CBCS SCHEME

USN | | | | | | | | | | | **15EC61**

Sixth Semester B.E. Degree Examination, Dec.2019/Jan.2020
Digital Communication

Time: 3 hrs. Max. Marks: 80

Note: Answer any FIVE full questions, choosing ONE full question from each module.

Important Note : 1. On completing your answers, compulsorily draw diagonal cross lines on the remaining blank pages.
2. Any revealing of identification, appeal to evaluator and /or equations written eg. 42+8 = 50, will be treated as malpractice.

Module-1

1 a. Define Hilbert transform. State the properties of it. **(04 Marks)**

b. Obtain the Hilbert transform of
   i) $x(t) = (\cos 2\pi Ft + \sin 2\pi Ft)$
   ii) $x(t) = e^{j2\pi Ft}$ **(04 Marks)**

c. Explain canonical representation of band pass signal. **(08 Marks)**

OR

2 a. Derive the expression for the complex low pass representation of bandpass systems. **(08 Marks)**

b. For the given data stream 11011100. Sketch the line code
   i) Unipolar NRZ
   ii) Polar NRZ
   iii) Unipolar RZ
   iv) Bipolar NRZ **(04 Marks)**

c. Draw the power spectra of NRZ unipolar and NRZ polar format. **(04 Marks)**

Module-2

3 a. Show that the energy of a signal is equal to squared length of the signal vector. **(08 Marks)**

b. Obtain the decision rule for maximum likelihood decoding and explain the correlation receiver. **(08 Marks)**

OR

4 a. Explain the correlation receiver using product integrator and matched filter. **(08 Marks)**

b. Three signals $s_1(t)$, $s_2(t)$ and $s_3(t)$ are shown in Fig.Q.4(b). Apply Gram Schmidt procedure to obtain an orthonormal basis for the signals. Express signals $s_1(t)$, $s_2(t)$ and $s_3(t)$ in terms of orthonormal basis functions. **(08 Marks)**

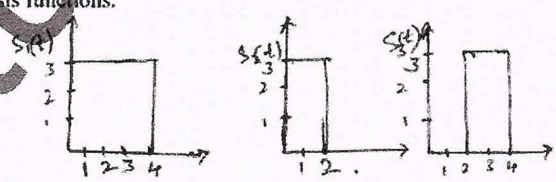

Fig.Q.4(b)

Module-3

5 a. With necessary diagrams, explain the generation and reception of BPSK signal. **(10 Marks)**

b. Given the binary data 10010011, draw the BPSK and DPSK waveforms. **(06 Marks)**

Scheme & Solution prepared by. Prof. Raghavendra. N

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OR

6 a. Derive the expression for error probability of BFSK. (08 Marks)
b. With block diagram explain generation and detection of DPSK. (08 Marks)

Module-4

7 a. What is ISI? Obtain the expression of output of a filter with intersymbol interference. (08 Marks)
b. Explain the Nyquist criterion for distortionless baseband binary transmission and obtain the ideal solution for zero ISI. (08 Marks)

OR

8 a. Draw and explain the time-domain and frequency domain of duobinary and modified duobinary signal. (08 Marks)
b. What is channel equalization? With a neat diagram, explain the concept of equalization using a linear transversal filter. (08 Marks)

Module-5

9 a. Draw the 4 stage linear feedback shift register with 1st and 4th state is connected to Modulo-2 adder. Output of Modulo-2 is connected to 1st stage input. Find the output PN sequence and write the autocorrelation function with initial state 1000. (06 Marks)
b. Explain the generation of direct sequence spread spectrum with relevant waveforms and spectrums. (07 Marks)
c. Write a short note on application of spread spectrum in wireless LAN's. (03 Marks)

OR

10 a. With necessary block diagram, explain the transmitter and receiver of frequency hop spread spectrum. (08 Marks)
b. With a neat block diagram, explain the CDMA system based on IS-95. (08 Marks)

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Scheme & Solution of Digital Communication

Dec 2019 / Jan. 2020 - 18EC61

Note: Answer any Five full Questions, choosing ONE full question from each module.

