UNIT III DIGITAL COMMUNICATION

INTRODUCTION:

2.2 Shannon Limit for Information Capacity

Definition of information capacity : It is an ability of the system to carry number of independent symbols in a given unit of time. The capacity is expressed in bits per second.

Shannon's limit for information capacity

The capacity of a white bandlimited gaussian channel is given as,

$$C = B\log_2\left(1+\frac{S}{N}\right) \text{ bits/sec} \quad \text{...(2.2.1)}$$

Here C is the channel capacity

B is the channel bandwidth

$\frac{S}{N}$ is the signal to noise power ratio.

2.3 Digital Amplitude Modulation or Amplitude Shift Keying

The amplitude shift keying is also called on-off keying (OOK). This is the simplest digital modulation technique. The binary input data is converted to unipolar NRZ signal. A product modulator takes this NRZ signal and carrier signal. The output of the product modulator is the ASK signal, which can be expressed mathematically as,

$$v(t) = d\sin(2\pi f_c t) \quad \text{... (2.3.1)}$$

Here f_c is the carrier frequency

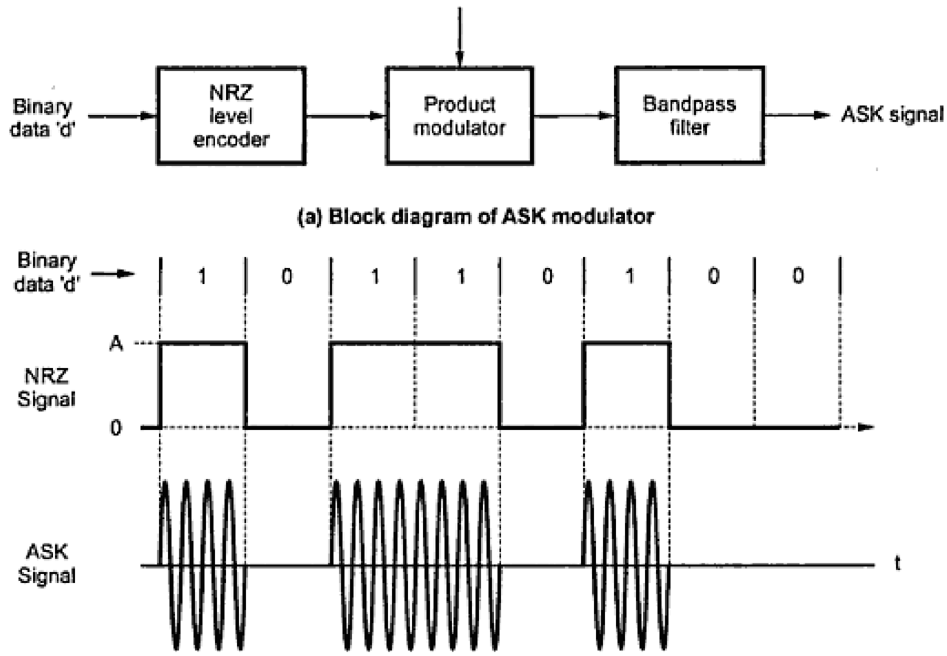
and d is the data bit, which is either 1 or 0.

Fig. 2.3.1 (a) shows the block diagram of the ASK modulator. The binary data sequence 'd' is given to the NRZ level encoder. This NRZ level encoder converts the input binary sequence to the signal suitable for product modulator. The product modulator also accepts a sinusoidal carrier of frequency f_c. The output of the product modulator is passed through a bandpass filter for bandwidth limiting. The output of the bandpass filter is the ASK signal. This signal and other waveforms are shown in Fig. 2.3.1 (b). Observe that the ASK signal has on-off nature. In equation 2.3.1 when $d=0$, $v(t)=0$; i.e. no ASK signal. And when $d=1$, $d=\sin(2\pi f_c t)$. The ASK is very sensitive to noise. It is used for very low bit rates less than around 100 *bps*. The only advantage of ASK is that it is very simple to implement.

Baud rate

For ASK, the ASK waveform is changed at the bit rate. Hence Baud rate is given as,

$$\text{Baud rate} = f_b \quad \text{... (2.3.2)}$$



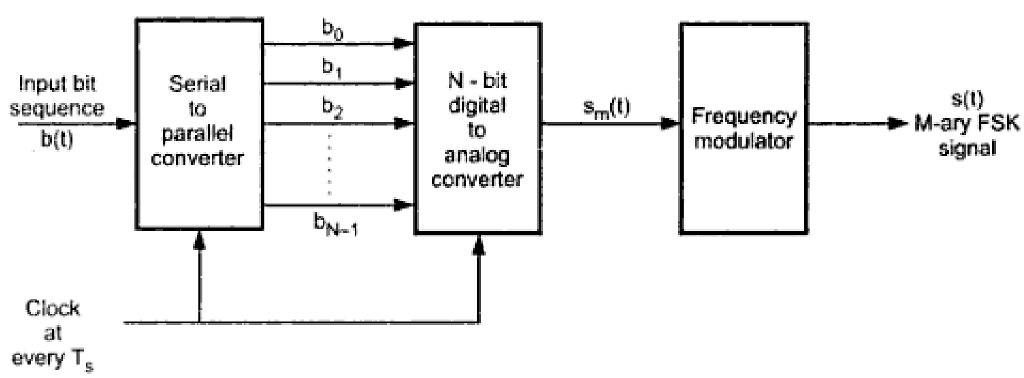
(b) Waveforms

Fig. 2.3.1 Amplitude shift keying (ASK)

FREQUENCY SHIFT KEYING:

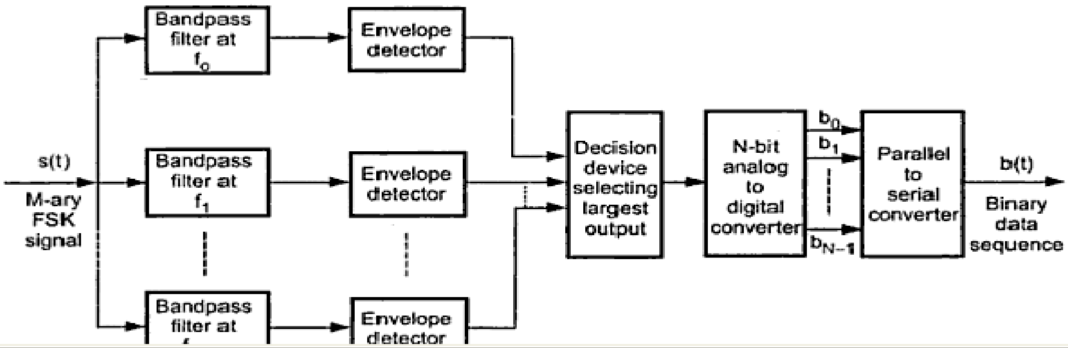
3.8.1.1 Transmitter

Fig. 3.8.1 shows the M-ary FSK transmitter. The 'N' successive bits are presented in parallel to digital to analog converter. These 'N' bits forms a symbol at the output of digital to analog converter. There will be total $2^N = M$ possible symbols. The symbol is presented every $T_s = NT_b$ period. The output of digital to analog converter is given to a frequency modulator. Thus depending upon the value of symbol, the frequency modulator generates the output frequency. For every symbol, the frequency modulator produces different frequency output. This particular frequency signal remains at the output for one symbol duration. Thus for 'M' symbols, there are 'M' frequency signals at the output of modulator. Thus the transmitted frequencies are $f_0, f_1, f_2, \ldots f_{M-1}$ depending upon the input symbol to the modulator.



