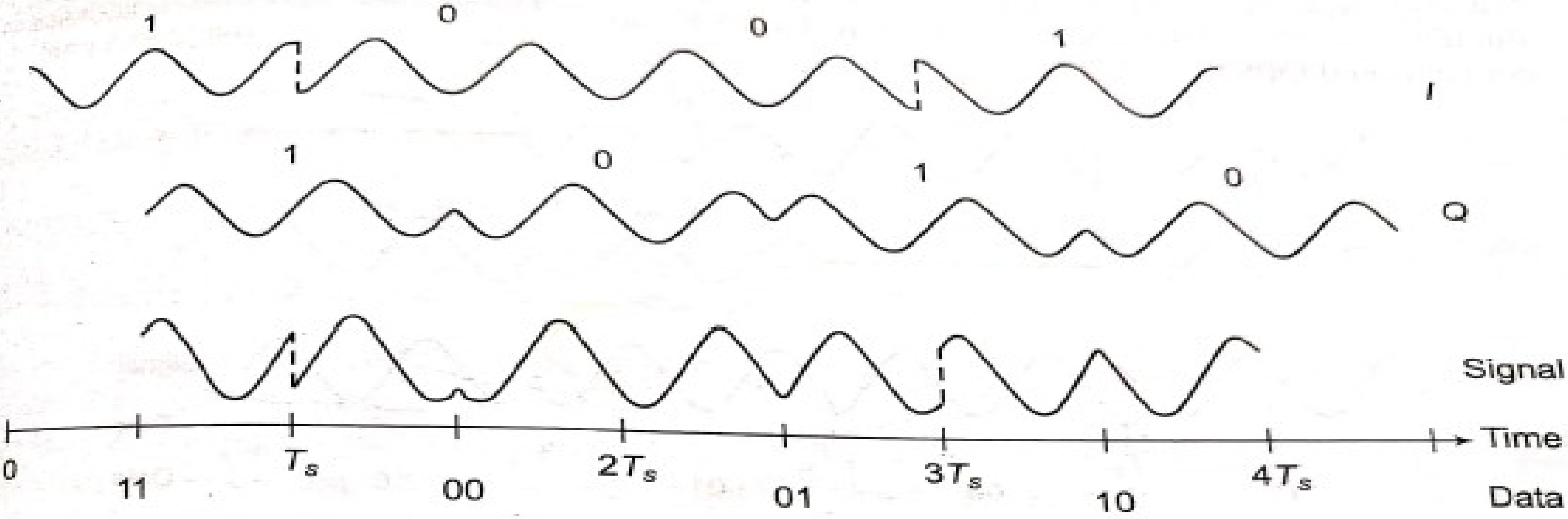


Figure 3.2.8 OQPSK Signal in Time Domain (Ts: Symbol duration)

The binary data shown here is 1 1 0 0 0 1 1 0. The odd bits contribute to the in-phase (I-channel) component as **1** 1 **0** 0 **0** 1 **1** 0, and the even bits with the half-period offset contribute to the quadrature-phase (Q-channel) component as 1 **1** 0 **0** 0 **1** 1 **0**. Note that there are abrupt changes in phase at some of the symbol-period boundaries but confined to ±90° only.

The modulated OQPSK signal has no 180° phase shift. Instead, it has ±90° maximum transition, which may result in reduction of out-of-band radiation. However, abrupt phase transitions are still present. Moreover, the minimum transmission bandwidth is twice that of QPSK for a given transmission data rate because the changes in the output phase occur at twice the data rate in either *I* or *Q* channels.[14]

3.2.4 Minimum Shift Keying (MSK)

Coherent binary FSK receivers use the phase information contained in the received signal only for synchronization, not for improved detection.

Minimum Shift Keying (MSK) is a special case of binary continuous-phase FSK modulation technique in which the change in carrier frequency from symbol *0* to symbol *1* or vice versa is exactly equal to one-half the bit rate of input data signal.

As a form of binary FSK, the MSK signal can be expressed by

$$\boxed{v_{MSK}(t) = \begin{cases} \sqrt{\dfrac{2E_b}{T_b}} \cos\left[2\pi f_1 t + \theta(0)\right] & \text{for binary 1} \\ \sqrt{\dfrac{2E_b}{T_b}} \cos\left[2\pi f_2 t + \theta(0)\right] & \text{for binary 0} \end{cases}}$$

where E_b is the transmitted signal energy per bit, T_b is the bit duration, the phase $\theta(0)$ denotes the value of the phase at time $t = 0$.

