

# **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 discrete 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.

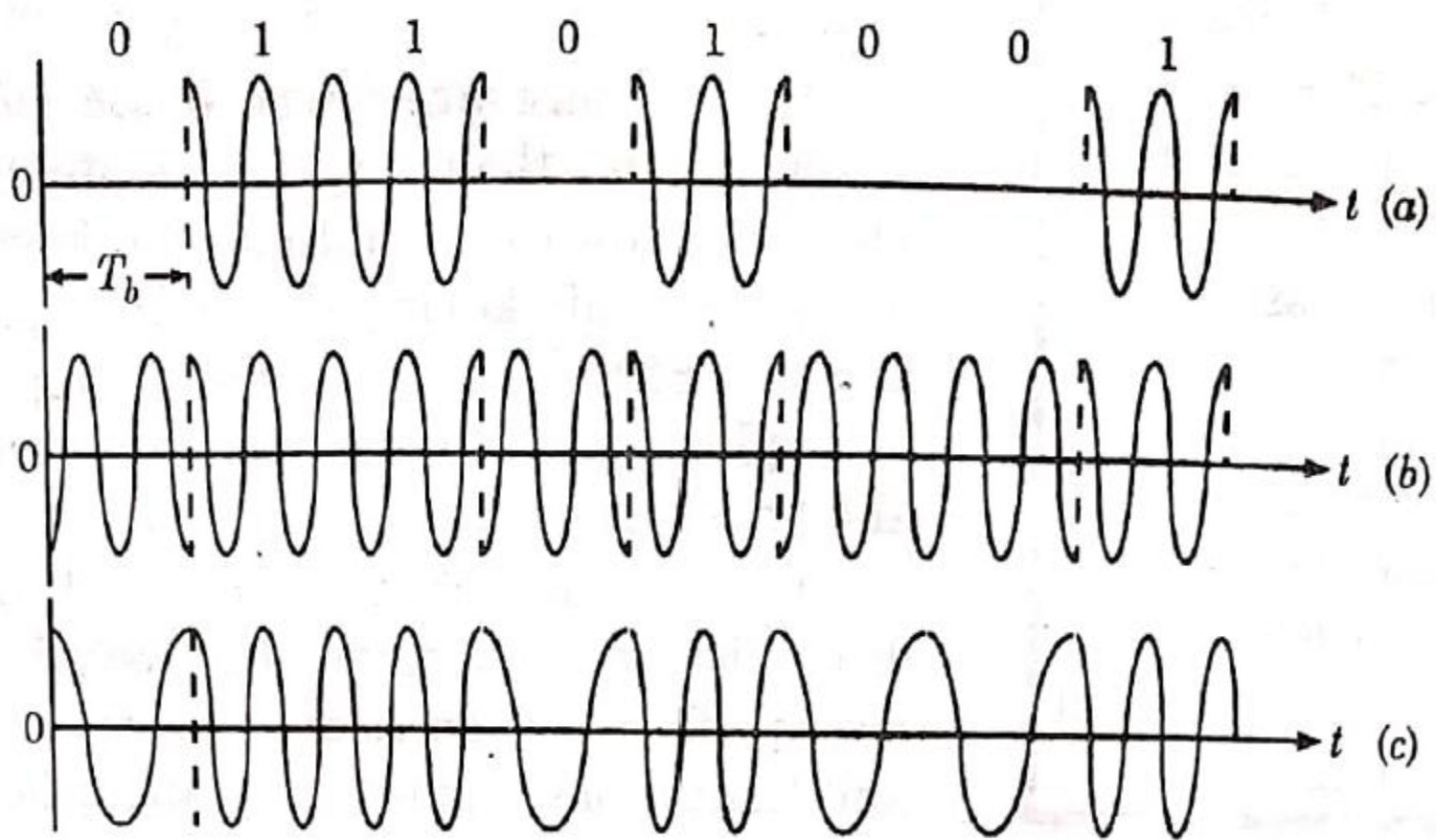
When we have to transmit a digital signal over a long distance, we need continuous-wave (CW) modulation. For this purpose, the transmission medium can be in form of radio, cable or other type of channel. Also, a carrier signal having some frequency  $f_c$  is used for modulation. Then the modulating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

Here, the word bandpass is used since the range of frequencies does not start from zero Hz to  $f_m$  Hz. In fact, the range of frequencies from zero Hz to  $f_m$  Hz is 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 to the designer of a digital communication system required for data transmission over a bandpass channel.

Every scheme offers system trade-offs of its own. However, the final choice made by the designer is determined by the way in which the available primary communication resources such as transmitted power and channel bandwidth are best exploited. In particular, the choice is made in favour of a scheme which possesses as many of the 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.

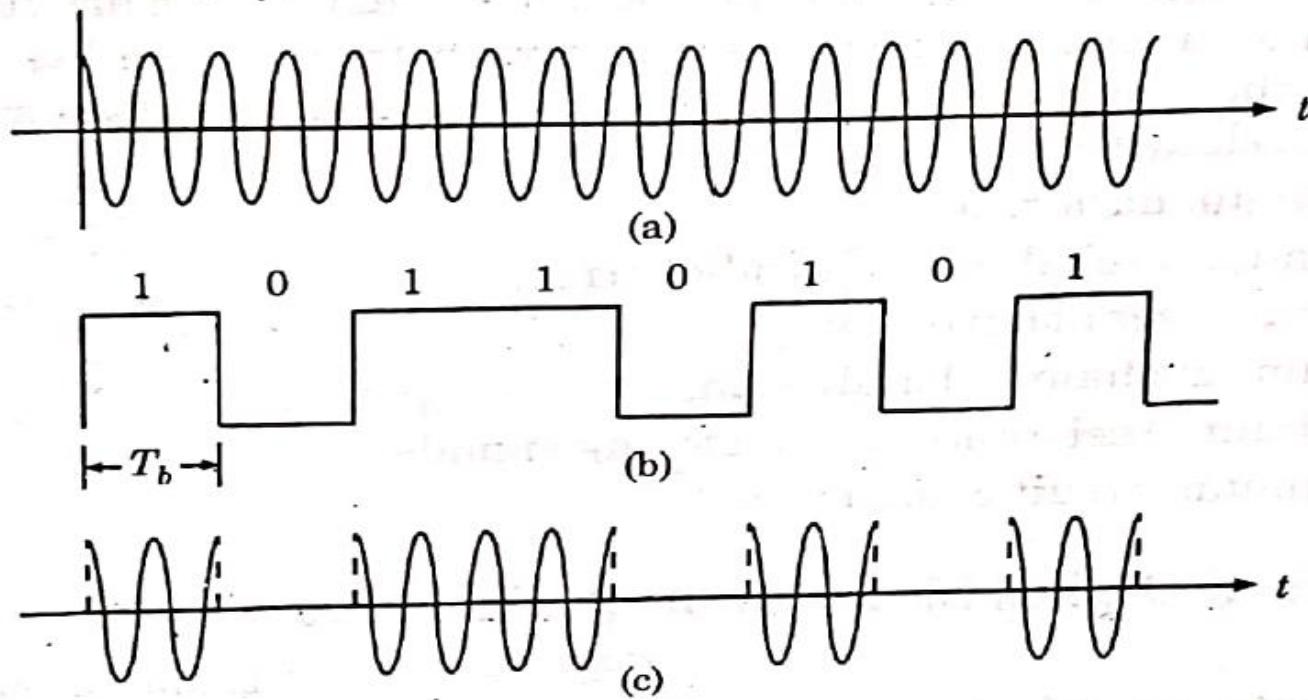
## **7.5. Coherent Binary Amplitude Shift Keying or On-Off Keying**

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

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

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



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

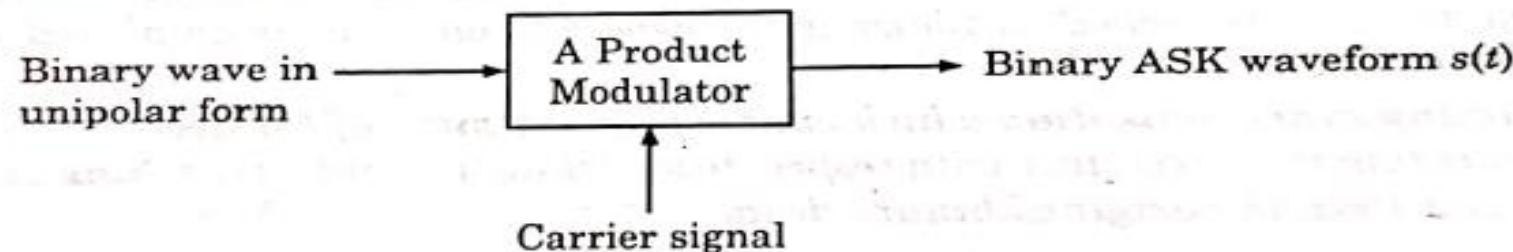
To transmit symbol '0', the signal  $s(t) = 0$  i.e., no signal is transmitted. Signal  $s(t)$  contains some complete cycles of carrier frequency ' $f_c$ '.

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.

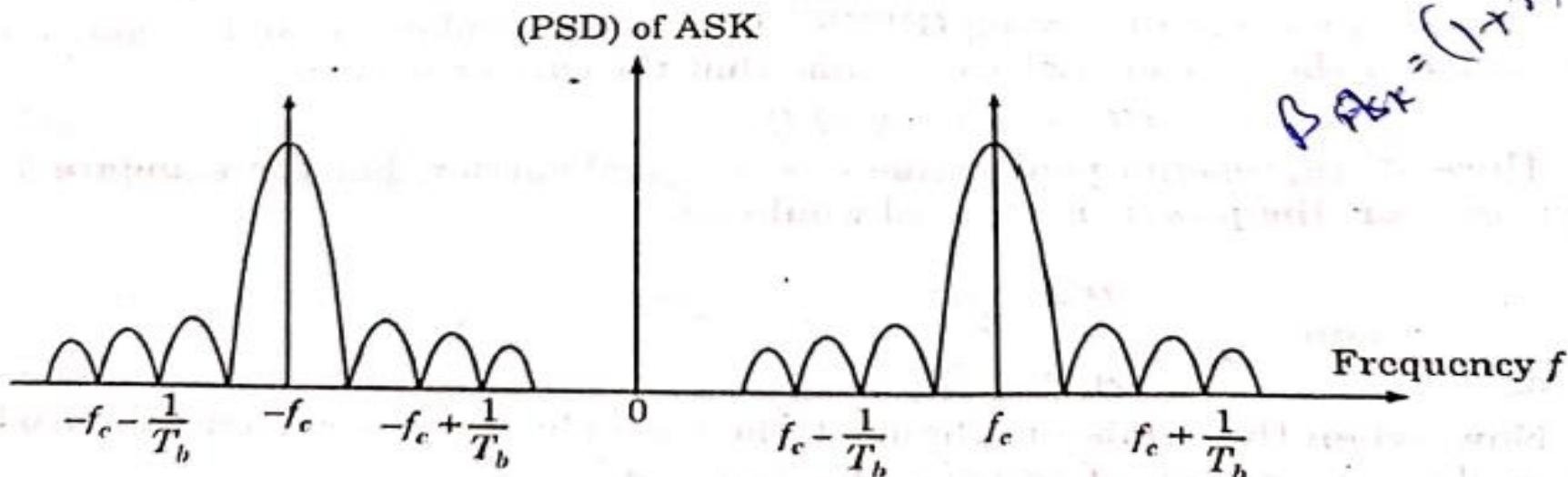
## 7.5.2. Generation of ASK Signal

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.