3.8.1.2 Receiver

Fig. 3.8.2 shows block diagram of M-ary FSK receiver. It is the extension of BFSK receiver of Fig. 3.8.1. The M-ary FSM signal is given to the set of 'M' bandpass filters. The center frequencies of those filters are $f_0, f_1, f_2, ...f_{M-1}$. These filters pass their particular frequency and alternate others. The envelope detectors outputs are applied to a decision device. The decision device produces its output depending upon the highest input. Depending upon the particular symbol, only one envelope detector will have higher output. The outputs of other detectors will be very low. The output of the decision device is given to 'N' bit analog to digital converter. The analog to digital converter output is the 'N' bit symbol in parallel. These bits are then converted to serial bit stream by parallel to serial converter. In some cases the bits appear in parallel. Then there is no need to use serial to parallel and parallel to serial converters.



3.8.2 Power Spectral Density and Bandwidth of M-ary FSK

We know that for M symbol $f_0, f_1, f_2 \ldots f_{m-1}$ frequencies are used for transmission. The probability of error is minimized by selecting those frequencies such that transmitted signals are mutually orthogonal. If those frequencies are selected as successive even harmonics of symbol frequency f_s, then transmitted signals will be orthogonal.

Let's say that the lowest carrier frequency f_0 is the k^{th} harmonic of symbol frequency i.e.,

$$f_0 = kf_s \qquad \ldots (3.8.1)$$

Then the other frequencies will be,

$$f_1 = (k+2) f_s, f_2 = (k+4) f_s \ \ldots \text{ etc} \qquad \ldots (3.8.2)$$

Thus every carrier frequency is separated by $2f_s$ from its nearest carriers. Fig.3.7.2 shows the power spectral density of BFSK (for two symbol FSK). In this plot the two symbol frequencies f_L and f_H are separated by $2f_s$ (Here $f_s = f_b$ for BFSK). The same principle of BFSK is extended to M-ary FSK. That is M-carriers are added with separation of $2f_s$ between the carrriers (Note here that f_s is symbol frequency and not f_b). Therefore power spectral density for M-ary FSK will be simply extension of BFSK. Fig. 3.8.3 shows the power spectral density of M-ary FSK.

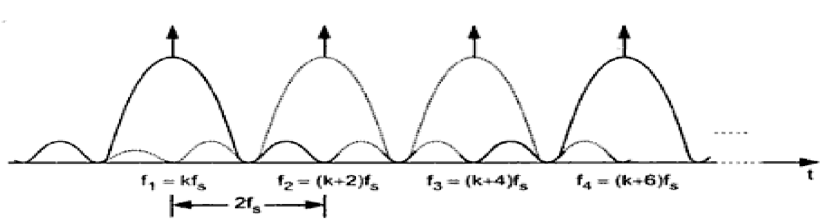


Fig. 3.8.3 Power spectral density M-ary FSK

Bandwidth of M-ary FSK :

From Fig. 3.8.3 it is clear that the width of one main lobe is $2f_s$. If there are M-symbols, then power spectral density spectrum will have M lobes. Therefore bandwidth of the system for M-symbols will be

$$BW = M \times (2f_s)$$

$$= 2Mf_s \quad \text{... (3.8.3)}$$

We know that $2^N = M$ and $f_s = \frac{f_b}{N}$ we can write the above equations,

$$BW = 2 \cdot 2^N \cdot \frac{f_b}{N} \quad \text{... (3.8.4)}$$

$$= \frac{2^{N+1} f_b}{N} \quad \text{... (3.8.5)}$$

4.2 Binary Phase Shift Keying (BPSK)

4.2.1 Principle of BPSK

- In binary phase shift keying (BPSK), binary symbol '1' and '0' modulate the phase of the carrier. Let the carrier be,

$$s(t) = A \cos(2\pi f_0 t) \quad \text{... (4.2.1)}$$

'A' represents peak value of sinusoidal carrier. In the standard 1Ω load register, the power dissipated will be,

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

$$\therefore \quad A = \sqrt{2P} \quad \text{... (4.2.2)}$$

- When the symbol is changed, then the phase of the carrier is changed by 180 degrees (π radians).
- Consider for example,

$$\text{Symbol '1'} \Rightarrow s_1(t) = \sqrt{2P} \cos(2\pi f_0 t) \quad \text{... (4.2.3)}$$

if next symbol is '0' then,

$$\text{Symbol '0'} \Rightarrow s_2(t) = \sqrt{2P} \cos(2\pi f_0 t + \pi) \quad \text{... (4.2.4)}$$

Since $\cos(\theta + \pi) = -\cos\theta$, we can write above equation as,

$$s_2(t) = -\sqrt{2P} \cos(2\pi f_0 t) \quad \text{... (4.2.5)}$$

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

$$\boxed{s(t) = b(t)\sqrt{2P} \cos(2\pi f_0 t)} \quad \text{... (4.2.6)}$$

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

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

4.2.2 Graphical Representation of BPSK Signal

Fig. 4.2.1 shows binary signal and its equivalent signal $b(t)$.