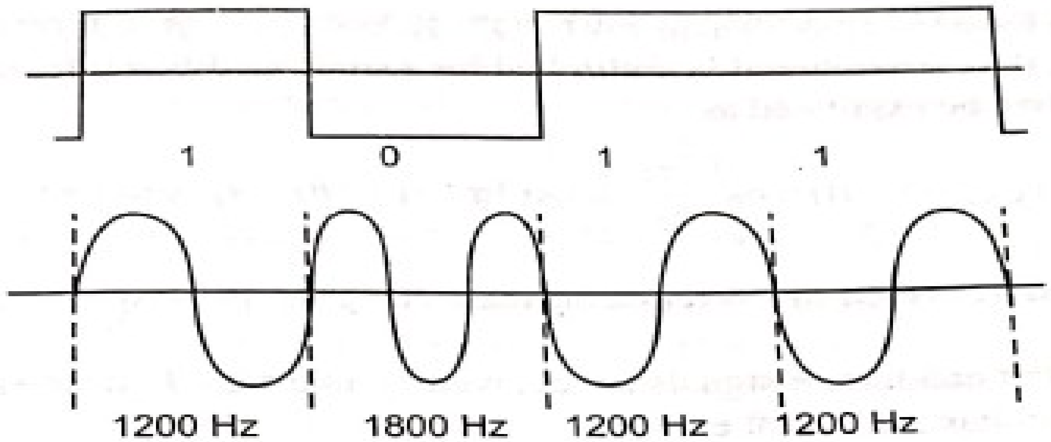


Figure 3.2.12 MSK Signal at 1200 baud for NRZ data 1 0 1 1

As an example, a 1200 baud baseband MSK data signal is composed of 1200 Hz and 1800 Hz frequencies for a binary 1 and 0, respectively. It can be seen that the modulated carrier contains no phase discontinuities and frequency changes occur at the carrier zero crossings only. Thus, MSK is a continuous phase modulation technique having modulation index ($\Delta f \times T_b$) = 0.5, where Δf is the difference in frequency for binary 1 and 0, and T_b is the bit duration. The two frequencies f_1 and f_2 satisfy the following equations:

$$f_1 = f_c + \frac{1}{4T_b}; \text{ and } f_2 = f_c - \frac{1}{4T_b}$$

Therefore, bandwidth of MSK signal, $BW_{MSK} = |f_1 - f_2|$

MSK signal produces orthogonal signaling and the phase information of the received signal is used in the detection process to improve the noise performance.

The PSD of MSK signal is shown in Figure 3.2.13.

7.33 shows the normalized spectral densities of MSK and Q... means maximum amplitudes of signals are scaled with respect to '1'.