Module - 1

1. a) Define Hilbert Transform. state the properties of it — (4M)

→ Hilbert Transform (HT) of a signal $x(t)$ is a signal $\hat{x}(t)$ whose frequency components lag the frequency components of $x(t)$ by $90°$. In other words, $\hat{x}(t)$ has exactly the same frequency components present in $x(t)$ with same amplitude except there is a $90°$ phase delay. HT does not involve a domain change. HT of a signal $x(t)$ is another signal denoted by $\hat{x}(t)$ in the same domain (i.e., time domain).

For example: HT of a signal

$$x(t) = A\cos(2\pi f_0 t + \theta)$$

is given by

$$\hat{x}(t) = A\cos(2\pi f_0 t + \theta - 90°)$$

$$\hat{x}(t) = A\sin(2\pi f_0 t + \theta)$$

- 2M

Properties of Hilbert Transform

HT has number of useful properties of its own.

(1) A signal $x(t)$ and its HT $\hat{x}(t)$ have the same magnitude spectrum $|X(f)| = |\hat{X}(f)|$.

(2). If $\hat{g}(t)$ is HT of $g(t)$, then HT of $\hat{g}(t)$ is $-g(t)$

w.k.t. HT is equivalent to passing $g(t)$ through a linear two port device with a transfer function of $-j\,\mathrm{sgn}(f)$.

A twice HT is therefore equivalent to passing g(t) through cascade of a linear two port devices.

The overall transfer function of a cascaded linear two port devices is given by.

$$[-j\,sgn(f)]^2 = -1 \text{ for all } f.$$

Hence we can say that HT of $\hat{g}(t)$ is $-g(t)$.

③ A signal g(t) & its HT $\hat{g}(t)$ are orthogonal over entire time interval $(-\infty, \infty)$. — 2M

1. b) obtain the Hilbert transform of

i) $x(t) = \cos 2\pi f t + t \sin 2\pi f t$

ii) $x(t) = e^{-j2\pi f t}$ — (4M)

→

i) $x(t) = \cos 2\pi f t + \sin 2\pi f t$

$$\hat{x}(t) = \cos\left(2\pi f t - \frac{\pi}{2}\right) + \sin\left(2\pi f t - \frac{\pi}{2}\right)$$

$$\hat{x}(t) = \sin(2\pi f t) - \cos(2\pi f t)$$

— 2M

ii) $x(t) = e^{-j2\pi f t}$

$$x(t) = \cos 2\pi f t - j \sin 2\pi f t$$

$$\hat{x}(t) = \cos\left(2\pi f t - \frac{\pi}{2}\right) - j \sin\left(2\pi f t - \frac{\pi}{2}\right)$$

$$\hat{x}(t) = \sin(2\pi f t) + j\cos(2\pi f t)$$

$$\hat{x}(t) = -j e^{-2\pi f t}$$

— 2M

(Raghavendran)

②

1.c) Obtain the canonical representation of band pass signals. — (8M)