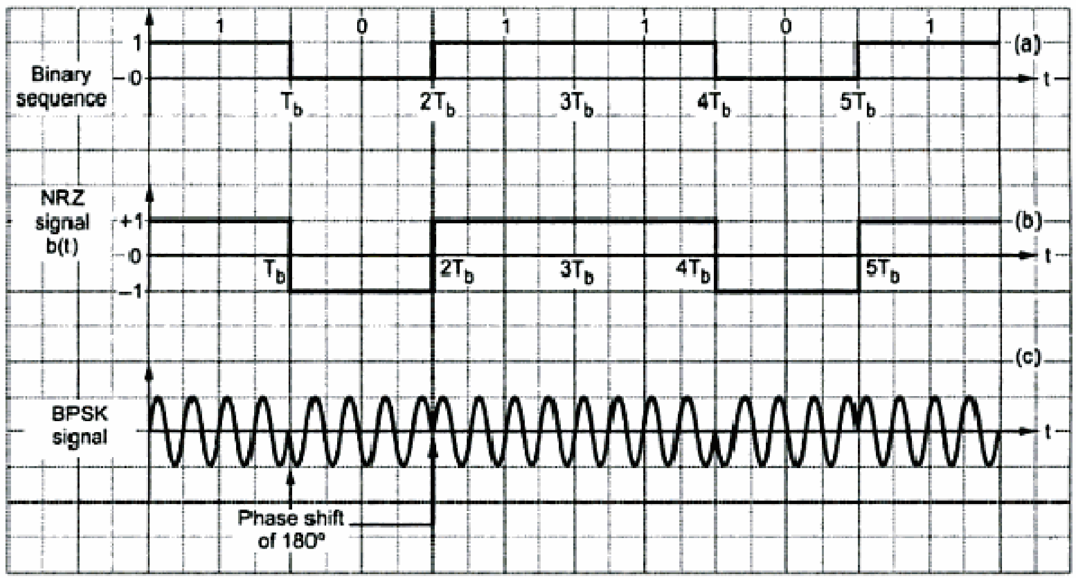


Fig. 4.2.1 (a) Binary sequence
(b) Its equivalent bipolar signal $b(t)$
(c) BPSK signal

As can be seen from Fig. 4.2.1 (b), the signal $b(t)$ is NRZ bipolar signal. This signal directly modulates carrier $cos(2\pi f_0 t)$.

4.2.3 Generation and Reception of BPSK Signal

4.2.3.1 Generation of BPSK Signal

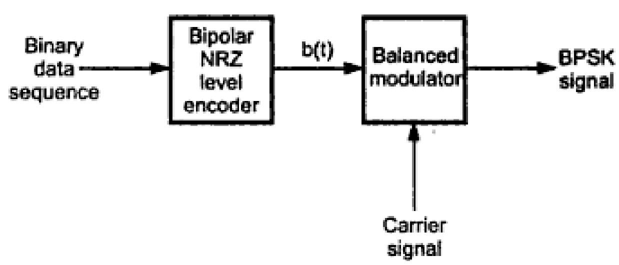


Fig. 4.2.2 BPSK generation scheme

- The BPSK signal can be generated by applying carrier signal to the balanced modulator.
- The baseband signal $b(t)$ is applied as a modulating signal to the balanced modulator. Fig. 4.2.2 shows the block diagram of BPSK signal generator.
- The NRZ level encoder converts the binary data sequence into bipolar NRZ signal.

4.2.3.2 Reception of BPSK Signal

Fig. 4.2.3 shows the block diagram of the scheme to recover baseband signal from BPSK signal. The transmitted BPSK signal is,

$$s(t) = b(t)\sqrt{2P}\,cos(2\pi f_0 t)$$

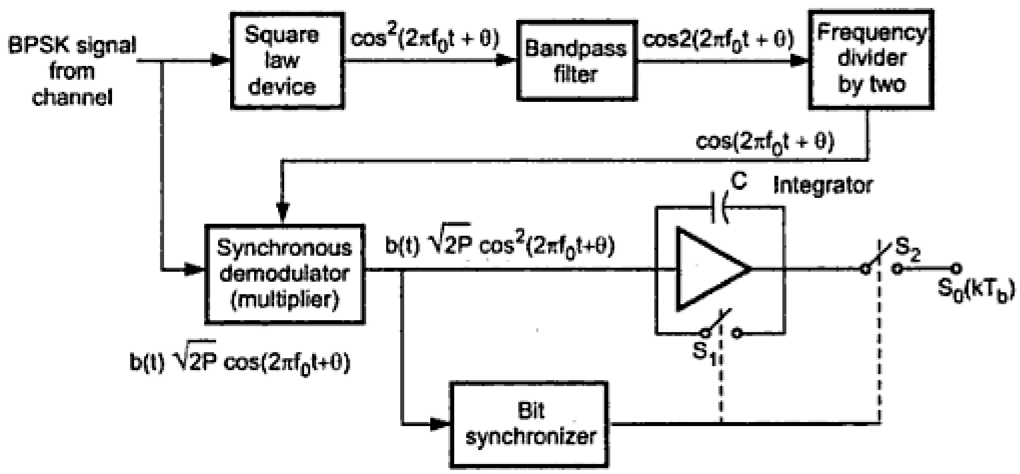


Fig. 4.2.3 Reception BPSK scheme

Operation of the receiver

1) **Phase shift in received signal** : This signal undergoes the phase change depending upon the time delay from transmitter to receiver. This phase change is normally fixed phase shift in the transmitted signal. Let the phase shift be θ. Therefore the signal at the input of the receiver is,

$$s(t) = b(t)\sqrt{2P}\cos(2\pi f_0 t + \theta) \qquad \text{... (4.2.7)}$$

2) **Square law device** : Now from this received signal, a carrier is separated since this is coherent detection. As shown in the figure, the received signal is passed through a square law device. At the output of the square law device the signal will be,

$$\cos^2(2\pi f_0 t + \theta)$$

Note here that we have neglected the amplitude, because we are only interested in the carrier of the signal.

We know that,

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

$$\therefore \quad \cos^2(2\pi f_0 t + \theta) = \frac{1+\cos 2(2\pi f_0 t + \theta)}{2}$$

4.2.4 Spectrum of BPSK Signals

Step 1 : Fourier transform of basic NRZ pulse.

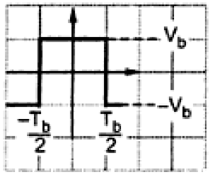


Fig. 4.2.4 NRZ pulse

We know that the waveform $b(t)$ is NRZ bipolar waveform. In this waveform there are rectangular pulses of amplitude $\pm V_b$. If we say that each pulse is $\pm \frac{T_b}{2}$ around its center as shown in Fig. 4.2.4. then it becomes easy to find fourier transform of such pulse. The fourier transform of this type of pulse is given as,

$$X(f) = V_b T_b \frac{\sin(\pi f T_b)}{(\pi f T_b)} \quad \text{By standard relations} \quad \dots (4.2.10)$$

Step 2 : PSD of NRZ pulse.