**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.

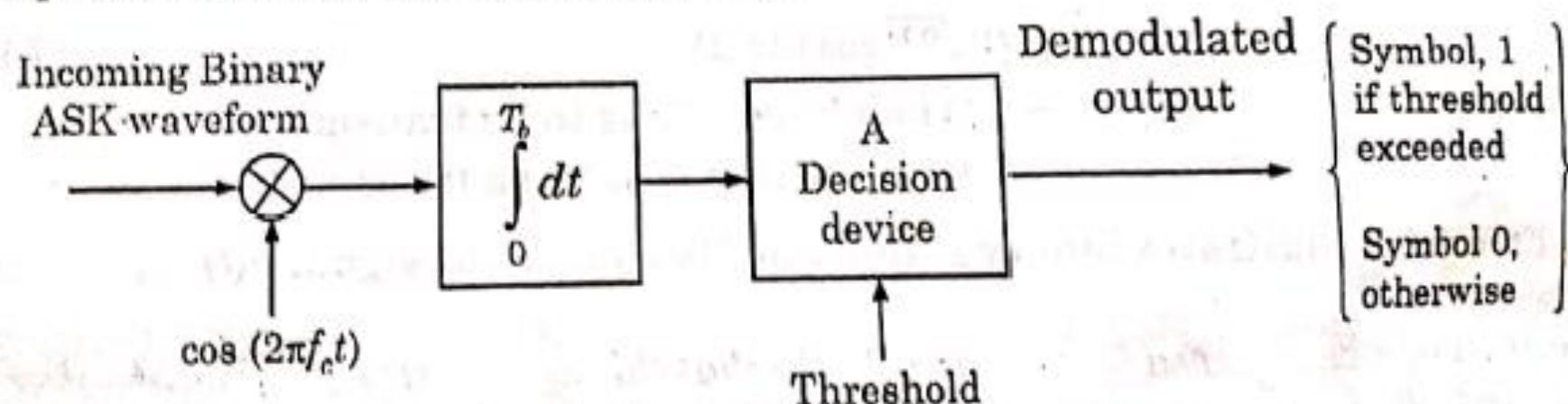


Fig. 7.6. *Coherent detection of binary ASK signals.*

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,

$$s(t) = A \cos(2\pi f_c t) \quad \dots(7.4)$$

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

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

or

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

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

Let us consider, for example,

For symbol '1', we have

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

If next symbol is '0', then we have

For symbol '0', we have

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

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

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

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

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

where

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

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

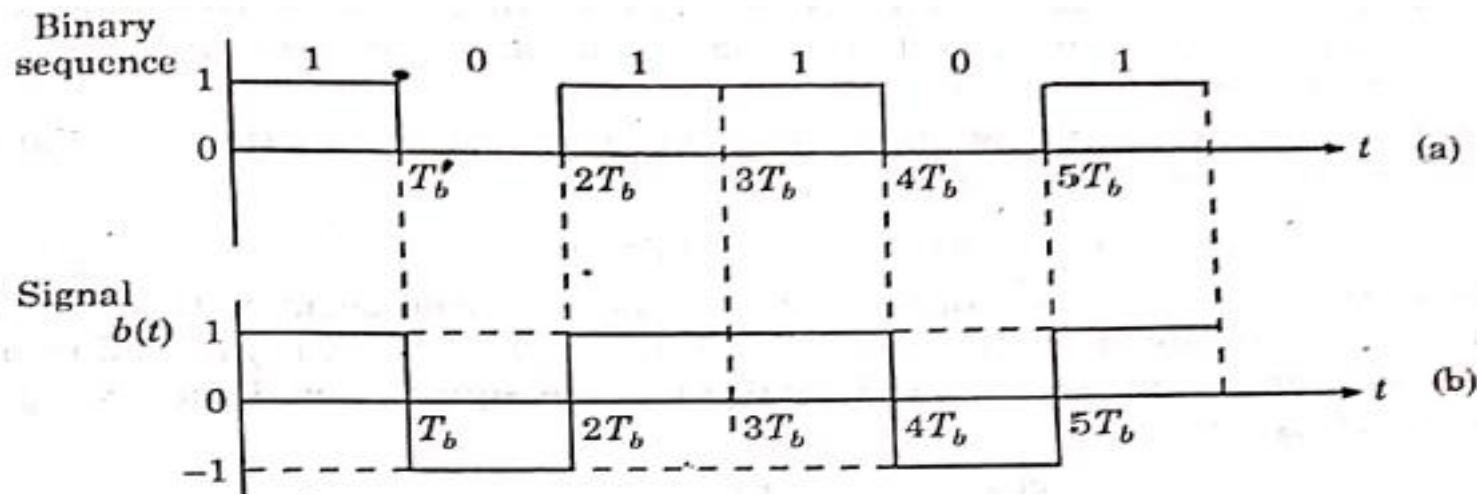


Fig. 7.7. (a) Binary sequence, (b) The corresponding bipolar signal  $b(t)$ .

### 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.

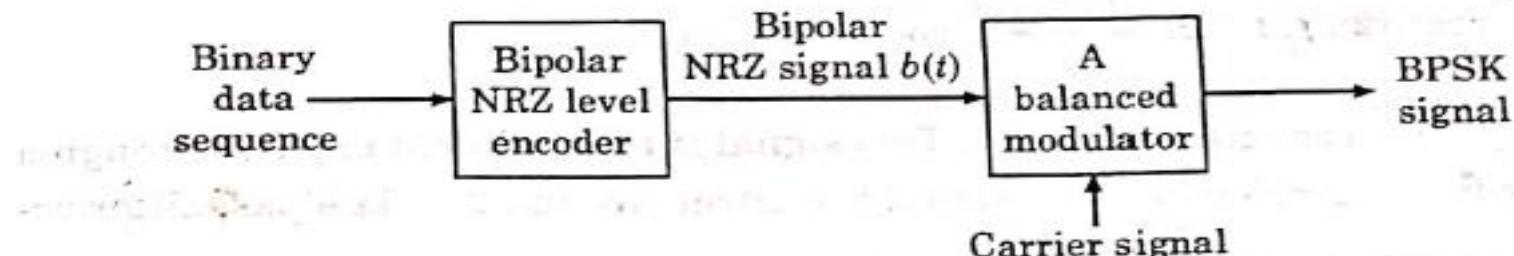


Fig. 7.8. Generation of BPSK.

### 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

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

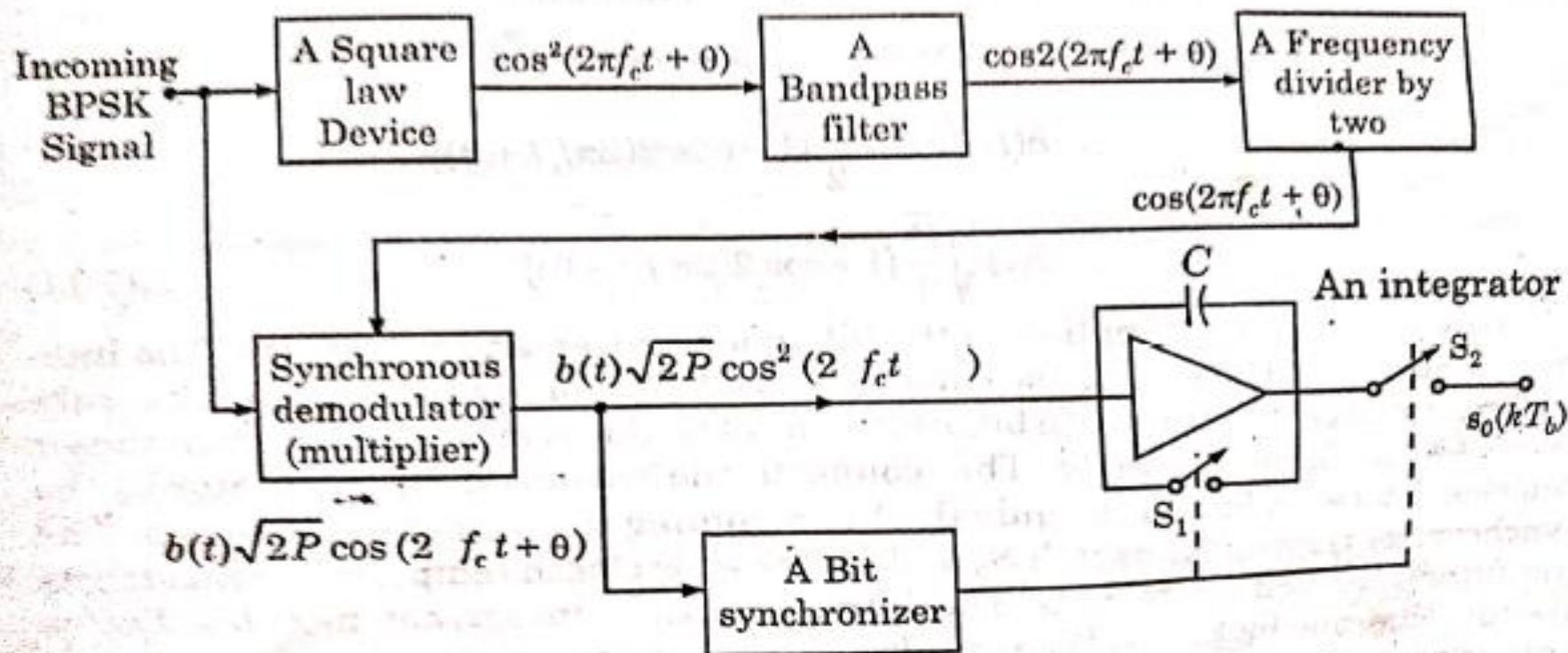


Fig. 7.9. Reception of baseband signal in BPSK signal.

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  $\theta$ . Because of this, the signal at the input of the receiver can be written as

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

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,

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

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

This signal is having frequency equal to  $2f_c$ . 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

$$\begin{aligned}
 b(t)\sqrt{2P} \cos(2\pi f_c t + \theta) \times \cos(2\pi f_c t + \theta) &= b(t)\sqrt{2P} \cos^2(2\pi f_c t + \theta) \\
 &= b(t)\sqrt{2P} \times \frac{1}{2} [1 + \cos 2(2\pi f_c t + \theta)] \\
 &= b(t) \sqrt{\frac{P}{2}} [1 + \cos 2(2\pi f_c t + \theta)] \quad \dots(7.11)
 \end{aligned}$$

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 temporarily. 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.

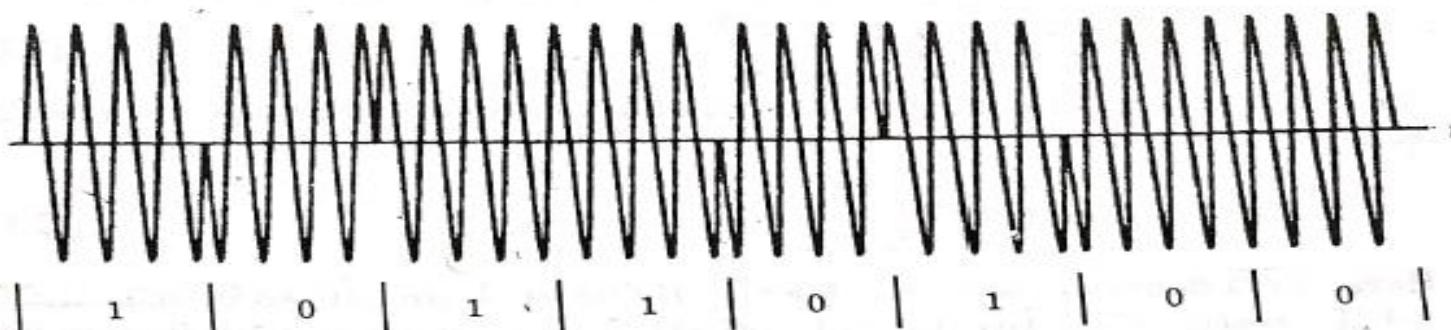


Fig. 7.10. The BPSK waveform.

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.

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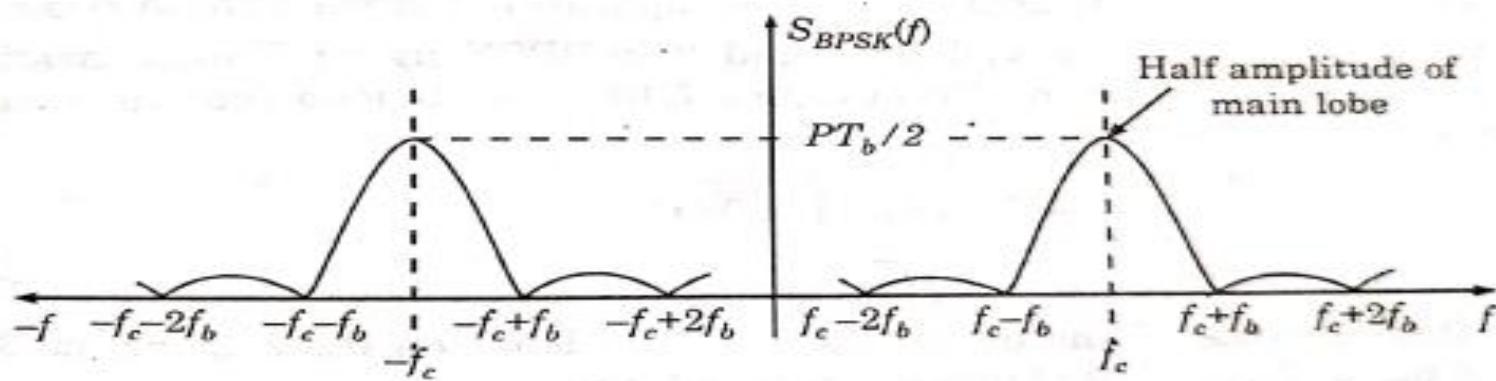


Fig. 7.12. Plot of power spectral density of BPSK signal.

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  $\Omega$ . Then we can write following equations.