→ We can express the bandpass signal $s(t)$ in terms of its complex envelope as

$$s(t) = \mathrm{Re}\left[\hat{s}(t)\, e^{j2\pi f_c t}\right] \quad -①$$

$$\hat{s}(t) = s_I(t) + j\, s_Q(t) \quad -②$$

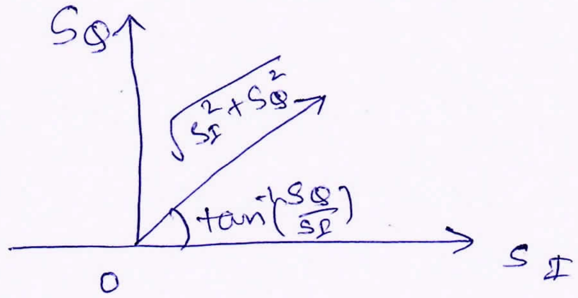
representing eqn ① in canonical form

$$s(t) = s_I(t)\cos(2\pi f_c t) - s_Q(t)\sin(2\pi f_c t)$$

where $s_I(t)$ — Inphase component

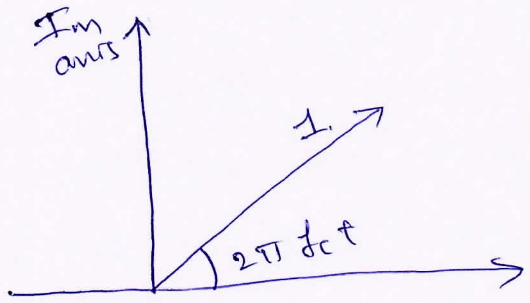
$s_Q(t)$ - quadrature component. — 1M.

Complex envelope $\hat{s}(t)$ may be pictured as a time varying phasor positioned at the origin of the (s_I, s_Q) plane in fig ① below:



— 1M

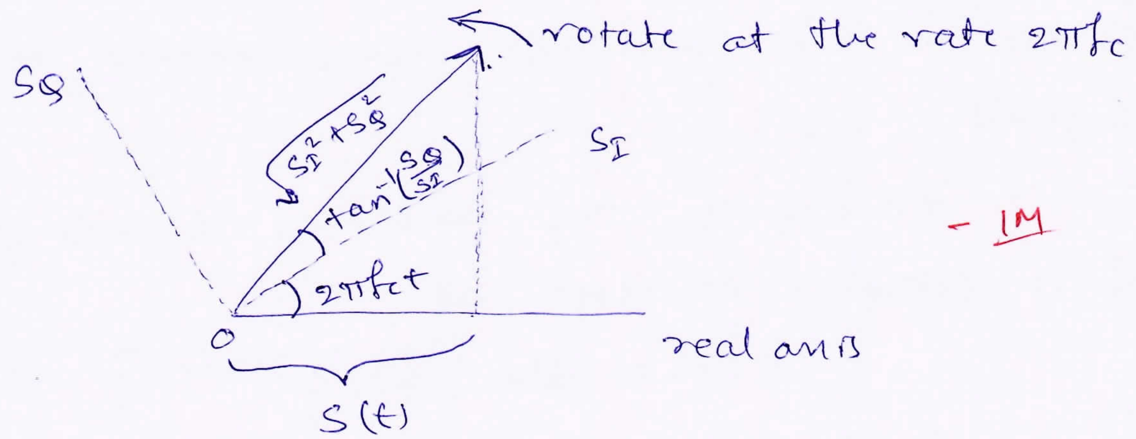
fig ① : phasor representation.



— 1M

fig ② : phasor representation of $e^{j2\pi f_c t}$

(Raghavendra. N)



fig(3): phasor representation for $S(t)$.

The above picture shows $\hat{S}(t)$ moves in (S_I, S_Q) plane while at the same time the plane itself rotates about the origin.

The original bandpass signal $s(t)$ is the projection of this time varying phasor on a real axis.

Alternatively we may define $\hat{S}(t)$ in polar form as

$$\hat{S}(t) = a(t)\, e^{[j\phi(t)]}$$ — 1M

where $a(t)$ and $\phi(t)$ are both real-valued low-pass functions.

$$\hat{S}(t) = a(t)\cos\left[2\pi f_c t + \phi(t)\right]$$

where $a(t)$ = natural envelope of bandpass signal

$\phi(t)$ = phase of the signal.