For large number of such positive and negative pulses the power spectral density $S(f)$ is given as

$$S(f) = \frac{\overline{|X(f)|^2}}{T_s} \quad \dots (4.2.11)$$

Here $\overline{X(f)}$ denotes average value of $X(f)$ due to all the pulses in $b(t)$. And T_s is symbol duration. Putting value of $X(f)$ from equation 4.2.10 in equation 4.2.11 we get,

Plot of PSD

- Equation 4.2.12 gives power spectral density of the NRZ waveform. For one rectangular pulse, the shape of $S(f)$ will be a sinc pulse as given by equation 4.2.12. Fig. 4.2.5 shows the plot of magnitude of $S(f)$.

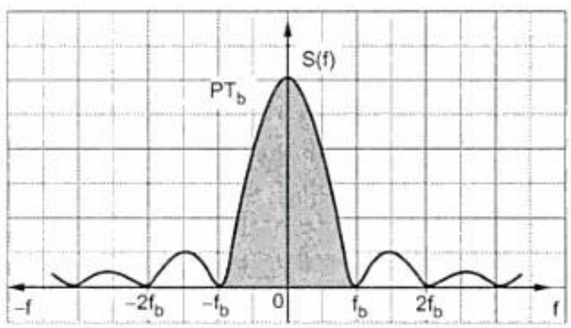


Fig. 4.2.5 Plot of power spectral density of NRZ baseband signal

Above figure shows that the main lobe ranges from $-f_b$ to $+f_b$. Here $f_b = \frac{1}{T_b}$.

Since we have taken $\pm V_b = \pm\sqrt{P}$ in equation 4.2.12, the peak value of the main lobe is PT_b.

- Now let us consider the power spectral density of BPSK signal given by equation 4.2.13. Fig. 4.2.6 shows the plot of this equation. The figure thus clearly shows that there are two lobes ; one at f_0 and other at $-f_0$. The same spectrum of Fig. 4.2.5 is placed at $+f_0$ and $-f_0$. But the amplitudes of main lobes are $\frac{PT_b}{2}$ in Fig. 4.2.6.

4.2.6 Bandwidth of BPSK Signal

The spectrum of the BPSK signal is centered around the carrier frequency f_0.

If $f_b = \frac{1}{T_b}$, then for BPSK the maximum frequency in the baseband signal will be

f_b see Fig. 4.2.6. In this figure the main lobe is centered around carrier frequency f_0 and extends from $f_0 - f_b$ to $f_0 + f_b$. Therefore Bandwidth of BPSK signal is,

$$BW = \text{Highest frequency} - \text{Lowest frequency in the main lobe}$$

$$= f_0 + f_b - (f_0 - f_b)$$

$$\therefore \quad \boxed{BW = 2f_b} \quad \text{... (4.2.21)}$$

Thus the minimum bandwidth of BPSK signal is equal to twice of the highest frequency contained in baseband signal.

4.4 Quadrature Phase Shift Keying (QPSK)

Principle

- In communication systems we know that there are two main resources, i.e. transmission power and the channel bandwidth. The channel bandwidth depends upon the bit rate or signalling rate f_b. In digital bandpass transmission, a carrier is used for transmission. This carrier is transmitted over a channel.
- If two or more bits are combined in some symbols, then the signalling rate is reduced. Therefore the frequency of the carrier required is also reduced. This reduces the transmission channel bandwidth. Thus because of grouping of bits in symbols, the transmission channel bandwidth is reduced.
- In quadrature phase shift keying, two successive bits in the data sequence are grouped together. This reduces the bits rate of signalling rate (i.e. f_b) and hence reduces the bandwidth of the channel.
- In BPSK we know that when symbol changes the level, the phase of the carrier is changed by 180°. Since there were only two symbols in BPSK, the phase shift occurs in two levels only.
- In QPSK two successive bits are combined. This combination of two bits forms four distinct symbols. When the symbol is changed to next symbol the

Since $b_o(t)$ and $b_e(t)$ cannot change at the same time, the phase change in QPSK signal will be maximum $\pi/2$. This is clear from Fig. 4.4.3.