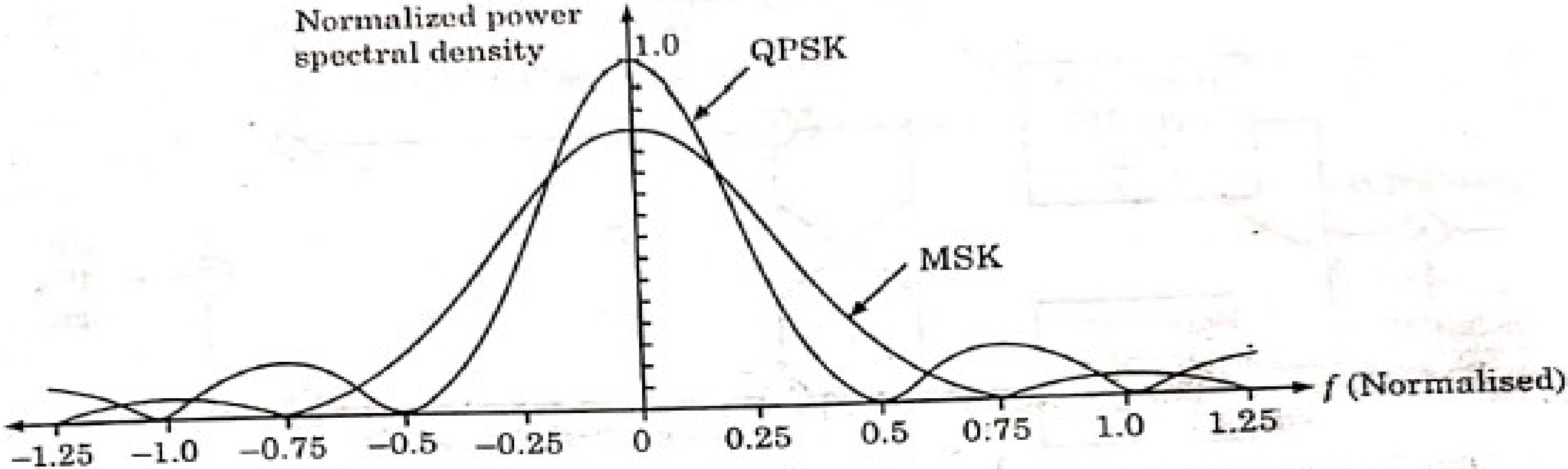


Fig. 7.33. *Power spectral densities (psd) of MSK and QPSK.*

The above plots show that the main lobe in MSK is wider than QPSK. The sides lobes in MSK are very small compared to QPSK.

Bandwidth Calculation of MSK

From figure 7.33, we observe that the width of main lobe in MSK is ± 0.75 *i.e.*,

$$fT_b = \pm 0.75$$

or

$$f = \pm 0.75 \, f_b$$

Hence, bandwidth will be equal to width of the main lobe *i.e.*,

$$BW = 0.75 \, f_b - (-0.75 \, f_b) = 1.5 \, f_b \qquad ...(7.100$$

Thus, the BW of MSK is higher than that of QPSK.

7.14.3. Generation of MSK

Figure 7.34 shows the block diagram of MSK transmitter. The two sinusoidal signals $\sin(2\pi f_c t)$ and $\cos(2\pi t/4T_b)$ are mixed (*i.e.*, multiplied). The bandpass filters then pass only sum and difference components $f_c + \frac{f_b}{4}$ and $f_c - \frac{f_b}{4}$. The outputs of bandpass filters (BPFs) are then added and subtracted such that two signals $x(t)$ and $y(t)$ are generated. Signal $x(t)$ is multiplied by $\sqrt{2P_s}\, b_0(t)$ and $y(t)$ is multiplied by $\sqrt{2P_s}\, b_e(t)$. The outputs of the multiplies are then added to give final MSK signal. Thus the block diagram of figure 7.34 is the step to step implementation of equation (7.73).