$$\therefore\ a(t) = |\hat{S}(t)| = |S_+(t)| = \sqrt{S_I^2(t) + S_Q^2(t)}$$

and $$\phi(t) = \tan^{-1}\left[\frac{S_Q(t)}{S_I(t)}\right]$$

and $$S_I(t) = a(t)\cos[\phi(t)]$$

$$S_Q(t) = a(t)\sin[\phi(t)]$$ — 3M

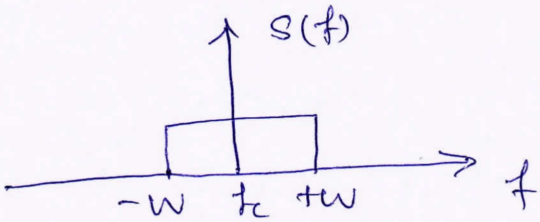
(Raghavendra.N)

3

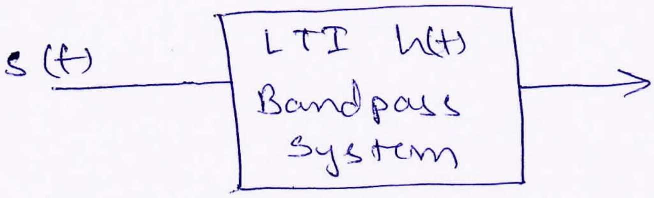
2. a) Derive the expression for the complex low pass representation of bandpass systems. — (8M)

→ Let $s(t)$ is a narrowband signal.

$S(f)$ its fourier transform



we also assume that $f_c > W$



— 1M

Let the signal $s(t)$ be applied to a linear-time invariant bandpass system. with impulse response $h(t)$ and freq. response $H(f)$.

The system BW $2B$ is usually narrower than or equal to the input signal BW $2W$.

Representing the band pass impulse response $h(t)$ in terms of 2 quadrature components.

$$h(t) = h_I(t)\cos(2\pi f_c t) - h_Q(t)\sin(2\pi f_c t) \quad -(1)$$

— 2M

Corresponding complex impulse response is

$$\hat{h}(t) = h_I(t) + j\,h_Q(t). \quad -(2)$$

Complex envelope $\hat{s}(t)$ of a bandpass signal $s(t)$ is

$$s(t) = \mathrm{Re}\left[\hat{s}(t)\, e^{j2\pi f_c t}\right]$$

— 2M

Expressing $h(t)$ in terms of $\hat{h}(t)$

(Raghavendran)

$$h(t) = Re\left[\hat{h}(t)\, e^{j2\pi f_c t}\right] \quad -(3)$$

then

$$2\,h(t) = \hat{h}(t)\, e^{j2\pi f_c t} + \hat{h}^*(t)\, e^{-j2\pi f_c t}$$

Here $\hat{h}^*(t)$ is complex conjugate of $\hat{h}(t)$ —(4)

— 1M

Applying FT to bothsides of eqn (4) and using complex conjugation property, we get

$$2\,H(f) = \hat{H}(f - f_c) + \hat{H}^*(-f - f_c)$$

since $\hat{H}^*(f - f_c) = 0$ for $f > 0$ then

$$\hat{H}(f - f_c) = 2\,H(f) \quad , \quad f > 0.$$

— 1M

The above eqn shows that, for a band pass freq. response $H(f)$, we can find complex low-pass freq response $\hat{H}(f)$ by taking i) $H(f)$ for $f > 0$, ii) shifting it to the origin, iii) scaling it by a factor of 2.