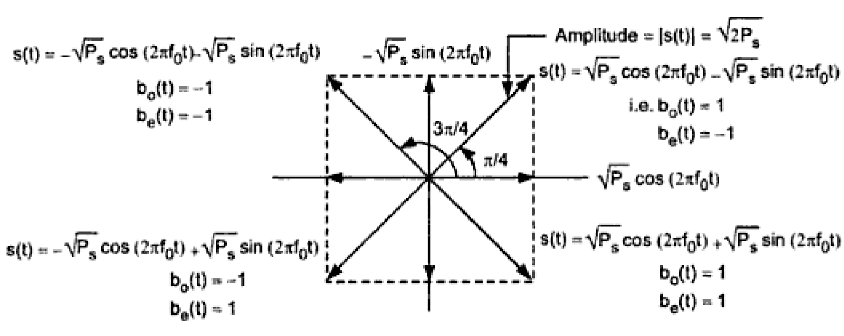


Fig. 4.4.3 Phasor diagram of QPSK signal

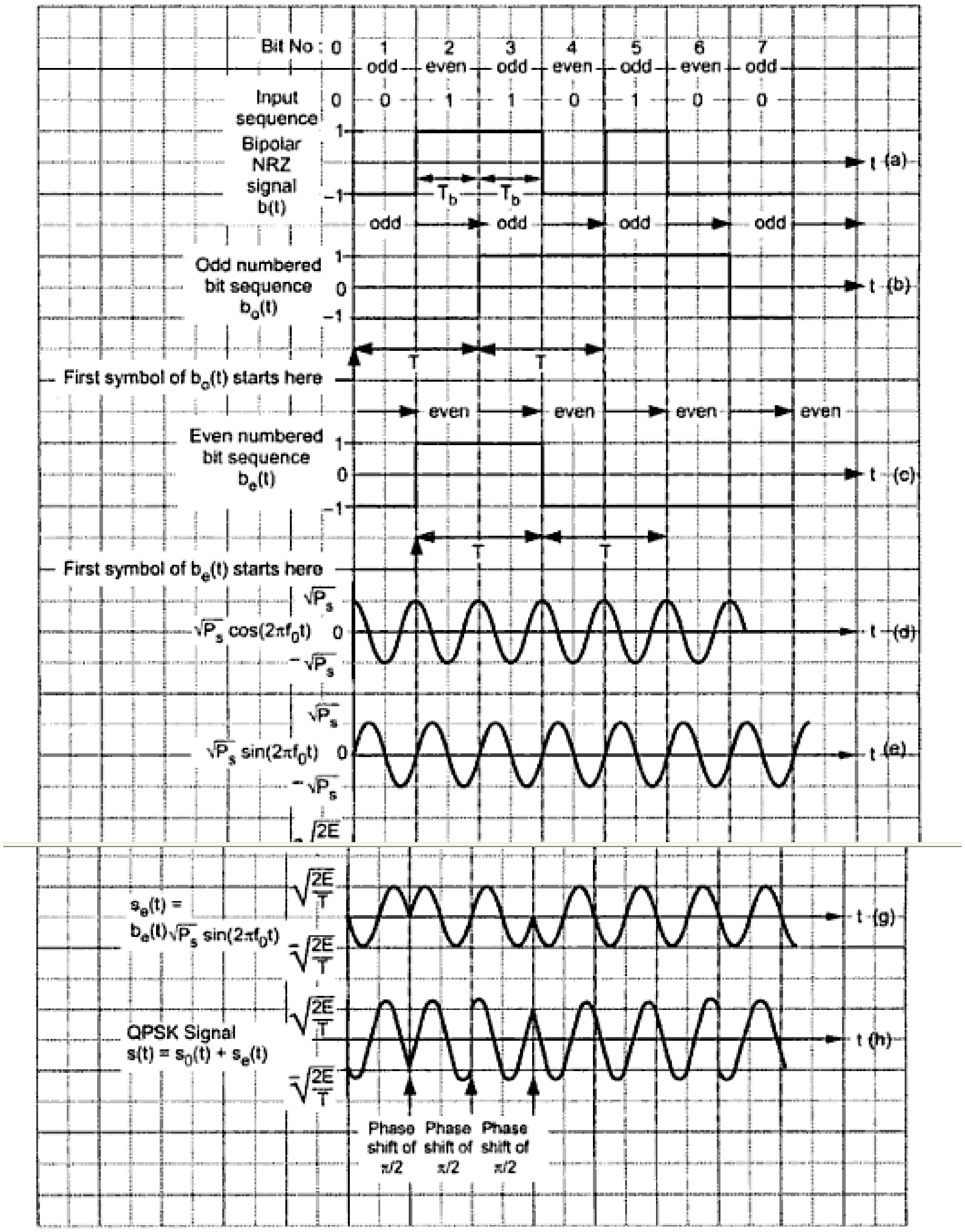


Fig. 4.4.2 QPSK waveforms (a) Input sequence and its NRZ waveform (b) Odd numbered bit sequence and its NRZ waveform (c) Even numbered bit sequence and its NRZ waveform (d) Basis function $\phi_1(t)$ (e) Basis function $\phi_2(t)$ (f) Binary PSK waveform for odd numbered channel (g) Binary PSK waveform for even numbered channel (h) Final QPSK waveform representing equation

4.4.1.3 The QPSK Receiver

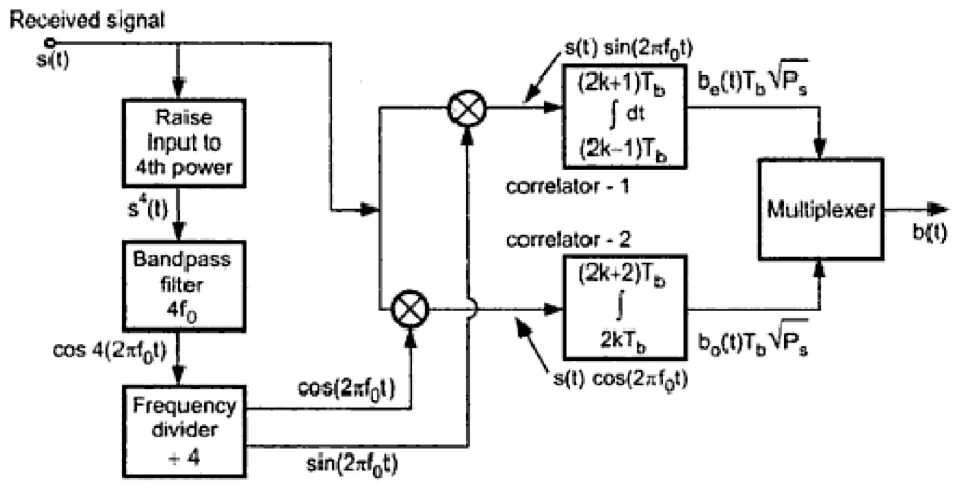


Fig. 4.4.4 QPSK receiver

Fig. 4.4.4 shows the QPSK receiver. This is synchronous reception. Therefore coherent carrier is to be recovered from the received signal $s(t)$.

Operation

Step 1 : Isolation of carrier

The received signal $s(t)$ is first raised to its 4^{th} power, i.e. $s^4(t)$. Then it is passed through a bandpass filter centered around $4f_0$. The output of the bandpass filter is a coherent carrier of frequency $4f_0$. This is divided by 4 and it gives two coherent quadrature carriers $\cos(2\pi f_0 t)$ and $\sin(2\pi f_0(t)$.

Step 2 : Synchronous detection

These coherent carriers are applied to two synchronous demodulators. These synchronous demodulators consist of multiplier and an integrator.

Step 3 : Integration over two bits interval

The incoming signal is applied to both the multipliers. The integrator integrates the product signal over two bit interval (i.e. $T_s = 2T_b$).

Step 4 : Sampling and multiplexing odd and even bit sequences

At the end of this period, the output of integrator is sampled. The outputs of the two integrators are sampled at the offset of one bit period, T_b. Hence the output of

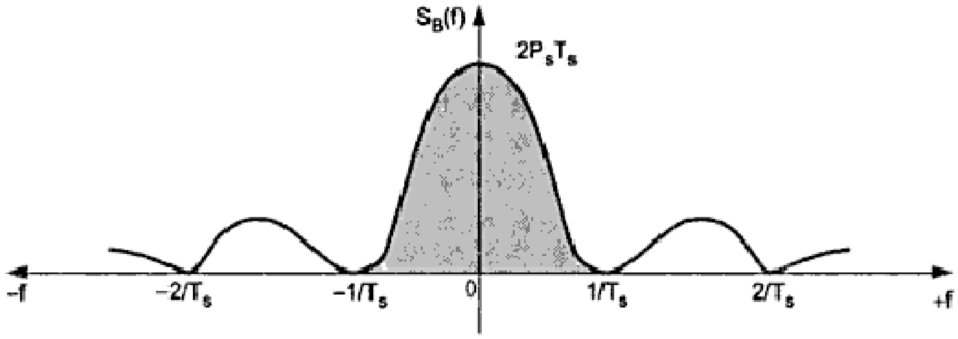


Fig. 4.4.7 Plot of power spectral density of QPSK signal

$$BW = \text{Highest frequency} - \text{Lowest frequency in main lobe}$$

$$= \frac{1}{T_s} - \left(-\frac{1}{T_s}\right) \text{since carrier frequency } f_0 \text{ cancels out}$$

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

We know that $T_s = 2T_b$

$$\therefore \quad BW = \frac{2}{2T_b} = \frac{1}{T_b} = f_b$$

which is same as we obtained in equation 4.4.24.