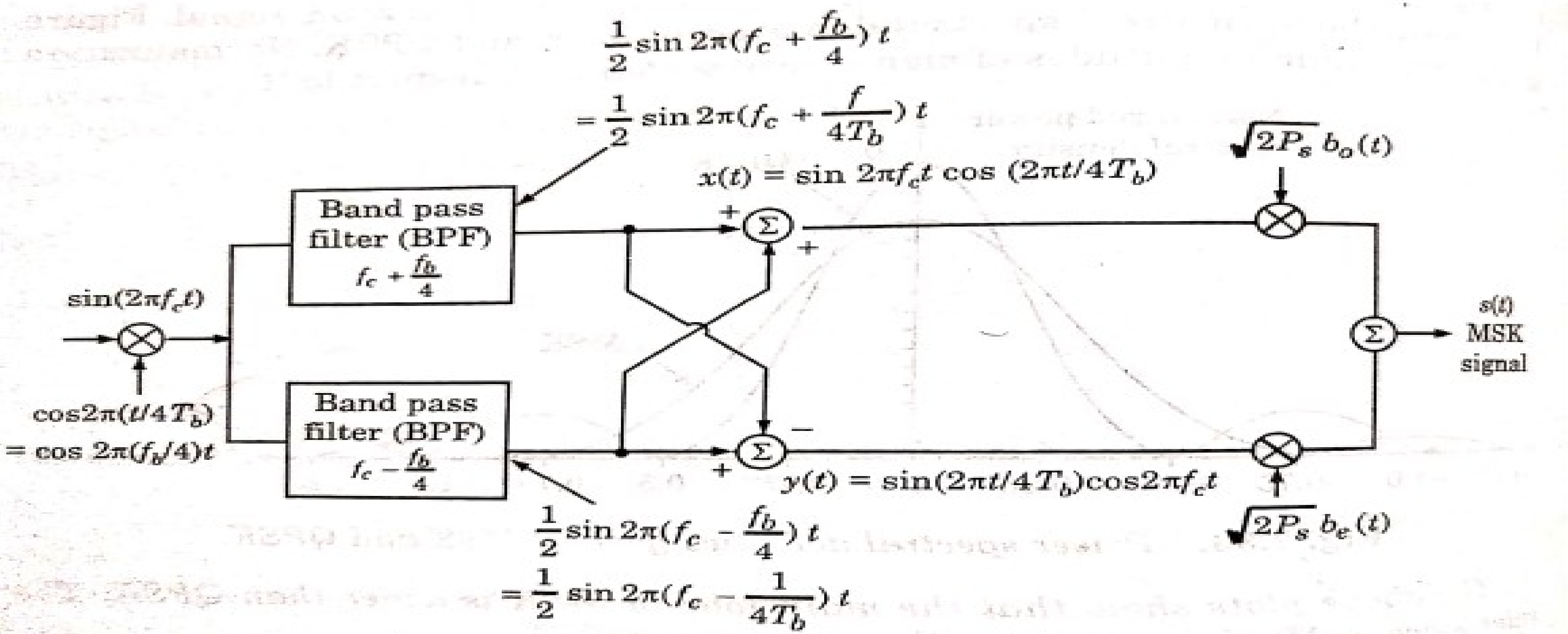


Fig. 7.34. *MSK transmitter block diagram.*

7.14.4. Reception of MSK (*i.e.* Detection of MSK)

Figure 7.35 shows the block diagram of MSK receiver. MSK uses synchronous detection. The signals $x(t)$ and $y(t)$ are multiplied with the received MSK signal. Here $x(t)$ and $y(t)$ have same values as shown in transmitter block diagram of figure 7.35. The outputs of the multipliers are $b_e(t)$ and $b_0(t)$. The integrators integrate over the period of $2T_b$. For the upper correlator, the sampling switch samples output of integrator at $t = (2k + 1)T_b$. Then the decision device decides whether $b_0(t)$ is + 1 or − 1. Similarly, lower correlator output is $b_e(t)$. The outputs of two decision devices are staggered by T_b. The switch S_3 operates at $t = kT_b$ and simply multiplexes the two correlator outputs.