$\hat{H}(f)$ is decomposed into in-phase & quadrature phase components as

$$\hat{H}(f) = \hat{H}_I(f) + j\,\hat{H}_Q(f)$$

where $\hat{H}_I(f) = \frac{1}{2}\left[\hat{H}(f) + \hat{H}^*(f)\right]$

$$\hat{H}_Q(f) = \frac{1}{2j}\left[\hat{H}(f) - j\,\hat{H}^*(f)\right]$$

Using IFT, we get

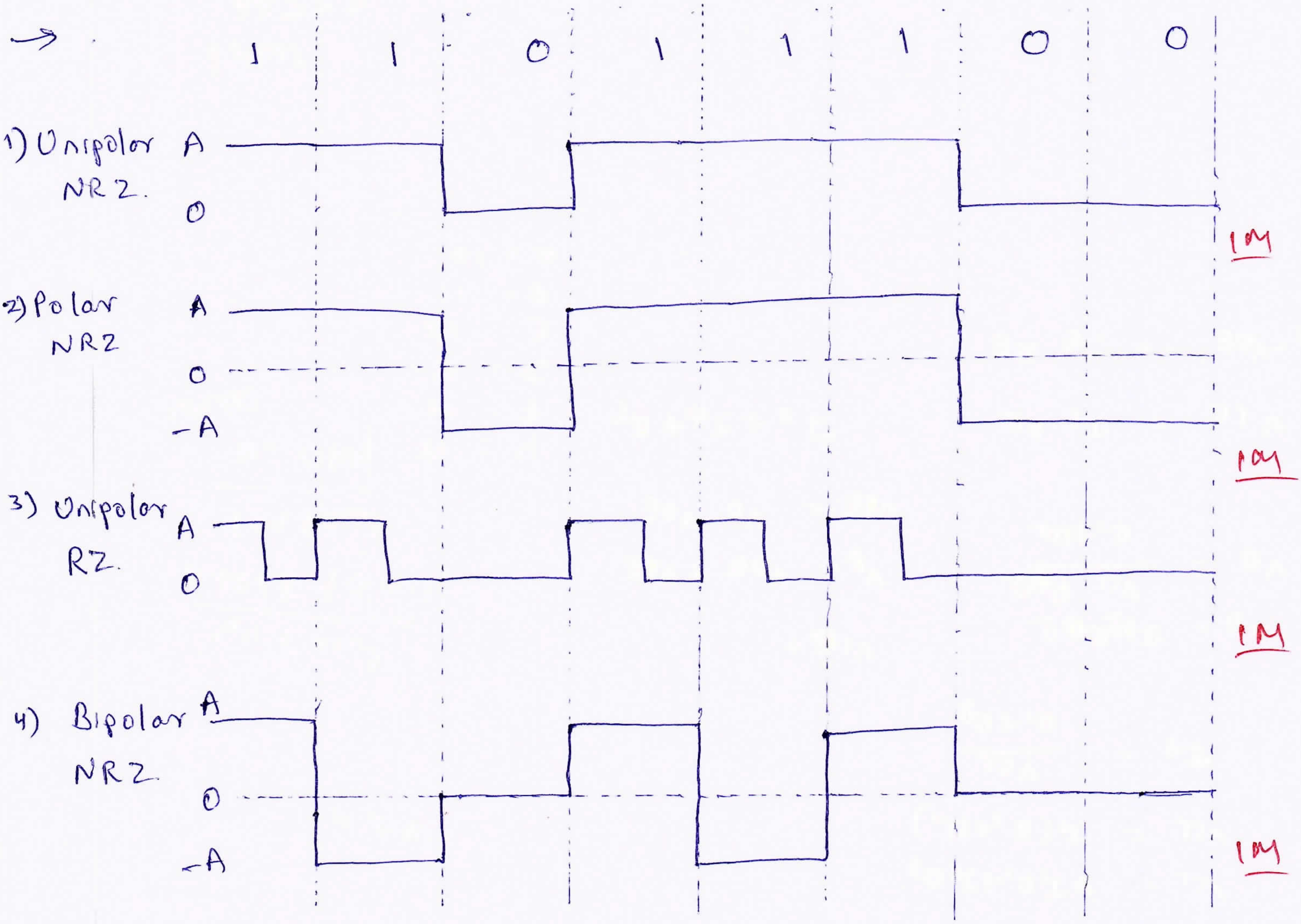
$$\hat{h}(t) = \int_{-\infty}^{\infty} \hat{H}(f)\, e^{j2\pi f_c t}\, df \, .$$

— 1M

(Prof R N)

2.b). For the given data stream 11011100. Sketch the line code. i) Unipolar NRZ ii) Polar NRZ iii) Unipolar RZ iv) Bipolar NRZ — (4M)