4.4.5 Advantages of QPSK

QPSK has some definite advantages and disadvantages as compared to BPSK and DPSK.

Advantages :

1) For the same bit error rate, the bandwidth required by QPSK is reduced to half as compared to BPSK.
2) Because of reduced bandwidth, the information transmission rate of QPSK is higher.
3) Variation in OQPSK amplitude is not much. Hence carrier power almost remains constant.

2.9.1 Bandwidth Efficiency (Information Density)

***Definition** : It is the ratio of transmission bit rate to minimum required bandwidth.*

i.e.,

$$\text{BW efficiency} = \frac{\text{Transmission rate (Bits / sec)}}{\text{Minimum bandwidth (cycles / sec)}}$$

$$= \frac{\text{Transmission rate}}{\text{Minimum bandwidth}} \text{ bits/cycle} \quad \ldots(2.9.1)$$

- When bandwidth efficiency is normalized to 1-Hz bandwidth, it gives number of bits that can be propagated per hertz of bandwidth.
- Bandwidth efficiency is used to compare the performance of digital modulation techniques.

2.10 Carrier Synchronization (Carrier Recovery)

The carrier synchronization is required in coherent detection methods to generate a coherent reference at the receiver. In this method the data bearing signal is modulated on the carrier in such a way that the power spectrum of the modulated carrier signal contains a discrete component at the carrier frequency. That is the fourier transform of the modulated signal contains one component at f_c also. Then the phase locked loop can be used to track this component f_c. The output frequency of phase locked loop is thus locked to the carrier frequency f_c in the transmitted signal. This output frequency of phase locked loop is used as a coherent reference signal for detection in the receiver.

2.10.1 Carrier Synchronization using Mth Power Loop

Fig. 2.10.1 shows the block diagram of carrier recovery or carrier synchronization circuit.

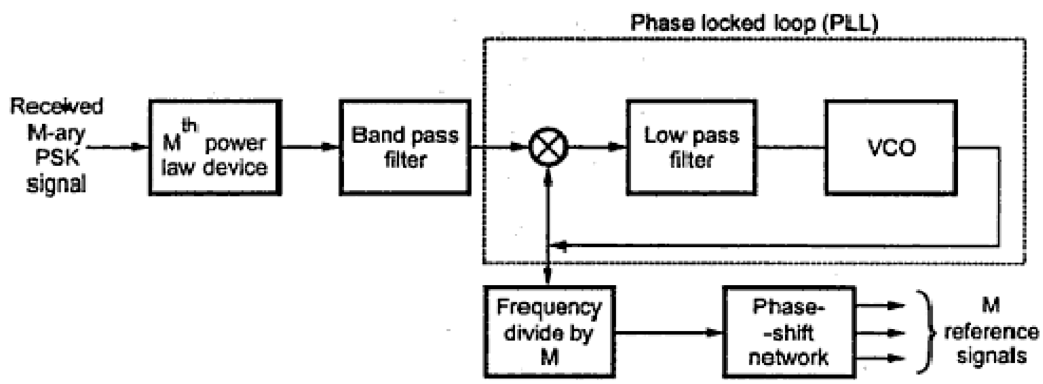


Fig. 2.10.1 Block diagram of Mth power loop

Fig. 2.10.1 shows the block diagram of carrier recovery circuit for M-ary PSK. This circuit is called the M^{th} power loop. When M = 2, then it is called squaring loop. When M=2, the M-ary PSK is then called as binary PSK. As shown in diagram, the input signal is first raised to the M^{th} power by the M^{th} power law device. Then the signal is passed through a bandpass filter. The bandpass filter is tuned to the carrier frequency f_c. The phase locked loop consists of a phase detector, low-pass filter and VCO. The phase locked loop tracks the carrier frequency. Then the output of a voltage controlled oscillator (VCO) is the carrier frequency. The output frequency of VCO is

divided by M. This is done because the M^{th} power of the input signal multiplies carrier frequency by M. The phase shift network then separates 'M' reference signals for the 'M' correlation receivers. In this technique the power of the input signal is raised to some power say 'M'. Let us say M = 2, then the input signal is squared. Because of this, the sign of the recovered carrier is always independent of sign of the input signal carrier since it is squared. Therefore there can be 180° error in the output.

2.10.2 Costas Loop for Carrier Synchronization May / June - 2006

This is the alternative method for carrier synchronization. This is used for binary phase shift keying. The block diagram is shown in Fig. 2.10.2.

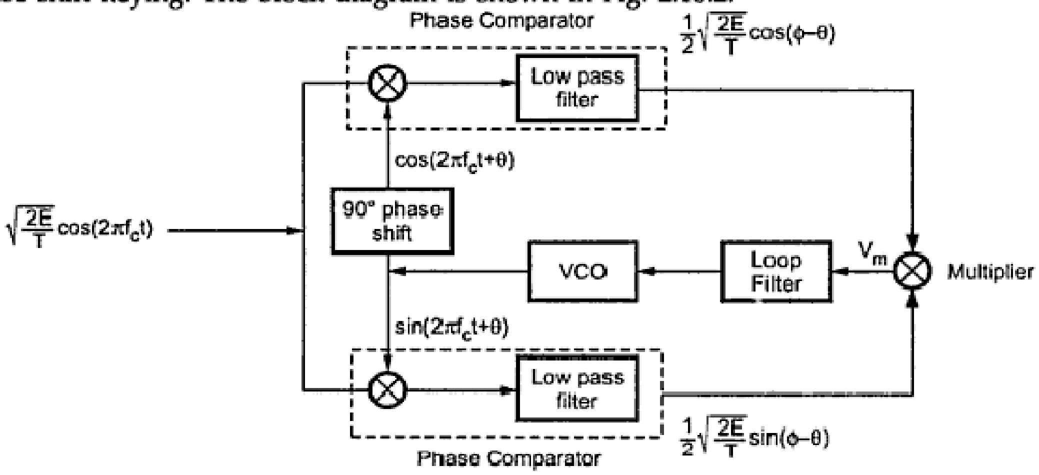


Fig. 2.10.2 The costas loop

As shown in Fig. 2.10.2 there are two phase locked loops. They have a common VCO and separate phase comparators. Let us assume that the VCO operates at the carrier frequency f_c with arbitrary phase angle θ. The BPSK signal is supplied to both the phase comparators. The low-pass filters remove the double frequency terms generated in the phase comparators and generate,

$\frac{1}{2}\sqrt{\frac{2E}{T}}\cos(\phi-\theta)$ and $\frac{1}{2}\sqrt{\frac{2E}{T}}\sin(\phi-\theta)$. The multiplier output is given as,

$$V_m = \frac{1}{4}\times\frac{2E}{T}\sin(\phi-\theta)\cos(\phi-\theta) \quad \text{... (2.10.1)}$$

$$= \frac{E}{2T}\cdot\frac{1}{2}\sin 2(\phi-\theta) \quad \text{..(2.10.2)}$$

$$= \frac{E}{4T}\sin 2(\phi-\theta) \quad \text{... (2.10.3)}$$

The power 'P' of the signal over the period T is given by,

$$P = \frac{E}{T}$$

Therefore equation (2.10.3) can be written as,

$$V_m = \frac{P}{4}\sin 2(\phi-\theta) \quad \text{... (2.10.4)}$$

If there is some difference between the VCO frequency and the input carrier frequency then the phase difference $(\phi-\theta)$ is changed proportionally. The change in $(\phi-\theta)$ causes V_m to increase or decrease VCO frequency such that synchronization is maintained.