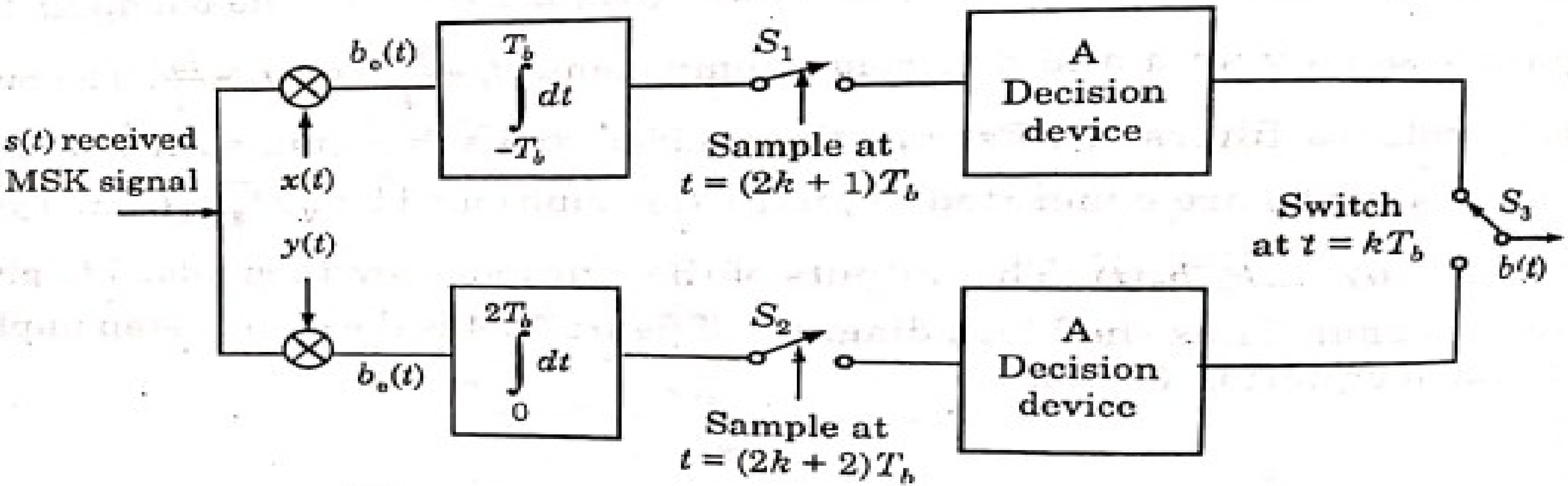
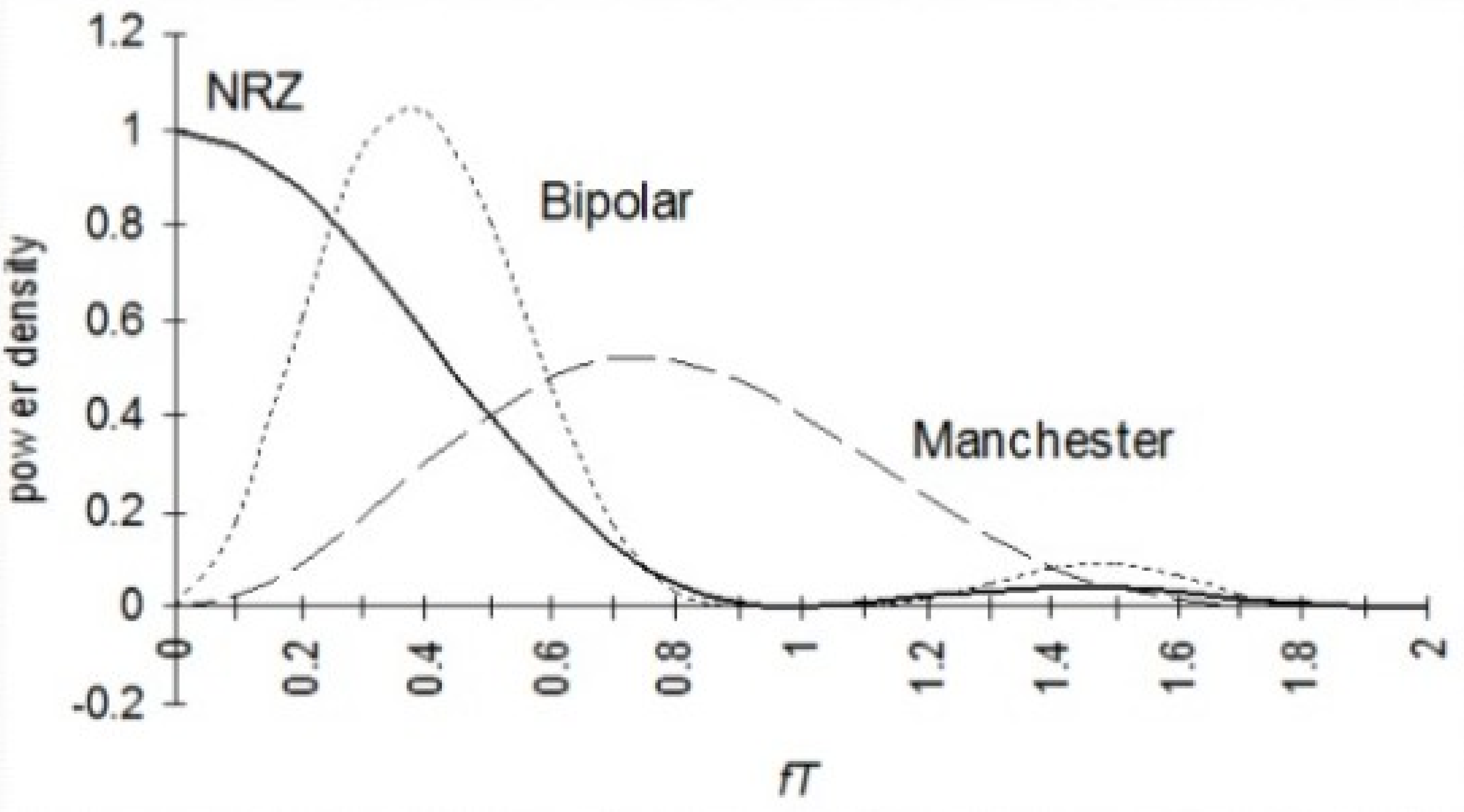


Fig. 7.35. *MSK receiver block diagram.*

PSD of Lines Codes:



Comparison of Line Codes:

Sr. No.	Parameters	Polar RZ	Polar NRZ	AMI	Manchester
1	Transmission of DC component	YES	YES	NO	NO
2	Signaling Rate	1/Tb	1/Tb	1/Tb	1/Tb
3	Noise Immunity	LOW	LOW	HIGH	HIGH
4	Synchronizing Capability	Poor	Poor	Very Good	Very Good
5	Bandwidth Required	1/Tb	1/2Tb	1/2Tb	1/Tb
6	Crosstalk	HIGH	HIGH	LOW	LOW

Various line-coding techniques can also be distinguished based on polarity of voltage levels used to represent the binary digits. Accordingly, line-coding techniques can be broadly classified in three types, as depicted in Figure 2.1.7.