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

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

Hence, there is increase or decrease in frequency by  $\Omega$ . 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

$$s(t) = \sqrt{2P_s} \cos[(2\pi f_c + d(t)\Omega)t] \quad \dots(7.27)$$

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

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

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

### 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.

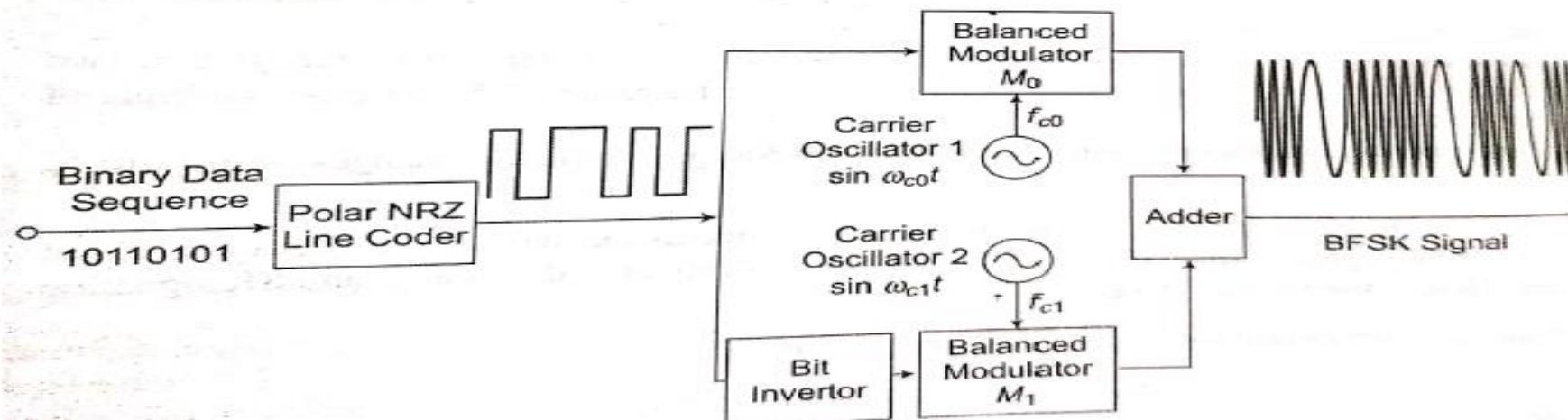


Figure 3.1.9 Functional Block Schematic of FSK Modulator

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  $\sqrt{P_s T_b}$ . Zero level is unaffected. Thus, the output

of the level shifters will be either  $\sqrt{P_s T_b}$  (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. \* \* \* \* \*  
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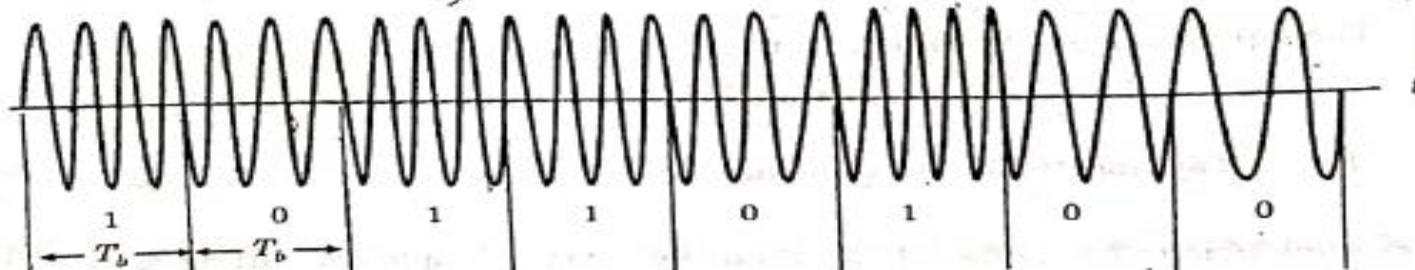


Fig. 7.15. The BFSK signal.

#### 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.

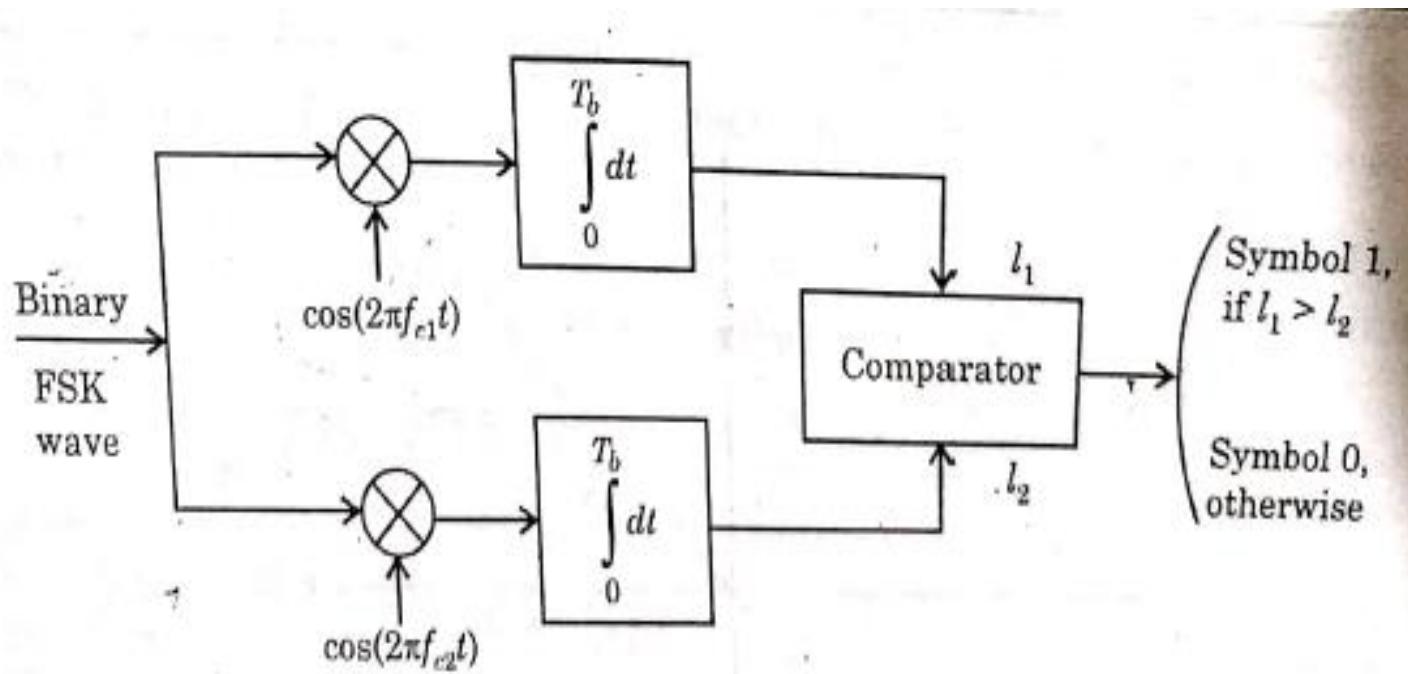


Fig. 7.17. Block diagram of BFSK receiver (detection of BFSK).

## 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.

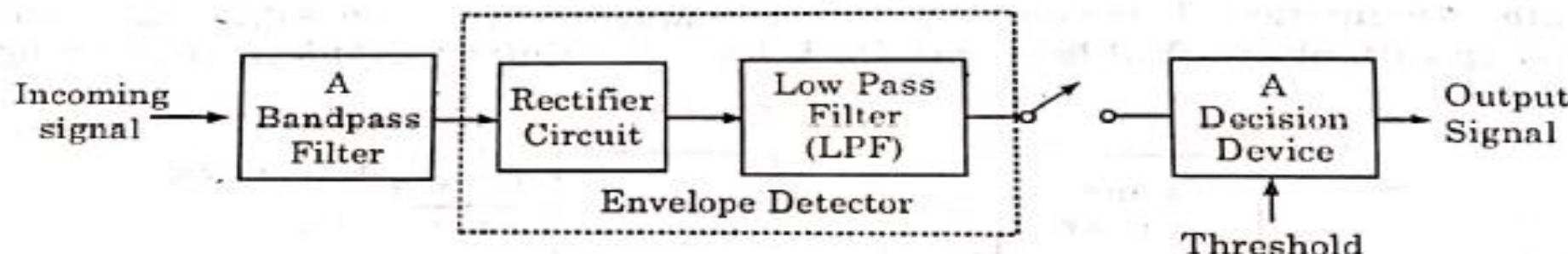


Fig. 7.20. Non-coherent ASK detector.

## 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.

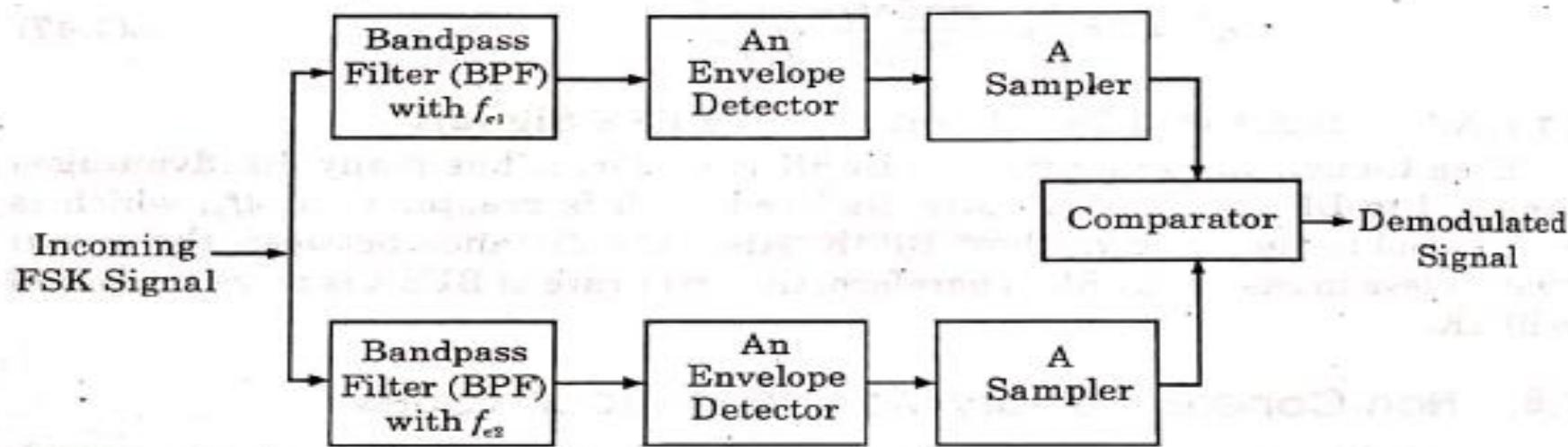


Fig. 7.21. Non-coherent detection of FSK binary signals.

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)*.

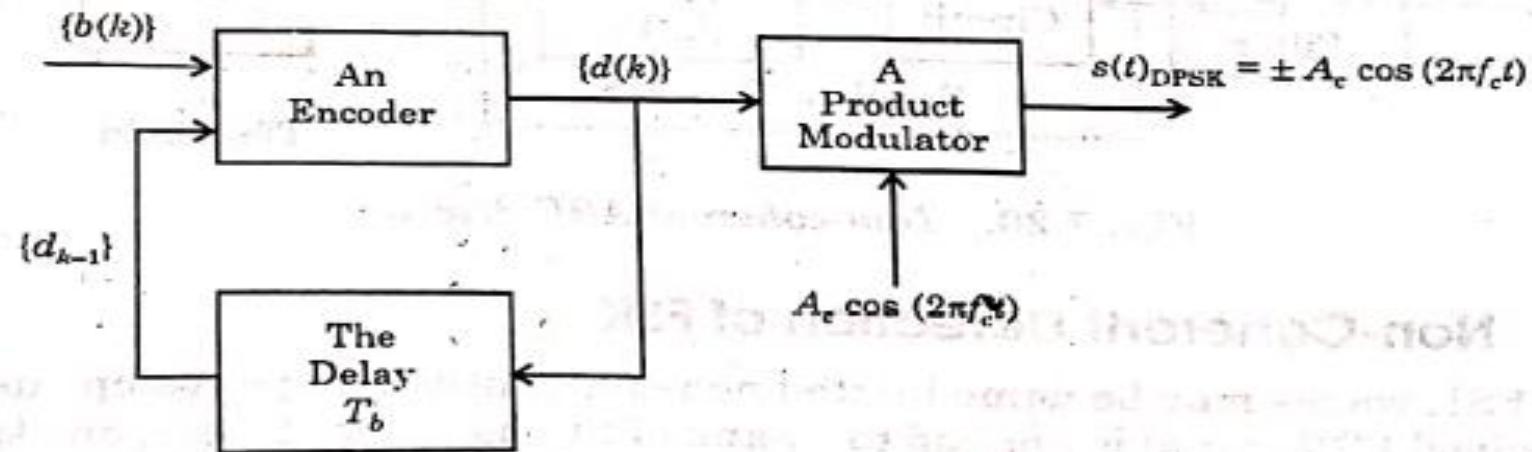


Fig. 7.22. Illustration of the scheme to generate DPSK signals.

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.