2.5 Differential Phase Shift Keying (DPSK)

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 b its 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.

2.5.1 DPSK Transmitter and Receiver

2.5.1.1 Transmitter / Generator of DPSK Signal

Fig. 2.5.1 shows the scheme to generate DPSK signal.

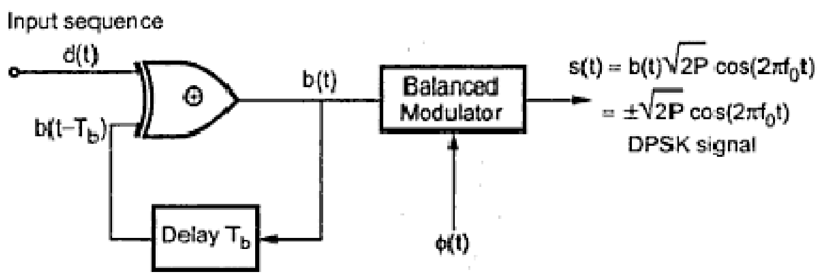


Fig. 2.5.1 Block diagram of DPSK generate or transmitter

The input sequence is $d(t)$. Output sequence is $b(t)$ and $b(t-T_b)$ is the previous output delayed by one bit period. Depending upon values of $d(t)$ and $b(t-T_b)$, exclusive OR gate generates the output sequence $b(t)$. Table 2.5.1 shows the truth table of this operation.

d(t)	*b (t − T_b)*	*b(t)*
0 (−1 V)	0 (−1 V)	0 (−1 V)
0 (−1 V)	1(1 V)	1(1 V)
1(1 V)	0 (−1 V)	1(1 V)
1(1 V)	1(1 V)	0 (−1 V)

Table 2.5.1 Truth table of exclusive OR gate

An arbitrary sequence $d(t)$ is taken. Depending on this sequence, $b(t)$ and $b(t-T_b)$ are found. These waveforms are shown in Fig. 2.5.2. The above Table 2.5.1 is used to derive the levels of these waveforms.

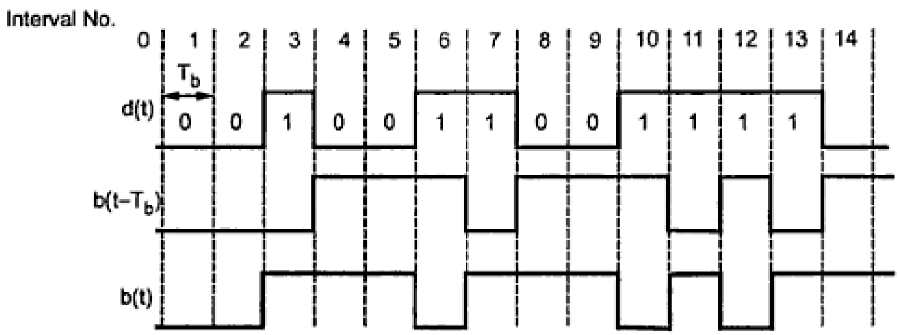


Fig. 2.5.2 DPSK waveforms

From the above waveform it is clear that $b(t-T_b)$ is the delayed version of $b(t)$ by one bit period T_b. The exclusive OR operation is satisfied in any interval i.e. in any interval $b(t)$ is given as,

$$b(t) = d(t) \oplus b(t-T_b) \quad \text{... (2.5.1)}$$

While drawing the waveforms the value of $b(t-T_b)$ is not known initially in interval no.1. Therefore it is assumed to be zero and then waveforms are drawn. We can write some important conclusions from the waveforms

1. Output sequence $b(t)$ changes level at the beginning of each interval in which $d(t)=1$ and it does not changes level when $d(t)=0$. Observe that $d(3)=1$, hence level of $b(3)$ is changed at the beginning of interval 3. Similarly in intervals 10, 11, 12 and 13 $d(t)=1$. Hence $b(t)$ is changed at the starting of these intervals. In interval 8 and 9 $d(t)=0$. Hence $b(t)$ is not changed in these intervals.
2. When $d(t)=0$, $b(t)=b(t-T_b)$ and
   When $d(t)=1$, $b(t)=\overline{b(t-T_b)}$
3. In interval no.1. we have assumed $b(t-T_b)=0$ and we obtained the waveform as shown in Fig. 2.5.2. If we assume $b(t-T_b)=1$ in interval no. 1, then the waveform of $b(t)$ will be inverted. But still $b(t)$ changes the level at the beginning of each interval in which $d(t)=1$.
4. The sequence $b(t)$ modulates sinusoidal carrier.
5. When $b(t)$ changes the level, phase of the carrier is changed. Since $b(t)$ changes its level only if $d(t)=1$; It shows that phase of the carrier is changed only if $d(t)=1$. In PSK phase of the carrier changes on both the symbol '1' and '0'. Whereas in DPSK phase of the carrier changes only on symbol '1'. This is the main difference between PSK and DPSK.
6. Always two successive bits of $d(t)$ are checked for any change of level. Hence one symbol has two bits.