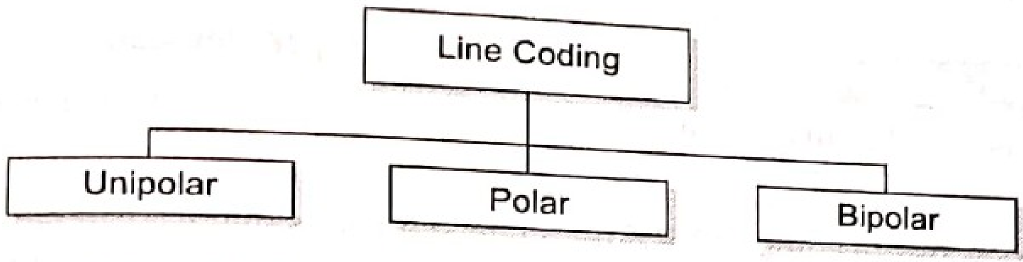


Figure 2.1.7 Classification of Line-Coding Techniques

Unipolar (represented by only one level +V or –V). Thus, Unipolar line coding can be Unipolar Non-Return-to-Zero (**UP-NRZ**) and Unipolar Return-to-Zero (**UP-RZ**).

Polar (represented by two distinct non-zero symmetrical but opposite voltage levels, +V and –V).[4]

Bipolar (also known as pseudo-ternary +V, –V, and 0 V or alternate mark inversion).

Polar line codes can be further classified as depicted in Figure 2.1.8.

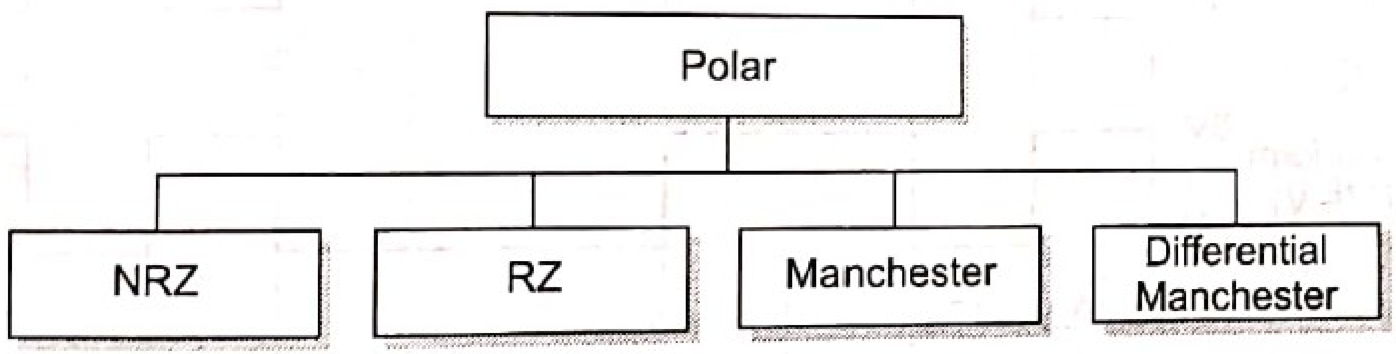


Figure 2.1.8 Classification of Polar Line Codes

Any one of several line coding techniques can be used for the electrical representation of a binary data stream:

- Unipolar Non-Return-to-Zero (NRZ) Line Code
- Unipolar Return-to-Zero (RZ) Line Code
- Polar Non-Return-to-Zero (NRZ) Line Code
- Manchester Polar Line Code
- Differential Manchester Polar Line Code
- Bipolar Non-Return-to-Zero Alternate Mark Inversion (BP-NRZ-AMI) Line Code
- Bipolar RZ Line Code
- Bipolar RZ-AMI Line Code
- High-Density Bipolar (HDB) NRZ-AMI Line Code
- Binary Eight Zeros Substitution (B8ZS) RZ-AMI Line Code

Now, we describe each one of these line codes briefly in order to develop better understanding

In *unipolar NRZ line code*, a binary symbol 0 is represented by 0 V for the entire bit interval T_b and the binary symbol 1 is represented by a constant voltage level, say $+V$ or $-V$, during its entire bit interval T_b (where bit rate is $1/T_b$), and, hence, called non-return-to-zero (NRZ).