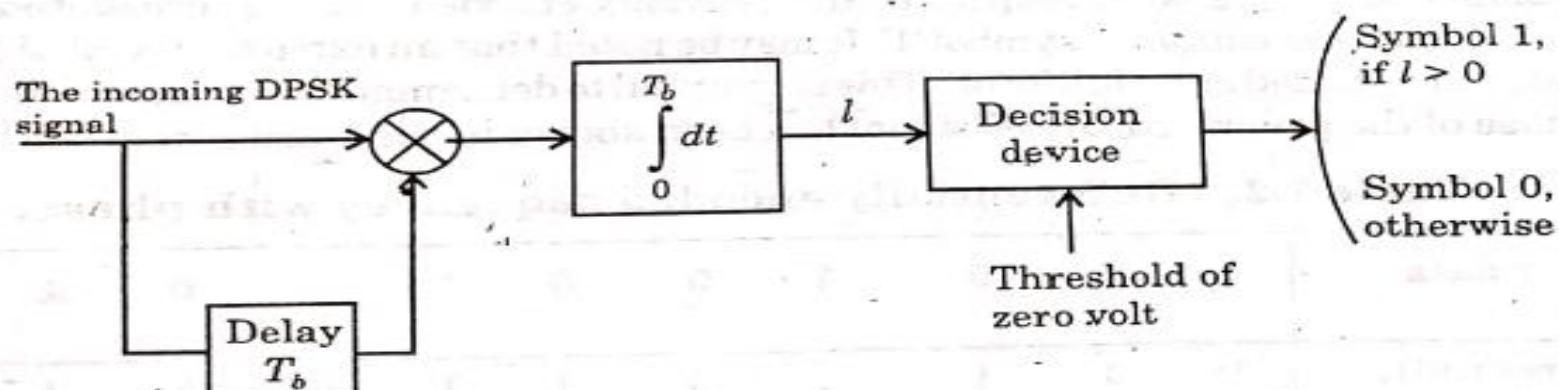
**Table 7.2.** Differentially encoded sequences with phase.

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 $\pi$ 0 0 $\pi$ 0 0 $\pi$ 0 0 0
Shifted differentially encoded data $\{d_{k-1}\}$	1 0 1 1 0 1 1 0 1 0 1 1
Phase of shifted DPSK	0 $\pi$ 0 0 $\pi$ 0 0 $\pi$ 0 0 0
Phase comparison output	- - + - - + - - + +
Detected binary sequence	0 0 1 0 0 1 0 0 1 1

### 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  $\cos(\phi)$ , here  $\phi$  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 binary 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  $180^\circ$ .

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  $\pi/4$ -*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  $90^\circ$  or  $180^\circ$  with another. Thus, in a QPSK modulator, each *dabit* code generates one of the four possible output phases ( $+45^\circ$ ,  $+135^\circ$ ,  $-45^\circ$ ,  $-135^\circ$ ), hence, the name “quadrature” meaning “four” phases.

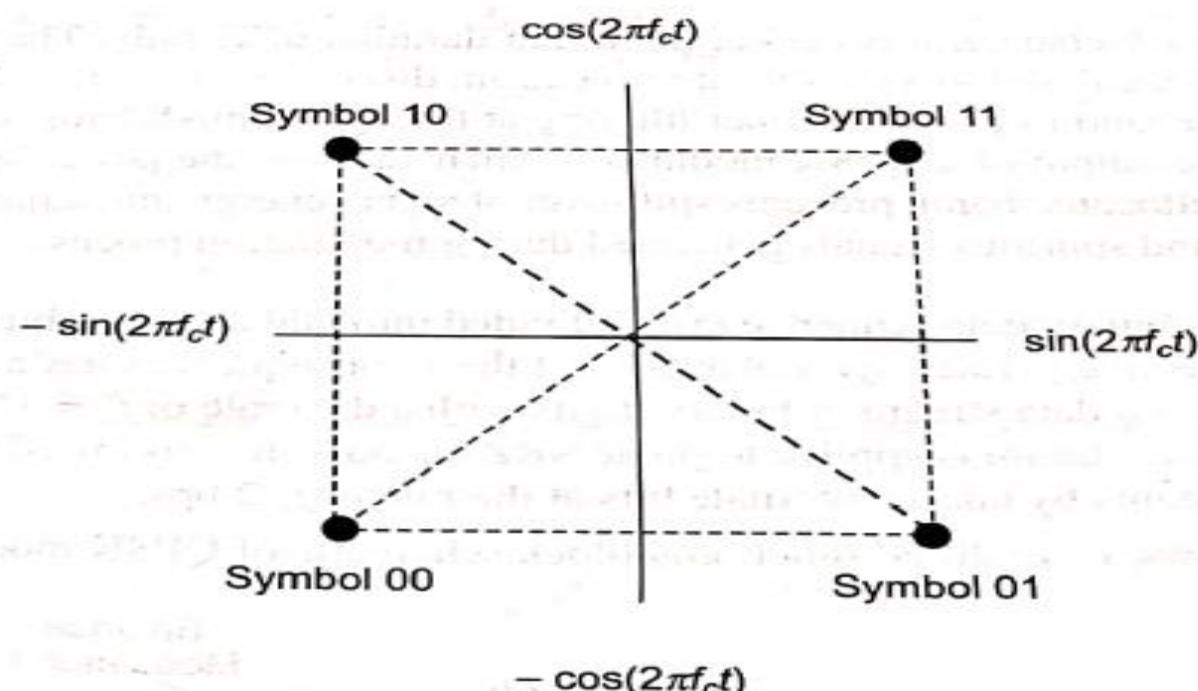
Mathematically, the QPSK signal for one symbol duration, consisting of two bits each, can be expressed as

$$S_{\text{QPSK}}(t) = \begin{cases} \sin(2\pi f_c t - 3\pi/4) & \text{for } 00 \\ \sin(2\pi f_c t - \pi/4) & \text{for } 01 \\ \sin(2\pi f_c t + 3\pi/4) & \text{for } 10 \\ \sin(2\pi f_c t + \pi/4) & \text{for } 11 \end{cases}$$

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	$-135^\circ$ or $-3\pi/4$ radians
$s_2$	0 1	$-45^\circ$ or $-\pi/4$ radians
$s_3$	1 0	$+135^\circ$ or $3\pi/4$ radians
$s_4$	1 1	$+45^\circ$ 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  $180^\circ$  transition (from 11 to 00 or from 10 to 01 or vice versa). Whereas it makes only  $90^\circ$  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.<sup>12</sup>

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.

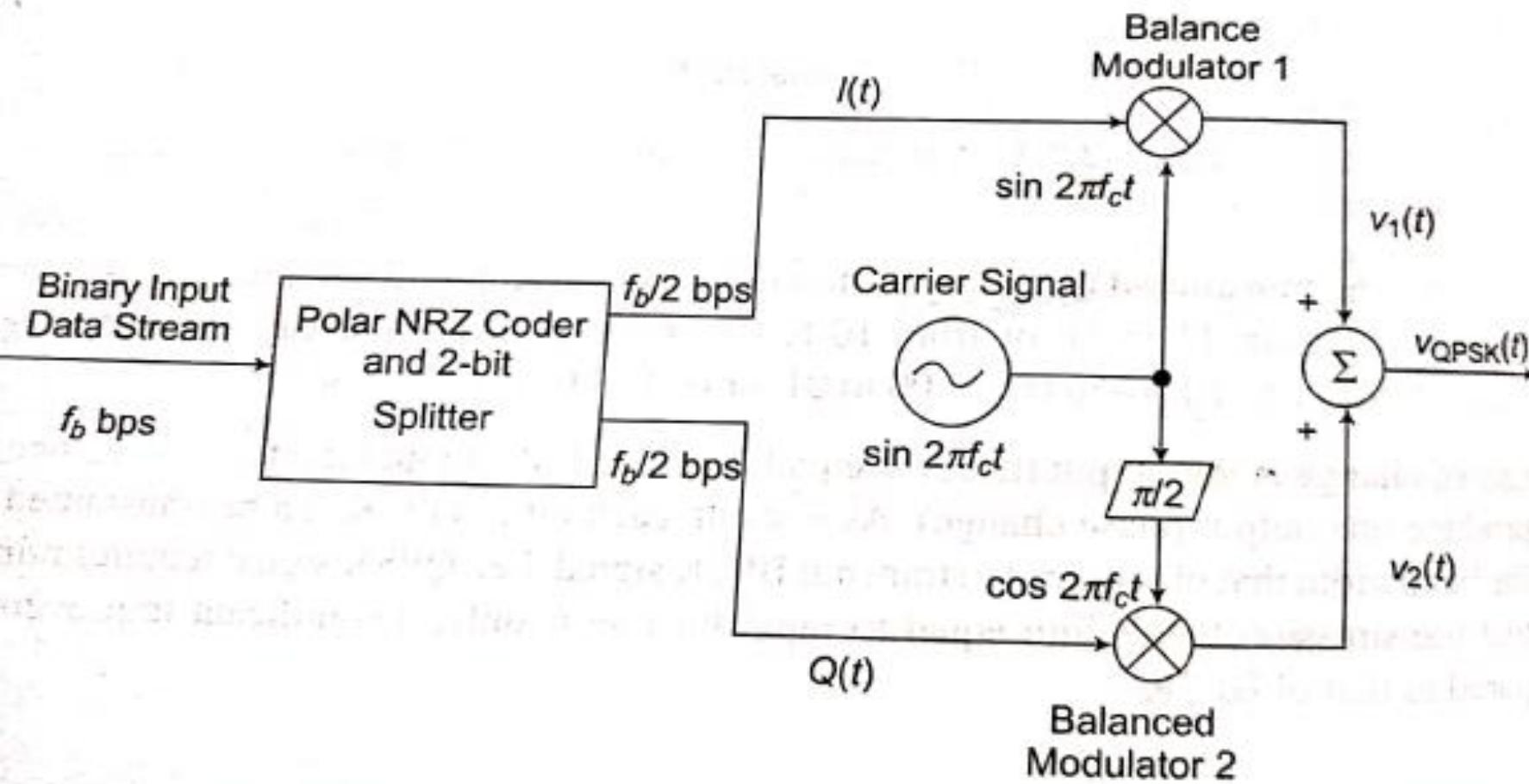


Figure 3.2.3 Functional Block Schematic of QPSK Modulator

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

$$\begin{aligned} & +\cos(2\pi f_c t) + \sin(2\pi f_c t), -\cos(2\pi f_c t) + \sin(2\pi f_c t), \\ & +\cos(2\pi f_c t) - \sin(2\pi f_c t), \text{ and } -\cos(2\pi f_c t) - \sin(2\pi f_c t). \end{aligned}$$

- 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:

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

Note... When  $d_1(t)$  and  $d_2(t)$  signals modulate the sinusoidal carrier signals of frequency  $f_c$ , the QPSK is shift keying.

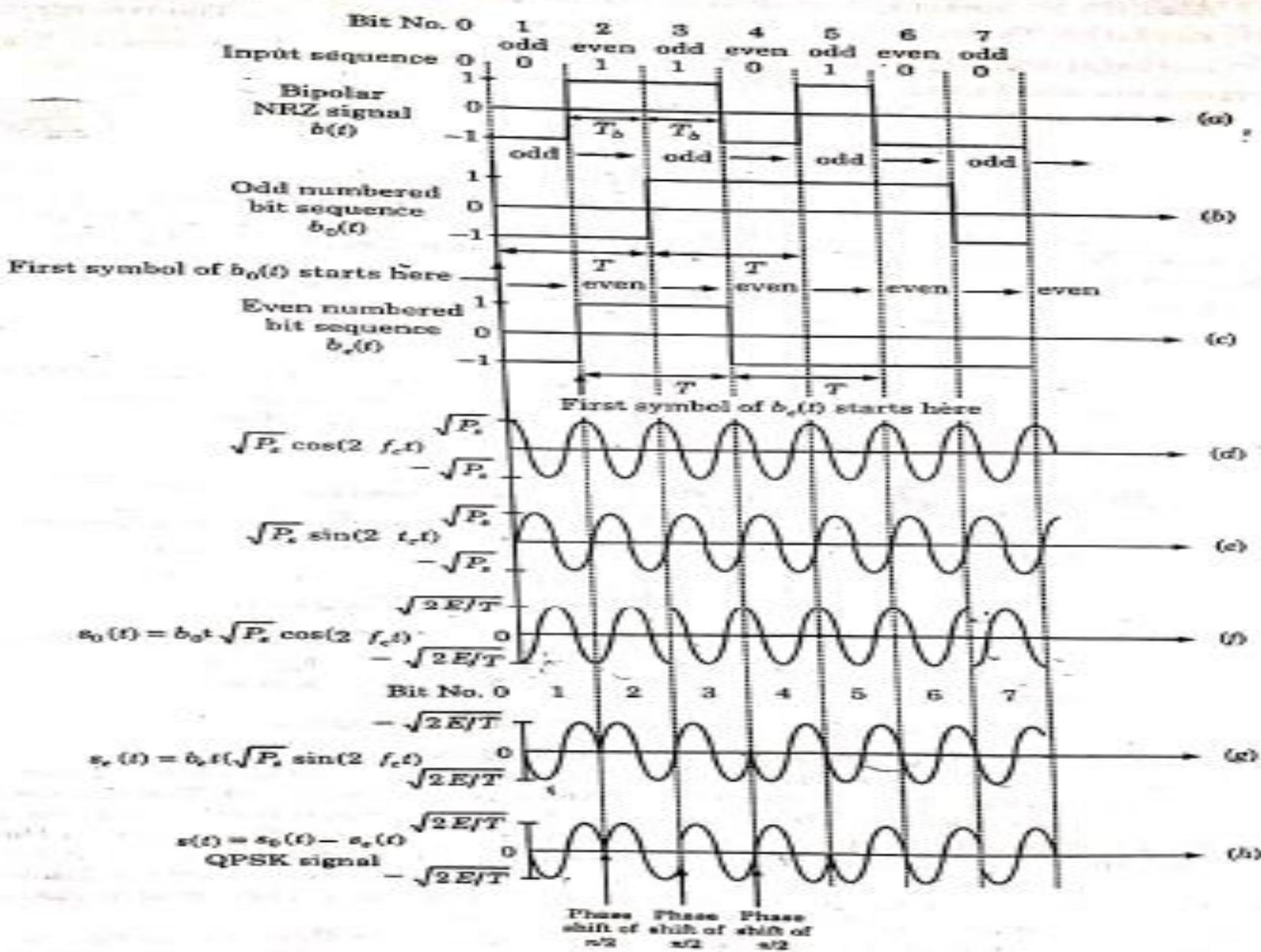


Fig. 7.25. QPSK waveforms. (a) Input sequence and its corresponding NRZ waveform. (b) Odd numbered bit sequence and its corresponding waveform. (c) Even numbered bit sequence and its NRZ waveform. (d) Basis function  $\phi_0(t)$ . (e) Basis function  $\phi_1(t)$ .