$\therefore$ Symbol duration (T) = Duration of two bits $(2T_b)$

i.e. $$T = 2T_b \quad \text{... (2.5.2)}$$

As shown in Fig. 2.5.1, the sequence $b(t)$ is applied to a balanced modulator. The balanced modulator is also supplied with a carrier $\sqrt{2P}\cos(2\pi f_0 t)$

The modulator output is,

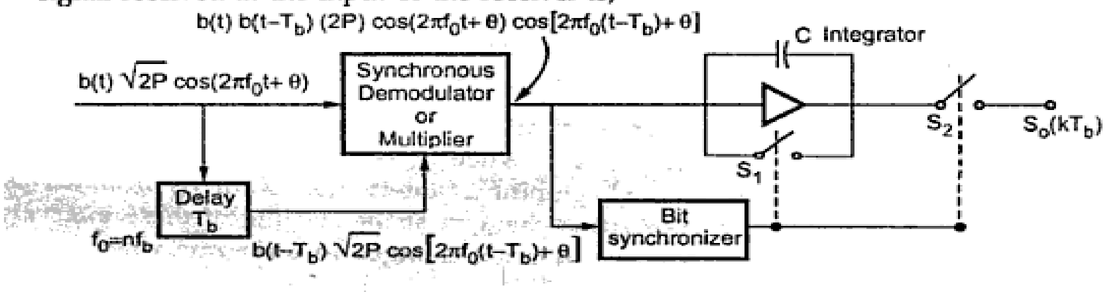
$$s(t) = b(t)\sqrt{2P}\cos(2\pi f_0 t) \quad \text{... (2.5.3)}$$

$$= \pm\sqrt{2P}\cos(2\pi f_0 t) \quad \text{... (2.5.4)}$$

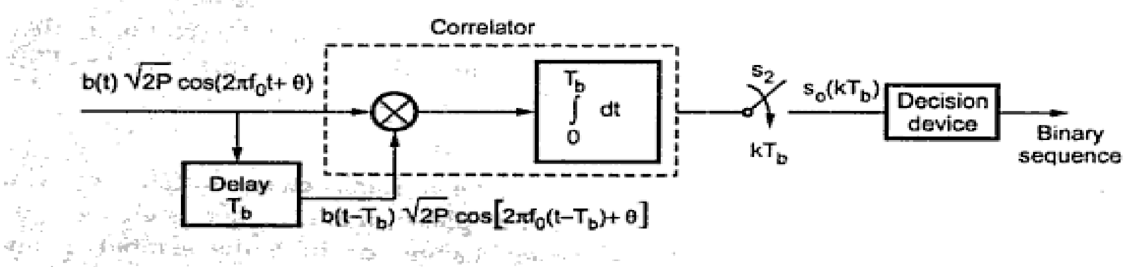
The above equation gives DPSK signal. Fig. 2.5.2 shows this DPSK waveforms. As shown in the waveforms the phase changes only when $d(t) = 1$.

2.5.1.2 DPSK Receiver

Fig. 2.5.3 shows the method to recover the binary sequence from DPSK signal. Fig. 2.5.3 (a) and (b) are equivalent to each other. Fig. 2.5.3(b) represents DPSK receiver using correlator. Fig. 2.5.3(a) shows multiplier and integrators separately. During the transmission, the DPSK signal undergoes some phase shift θ. Therefore the signal received at the input of the receiver is,



(a)



(b)

Received signal $= b(t)\sqrt{2P}\cos(2\pi f_0 t + \theta)$ ... (2.5.5)

This signal is multiplied with its delayed version by one bit. Therefore the outpu of the multiplier is,

Multiplier output $= b(t)\, b(t-T_b)\,(2P)\cos(2\pi f_0 t + \theta)\cos\left[2\pi f_0 (t-T_b) + \theta\right]$ ... (2.5.6)

We know that, $\cos(A)\cos(B) = \frac{1}{2}\left[\cos(A-B) + \cos(A+B)\right]$

Here $A = 2\pi f_0 t + \theta$ and $B = 2\pi f_0 (t - T_b) + \theta$

$\therefore$ Multiplier output $= b(t)\, b(t-T_b)\, P\left\{\cos 2\pi f_0 T_b + \cos\left[4\pi f_0\left(t - \frac{T_b}{2}\right) + 2\theta\right]\right\}$ ...(2.5.7)

f_0 is the carrier frequency and T_b is one bit period. T_b contains integral number o cycles of f_0. We know that,

$$f_b = \frac{1}{T_b}$$

If T_b contains 'n' cycles of f_0 then we can write,

$$f_0 = n\, f_b \quad \Rightarrow \quad f_0 = \frac{n}{T_b}$$

$\therefore \quad f_0 T_b = n$ ... (2.5.8)

Putting $f_0 T_b = n$ in first cosine term in equation (2.5.7) we get

$$\text{Multiplier output} = b(t)\, b(t-T_b)\, P\left\{\cos 2\pi n + \cos\left[4\pi f_0\left(t - \frac{T_b}{2}\right) + 2\theta\right]\right\}$$

This signal is given to the integrator. In the k^{th} bit interval, the integrator output can be written as,

$$s_o(kT_b) = b(kT_b)\, b\left[(k-1)T_b\right] P \int_{(k-1)T_b}^{kT_b} dt + b(kT_b)\, b\left[(k-1)T_b\right] P \int_{(k-1)T_b}^{kT_b} \cos\left[4\pi f_0\left(t-\frac{T_b}{2}\right)+2\theta\right] dt$$

The integration of the second term will be zero since it is integration of carrier over one bit duration. The carrier has integral number of cycles over one bit period hence integration is zero. Therefore we can write,

$$\begin{aligned} s_o(kT_b) &= b(kT_b)\, b\left[(k-1)T_b\right] P\left[kT_b-(k-1)T_b\right] \\ &= b(kT_b)\, b\left[(k-1)T_b\right] P T_b \qquad \text{... (2.5.10)} \end{aligned}$$

Here know that $PT_b = E_b$; i.e. energy of one bit. The product $b(kT_b)\, b\left[(k-1)T_b\right]$ decides the sign of PT_b.

The transmitted data bit $d(t)$ can be verified easily from product $b(kT_b)\, b\left[(k-1)T_b\right]$. We know from Fig. 2.5.2 when $b(t) = b(t-T_b)$, $d(t) = 0$. That is if both are $+1\,V$ or $-1\,V$ then $b(t)\, b(t-T_b) = 1$. Alternately we can write,

If $\quad b(t)\, b(t-T_b) = 1\,V \quad$ then $\quad d(t) = 0$

We know that $b(t) = \overline{b(t-T_b)}$ then $d(t) = 1$. That is $b(t) = -1\,V$, $b(t-T_b) = +1\,V$ and vice versa. Therefore $b(t)\, b(t-T_b) = -1$. Alternately we can write,

If $\quad b(t)\, b(t-T_b) = -1\,V, \quad$ then $\quad d(t) = 1$

The decision device is shown in Fig. 2.5.3 (b). We know that,

$$s_o(kT_b) = b(kT_b)\, b\left[(k-1)T_b\right] PT_b \qquad \text{... from equation 2.5.10}$$

If $$s_o(kT_b) = \begin{cases} -PT_b, \text{ then } d(t) = 1 \text{ and} \\ +PT_b, \text{ then } d(t) = 0 \end{cases}$$