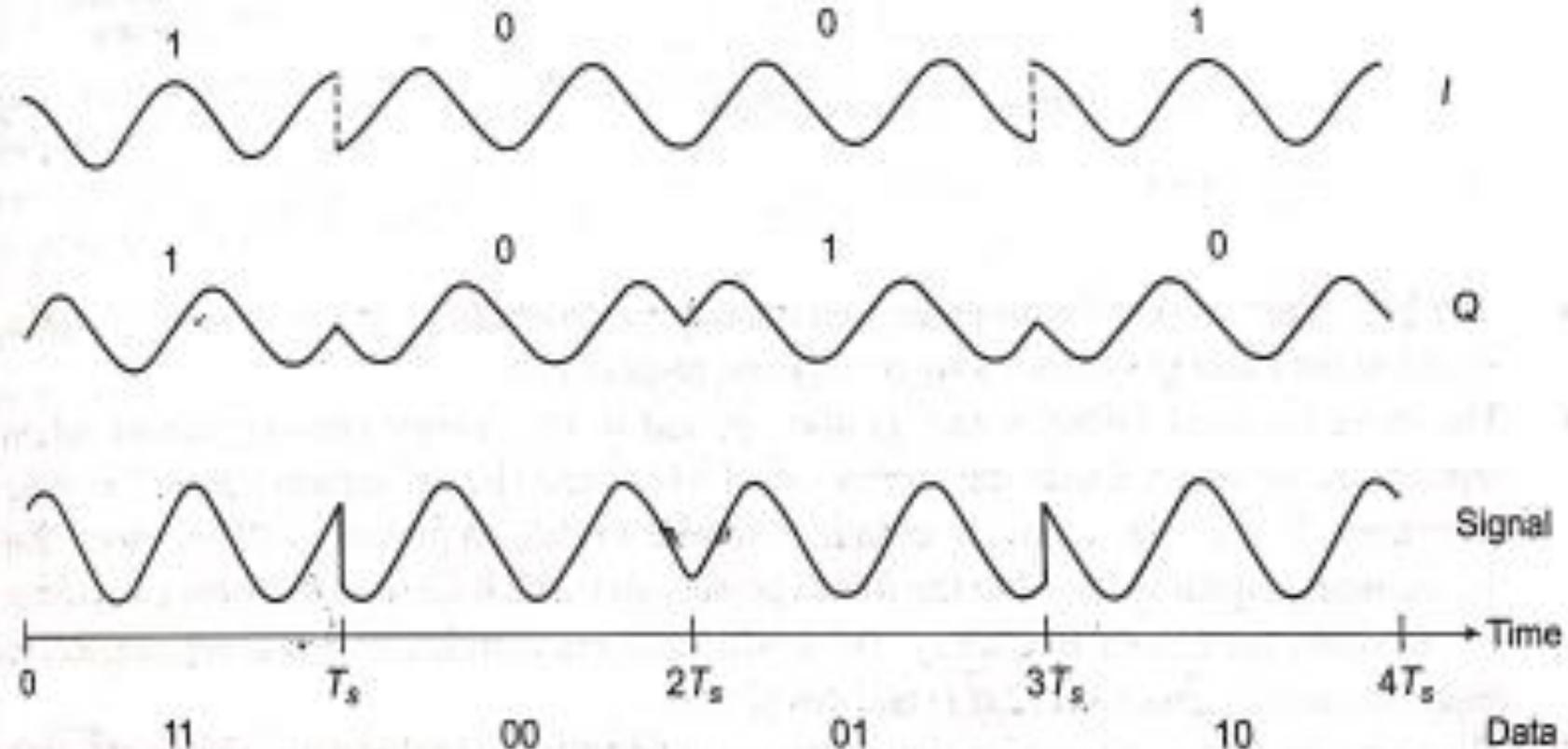


Figure 3.2.4 QPSK Signal in Time Domain ( $T_s$ : Symbol duration)

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.<sup>13</sup>

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

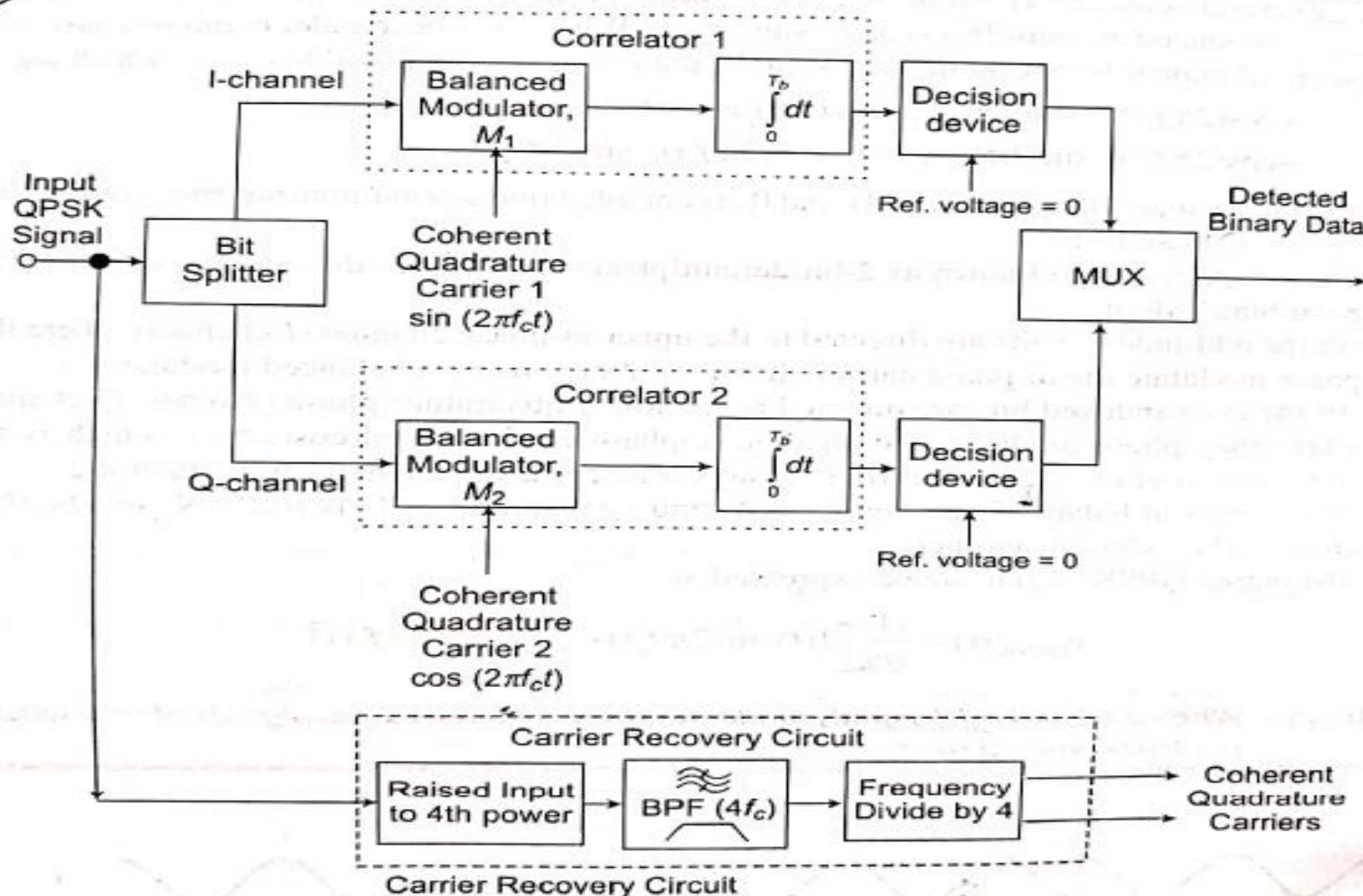


Figure 3.2.5 Functional Block Schematic of Coherent QPSK Detector

- 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  $T_s = 2T_b$ .

<sup>13</sup> 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,  $f_b = 1 \text{ Mbps}$  or  $1 \times 10^6 \text{ bps}$

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 =  $\frac{1}{T_s} = \frac{1}{2 \mu\text{s}} = 500 \text{ k symbols/second}$

**Ans.**

Minimum bandwidth,  $B_{QPSK} = f_b = 1 \text{ MHz}$

**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  $\times (1 + \text{roll-off factor})$

Given bit rate of a bipolar NRZ data sequence,  $f_b = 9600 \text{ bps}$

In QPSK, symbol rate =  $\frac{1}{2} \times \text{bit rate} = \frac{1}{2} \times 9600 \text{ bps} = 4800 \text{ symbols per second}$

(a) For given roll-off factor of 0.35, required bandwidth =  $4800 \times (1 + 0.35)$

Hence, required bandwidth = 6480 Hz

**Ans.**

(b) For given roll-off factor of 0.5, required bandwidth =  $4800 \times (1 + 0.5)$

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  $\pm 90^\circ$ , or sometimes even by  $\pm 180^\circ$ , depending upon whether the sign change of the in-phase and quadrature component occurs simultaneously or not. From the signal constellation diagram of QPSK, it can be seen 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 change of carrier phase by  $\pm 90^\circ$  only.
- On the other hand, whenever adjacent dibits in the binary data sequence differ in both the binary digits (e.g., 1 1 and 0 0, or 0 1 and 1 0), the transition involves a change of carrier phase by  $\pm 180^\circ$ .

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  $\pm 180^\circ$  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  $\pm 90^\circ$  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

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

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 dabit code. There is never more than a  $90^\circ$  phase shift in the output phase. In other words, the modulated OQPSK signal transitions are  $\pm 90$  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.

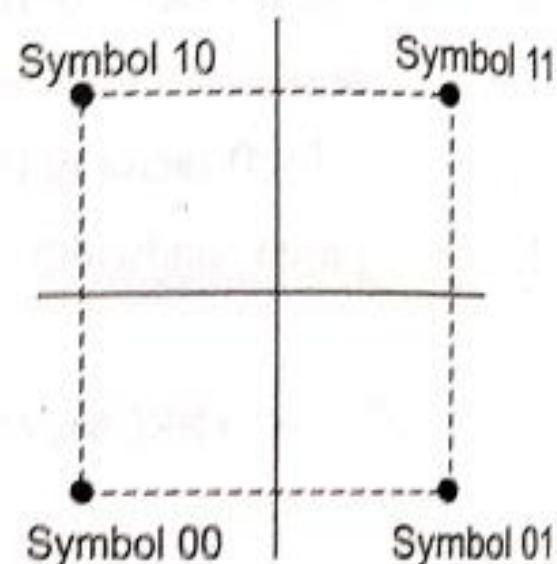


Figure 3.2.6 Constellation Diagram of OQPSK Signal

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.

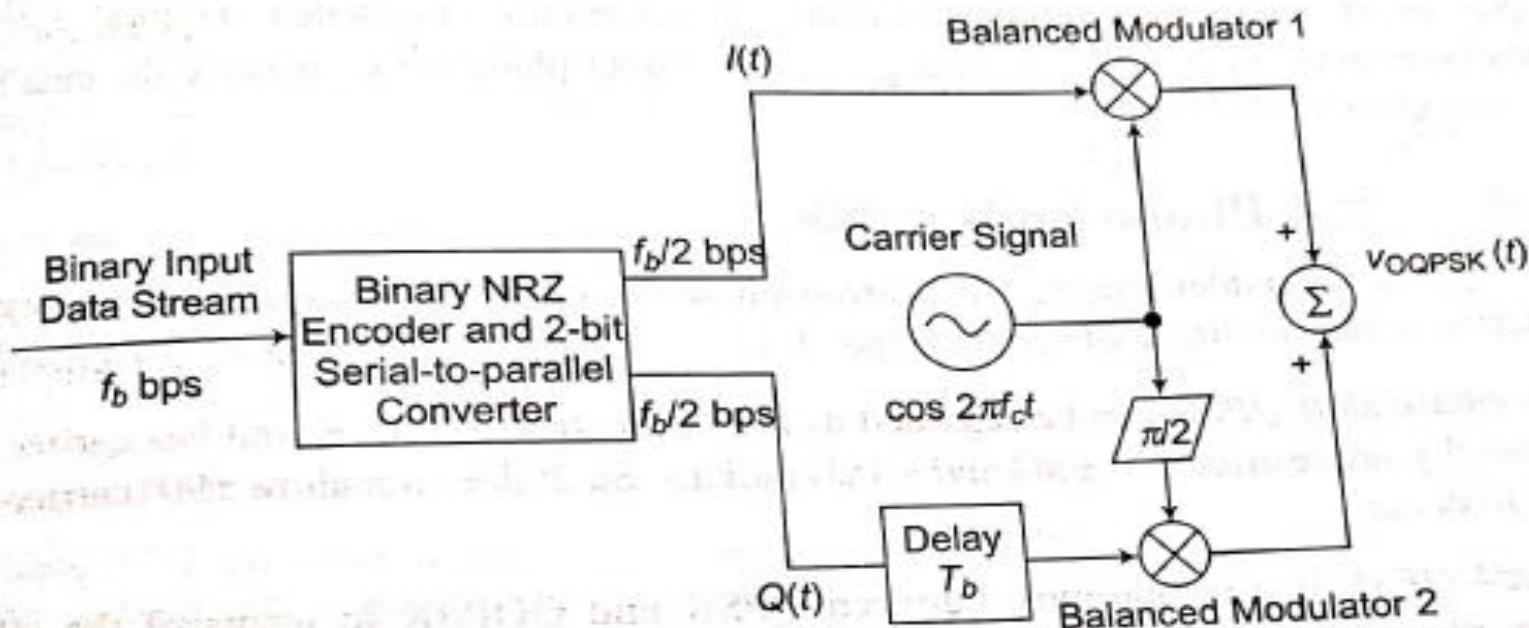
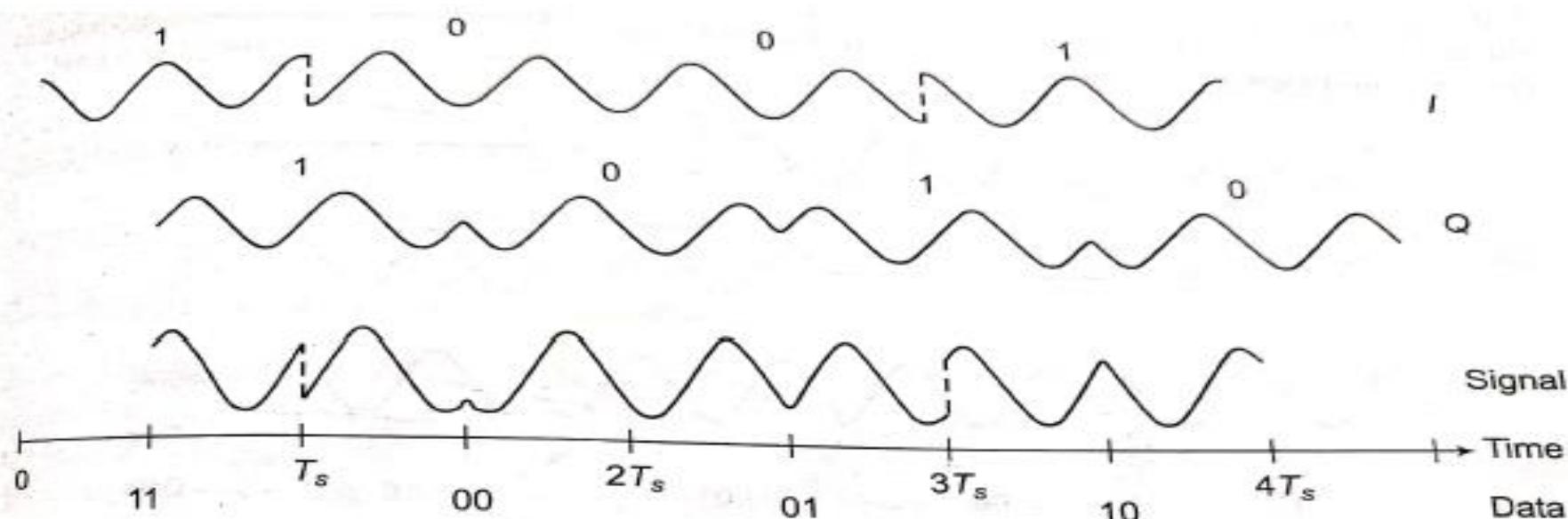


Figure 3.2.7 Functional Block Schematic of OQPSK Modulator

- 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 ( $T_s$ : 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  $\pm 90^\circ$  only.

The modulated OQPSK signal has no  $180^\circ$  phase shift. Instead, it has  $\pm 90^\circ$  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.<sup>14</sup>

### 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

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

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  $t = 0$ .

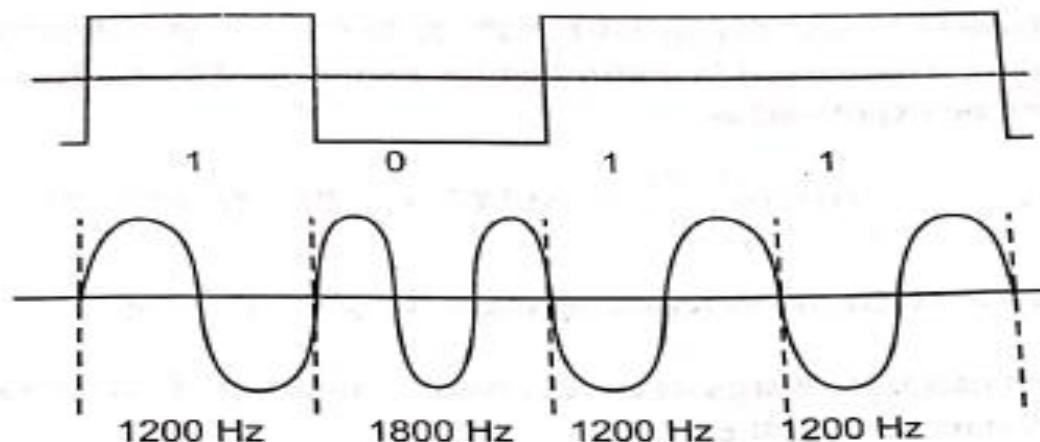


Figure 3.2.12 MSK Signal at 1200 baud for NRZ data 1 0 1 1

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  $\Delta f$  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 QPSK. It means maximum amplitudes of signals are scaled with respect to '1'.

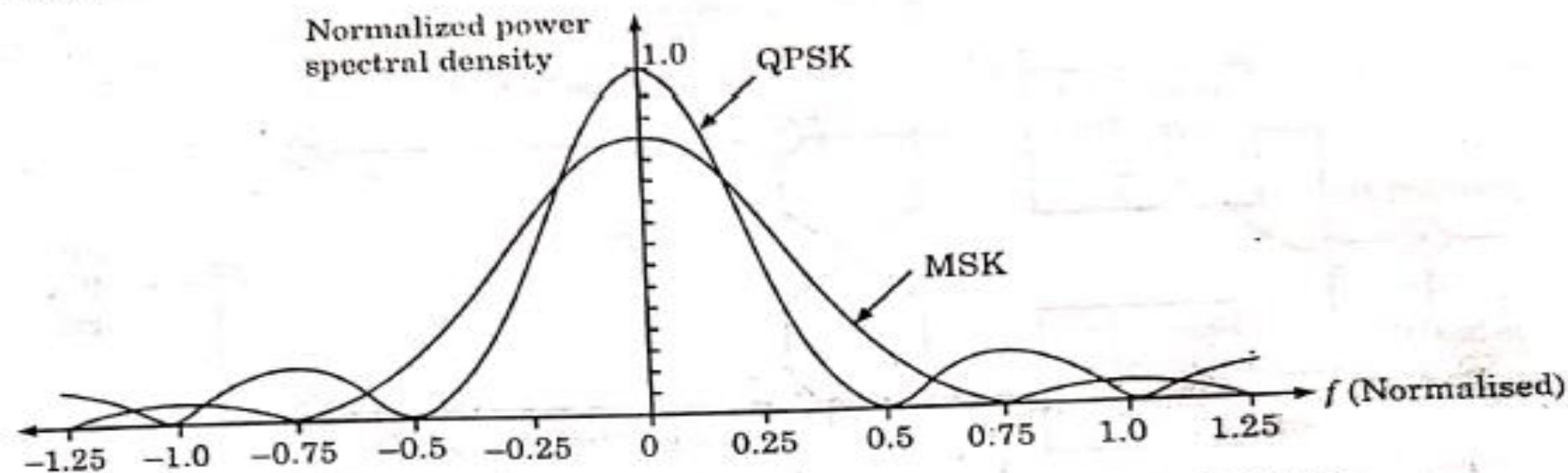


Fig. 7.33. Power spectral densities (psd) of MSK and QPSK.

The above plots show that the main lobe in MSK is wider than QPSK. The side 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  $\pm 0.75$  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.,

$$BW = 0.75 f_b - (-0.75 f_b) = 1.5 f_b \quad \dots(7.100)$$

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 multipliers 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).

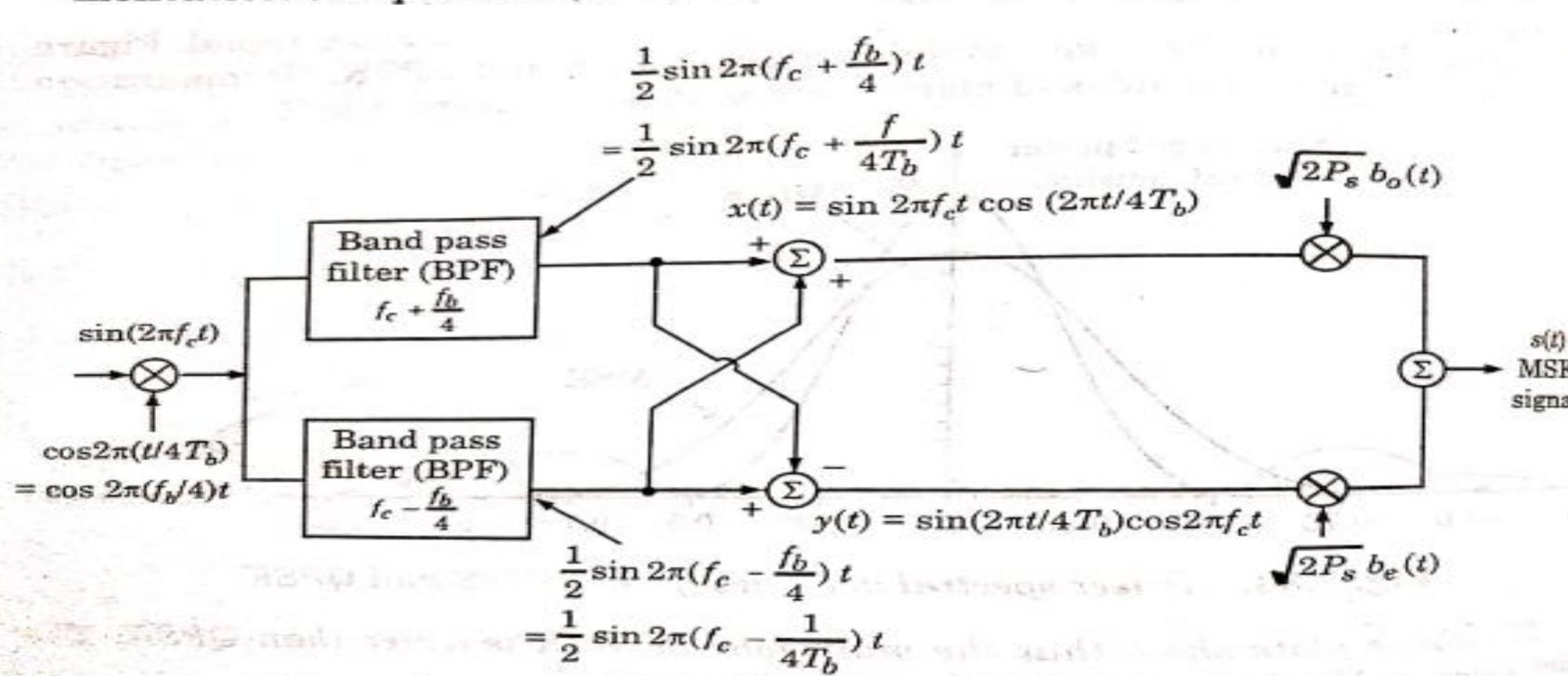


Fig. 7.34. MSK transmitter block diagram.

#### 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  $b_0(t)$  and  $b_e(t)$ . 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.

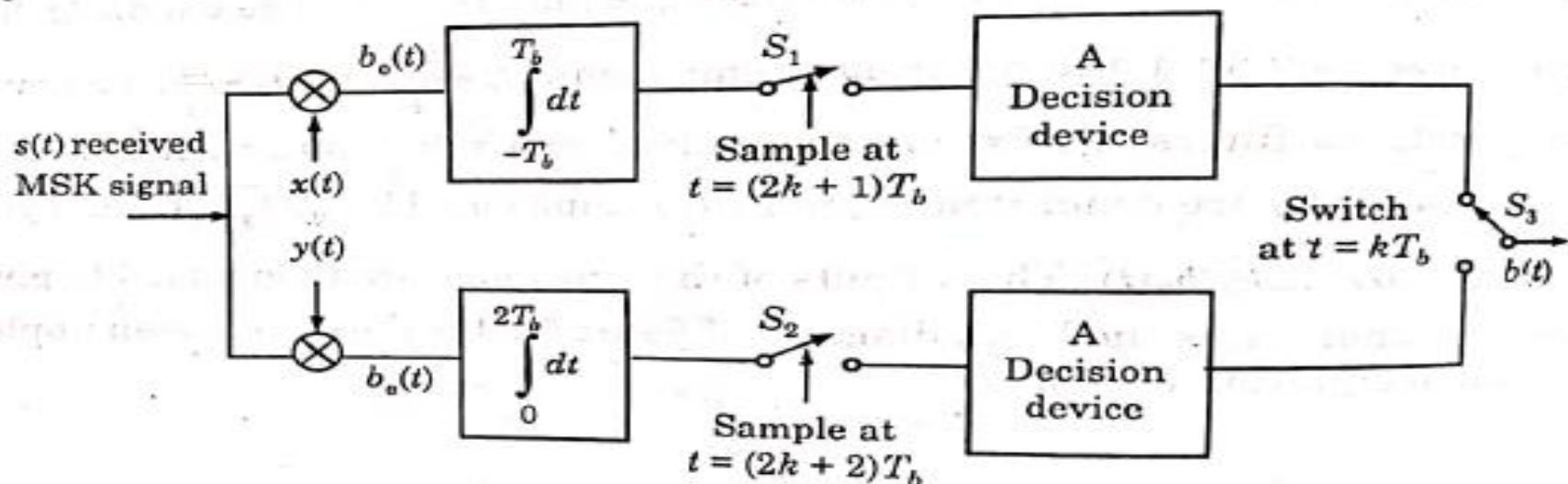


Fig. 7.35. MSK receiver block diagram.

### 3.2.5 Gaussian Minimum Shift Keying (GMSK)

MSK is a continuous phase modulation scheme where the modulated carrier contains no phase discontinuities and frequency changes occur at the carrier zero crossings. MSK is unique due to the relationship between the frequencies of a logical 0 and 1: the difference between the frequency of a logical 0 and a logical 1 is always equal to half the data rate. In other words, the modulation index is 0.5 for MSK.

The fundamental problem with MSK is that the spectrum is not compact enough to realize data rates approaching the RF channel bandwidth. A plot of the spectrum for MSK reveals side lobes extending well above the data rate. For wireless data transmission systems which require more efficient use of the RF channel bandwidth, it is necessary to reduce the energy of the MSK upper side lobes. As we have seen that the power spectral density of MSK falls off as the fourth power of the frequency. But it is still not adequate for multi-user wireless communication applications, which require very stringent specifications with regard to adjacent channel interference.

- So the PSD of an MSK signal is modified by passing the NRZ binary data stream through a pulse-shaping low-pass filter such as Gaussian filter before carrier modulation.
- The *Gaussian filter* has a narrow bandwidth with sharp cut-off characteristics and negligible overshoot in its impulse response.

In addition, this filter should permit the carrier-modulated signal in that it has a carrier phase of either 0 or  $\pi$  radians at even multiples of  $T_b$ , and either  $+\pi/2$  or  $-\pi/2$  radians at odd multiples of  $T_b$ .

Therefore, the modified MSK using such a Gaussian filter, is called ‘Gaussian-filtered MSK, or simply ‘*Gaussian MSK (GMSK)*’.

*Gaussian Minimum Shift Keying (GMSK)* is a special case of MSK in which a pre-modulation low-pass Gaussian filter is used as a pulse-shaping filter to reduce the bandwidth of the baseband signal before it is applied to the MSK modulator.

- If  $f_{3dB} < 1/T_b$ , then the intersymbol interference occurs since each data pulse overlaps with the data pulse in the next position in the symbol duration.

We know that the product of bandwidth and bit duration ( $BT_b$ ) plays an important role in the shape of the response of the pulse-shaping filter.

Since the pulse-shaping Gaussian filter acts on the input binary data which has been encoded in the form of polar NRZ waveforms, and is further used for frequency-shift keying, the shape of the filter response varies for different values of ( $BT_b$ ) product.

The pre-modulation Gaussian filtering introduces marginal intersymbol interference in the finally modulated signal. However, the degradation in performance can be minimized by selecting the appropriate 3 dB bandwidth-bit duration product ( $BT_b$ ) of the filter.

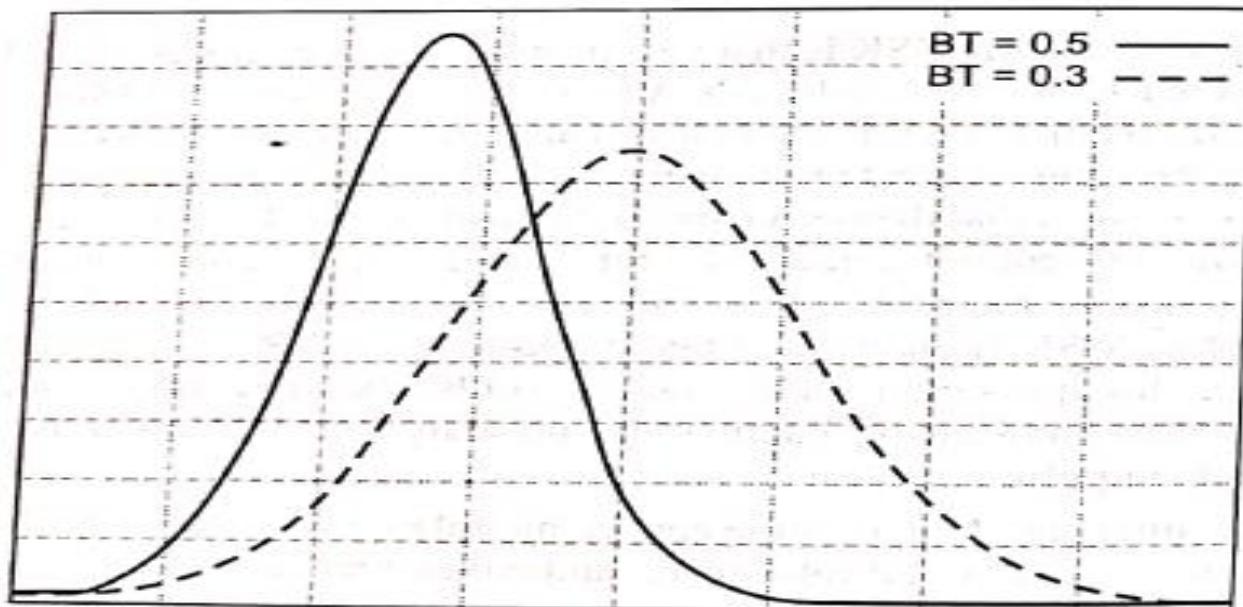


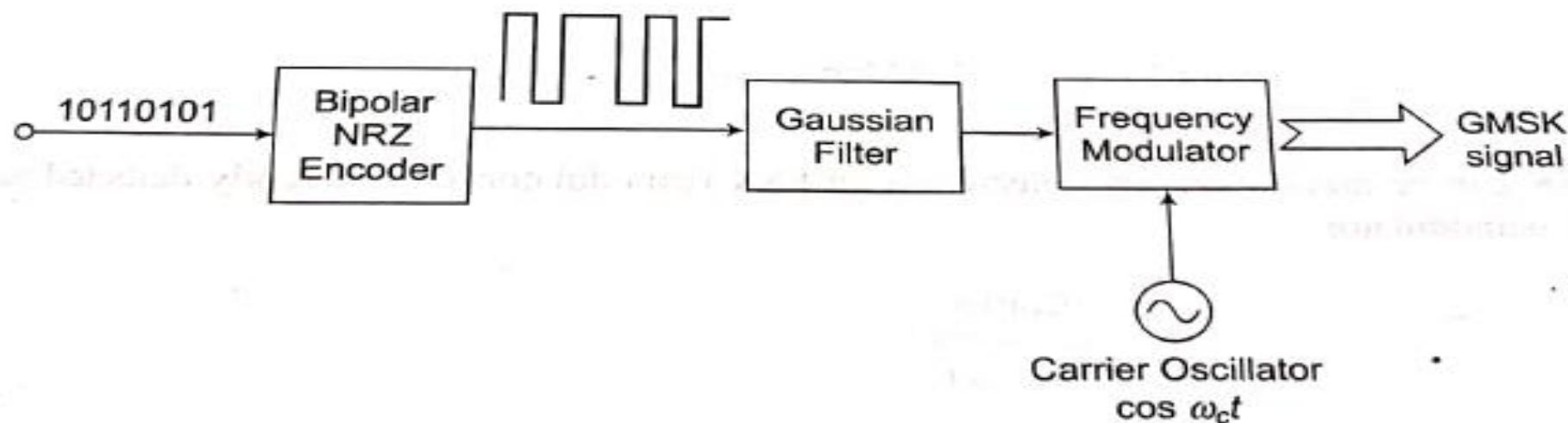
Figure 3.2.14 Gaussian Filter Impulse Response for  $B \times T_b$  (or, BT) = 0.3 and 0.5

For wireless data transmission systems, which require more efficient use of the RF channel, it is necessary to reduce the energy of the side lobes. We have discussed that the spectrum of MSK has side lobes which extend well above the data rate. The use of pre-modulation low-pass Gaussian filter having a narrow bandwidth with a sharp cut-off frequency, the side lobe energy is reduced. The resultant modulation scheme is known as Gaussian MSK.

The PSD of MSK signal has low side lobes because of the effect of the linear phase change. This enables to control adjacent channel interference. The power spectrum of GMSK with a  $(BT_b)$  value of infinity is equivalent to that of MSK. As the  $(BT_b)$  product decreases, the side lobe levels fall off very rapidly. For example, for a  $(BT_b) = 0.5$ , the peak of the second power value, increases the irreducible error rate generated by the Gaussian filter due to ISI. For  $(BT_b) = 0.3$ , adjacent bits or symbols will interfere with each other more than for  $(BT_b) = 0.5$ . In other words, MSK does not intentionally introduce ISI. But greater ISI allows the spectrum to be more compact, making demodulation more difficult. Hence, spectral compactness is the primary trade-off while selecting Gaussian pre-modulation filtered MSK. It implies that channel spacing can be closer for GMSK as compared to that for MSK for the same adjacent channel interference.

The channel spacing can be tighter for GMSK when compared to MSK for same adjacent channel interference.

Considering GMSK as a special case of FSK, it can be generated simply using a frequency modulator. Also considering GMSK as a special case of PSK, it can be generated simply using phase modulator. Both types of GMSK modulators are briefly described here.



**Figure 3.2.16** GMSK Modulator using Frequency Modulator

The binary data sequence is encoded using a bipolar NRZ encoder.

The resulting data stream is then applied through a Gaussian low-pass filter whose characteristics are Gaussian in nature.

The filtered signal acts as modulating signal which modulates the carrier signal in frequency modulator.

The output of the frequency modulator is a GMSK signal.

The bipolar NRZ encoded information binary digital data sequence is converted into two parallel data sequences using a serial-to-parallel converter.

The resulting data streams are then applied to in-phase path and quadrature path respectively, each comprising of an integrator and Gaussian filter.

Their outputs are applied to independent balanced modulators whose carrier signals are having a phase shift of  $90^\circ$  with each other. This operation is similar to phase modulation.

Both outputs are added to produce the desired GMSK signal.

It may be noted that in GMSK, each transmitted symbol spans several bit periods.

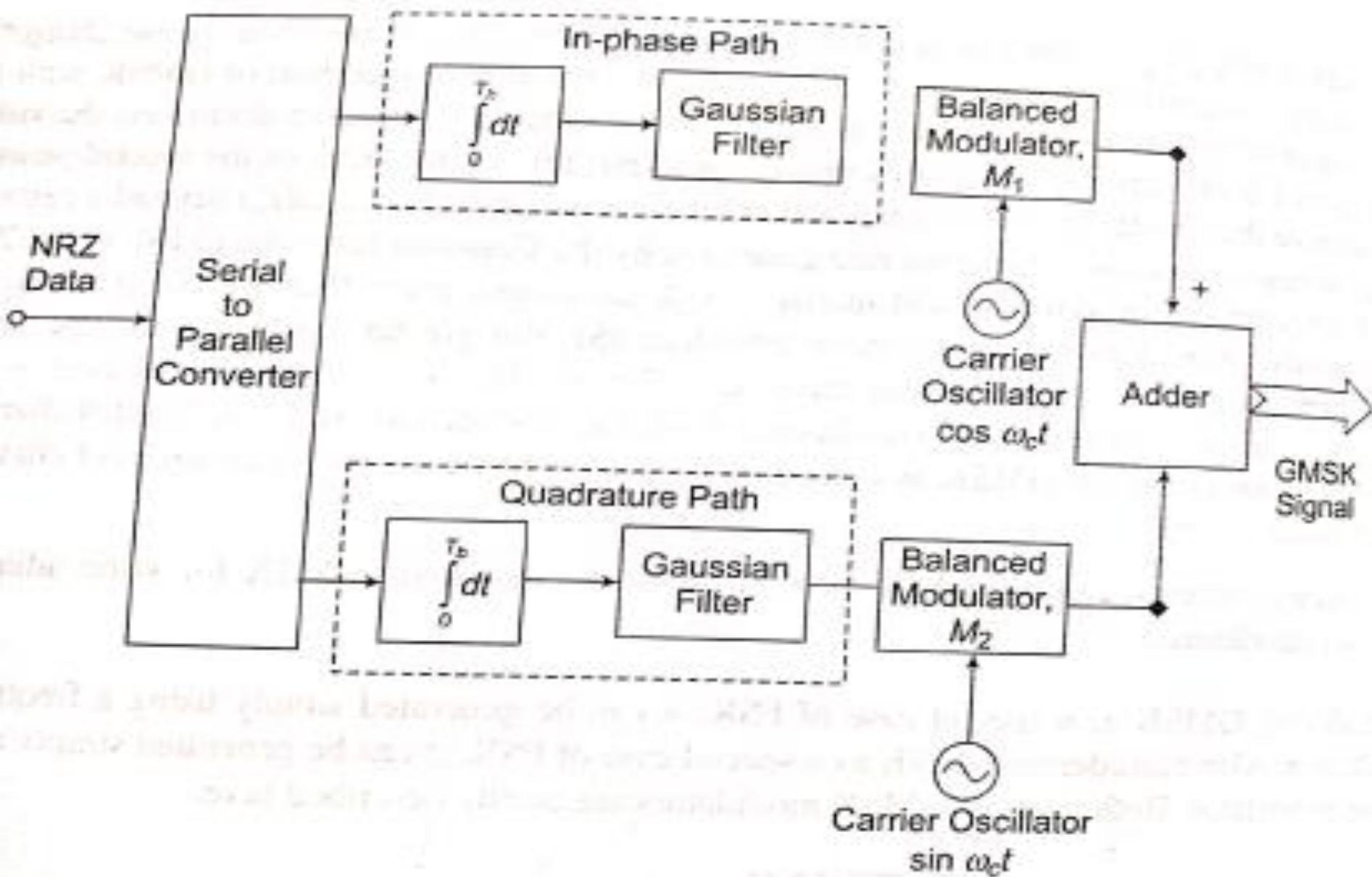


Figure 3.2.17 GMSK Modulator using Phase Modulator

GMSK can be non-coherently detected as in FSK demodulator, or coherently detected as in MSK demodulator.

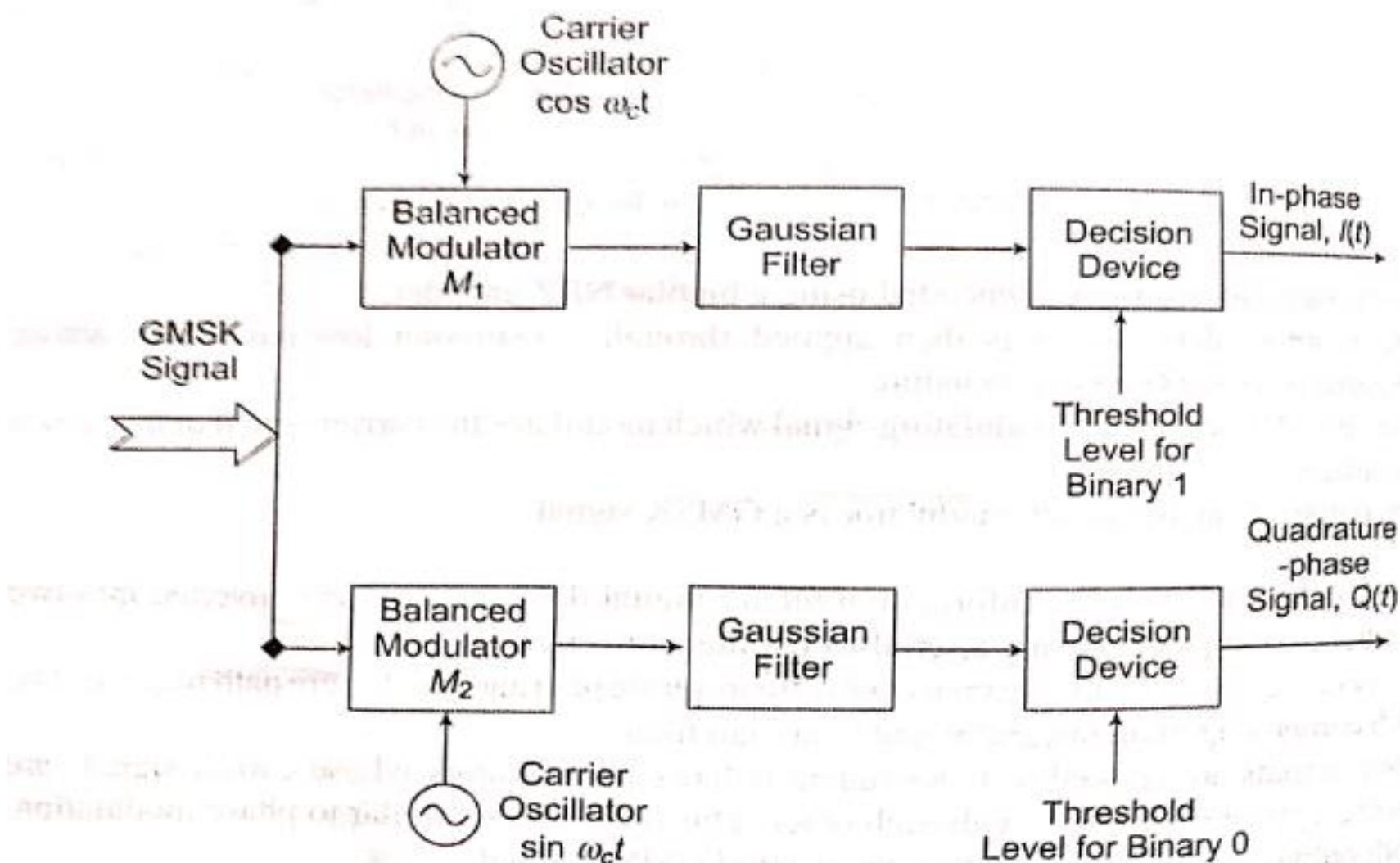
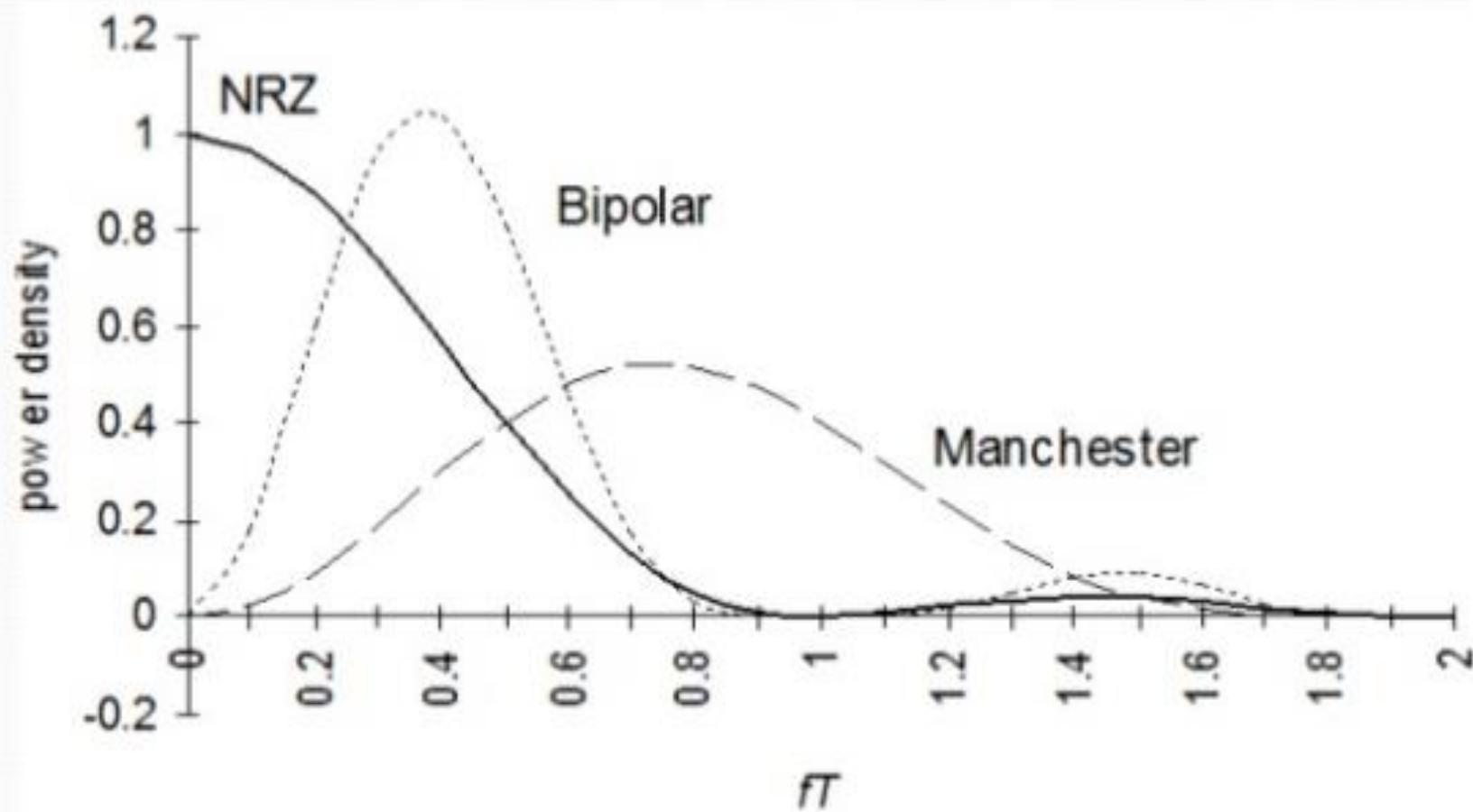


Figure 3.2.18 GMSK Demodulator

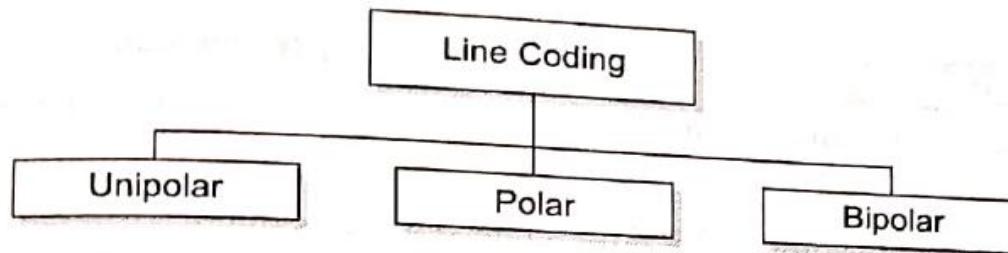
## PSD of Lines Codes:



## 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).<sup>4</sup>

**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.

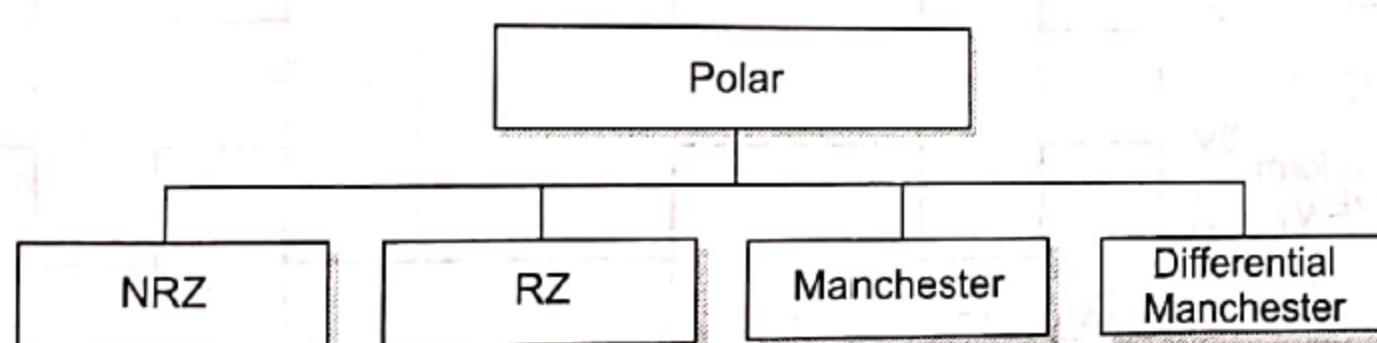


Figure 2.1.8 Classification of Polar Line Codes

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).