→



(RN)

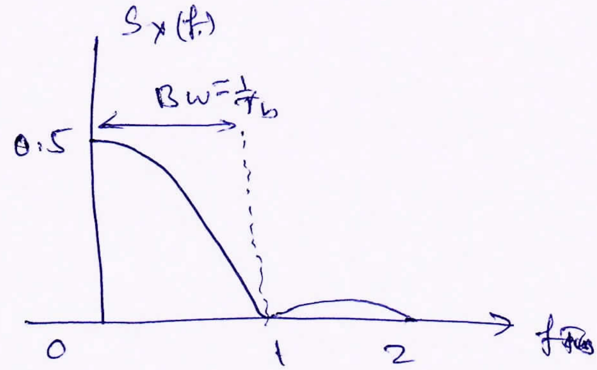
(8)

2 c). Draw the power spectra of NRZ unipolar and NRZ polar format. — (4M)

→ NRZ unipolar format

PSD of unipolar NRZ is given by.

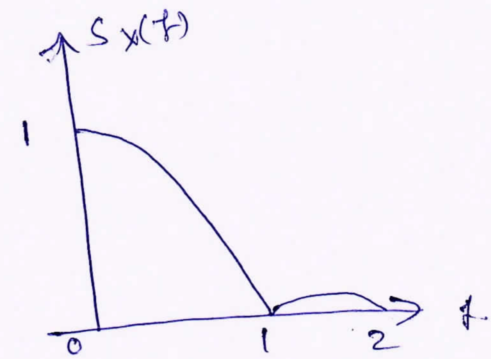
$$S_X(f) = \frac{A^2 T_b}{4} \operatorname{sinc}^2(f T_b) + \frac{A^2}{4} \delta(f)$$



— 2M

NRZ polar format

$$S_X(f) = A^2 T_b \operatorname{sinc}^2(f T_b)$$



— 2M

(RN)

6

Module - 2.

3. a). Explain the geometric representation of signals and express the energy of the signal in terms of the signal vector. → (8 M)

→ A set of orthonormal vectors are mutually orthogonal and all have unit energy. Whose inner product with itself results in 1. and with other vectors results in 0.

The essence of geometric representation of signals is to represent any set of M energy signals $s_i(t)$ as linear combinations of N orthonormal basis functions, where $N \leq M$. That is, given a set of real-valued energy signals $s_1(t), s_2(t), \ldots s_M(t)$, each of duration T seconds.

$$s_i(t) = \sum_{j=1}^{N} s_{ij}\,\phi_j(t) \quad , \quad \begin{cases} 0 \leq t \leq T \\ i = 1, 2, \ldots M \end{cases}$$

where the co-efficients

$$s_{ij} = \int_0^T s_i(t)\,\phi_j(t)\,dt \quad \begin{cases} i = 1, 2, \ldots M \\ j = 1, 2, \ldots N \end{cases}$$ — 2M

The real valued basis functions $\phi_1(t), \phi_2(t) \ldots \phi_N(t)$ form an orthonormal set, by which it means.

$$\int_0^T \phi_i(t)\,\phi_j(t)\,dt = \delta_{ij} = \begin{cases} 1 & \text{if } i = j \\ 0 & \text{if } i \neq j \end{cases}$$ — 1M

the set of co-efficients $\{s_{ij}\}_{j=1}^{N}$ may be viewed as an N-dimensional signal vector, denoted by s_i. We may state that each signal in the set $\{s_i(t)\}$ is completely determined by the signal vector.

(RN)

$$s_i = \begin{bmatrix} s_{i1} \\ s_{i2} \\ \vdots \\ s_{iN} \end{bmatrix} \quad ; i = 1, 2, \ldots M.$$

— 1M

The length of a signal vector s_i is denoted by $\|s_i\|$. The squared length of any signal vector s_i is defined as to be the inner product of s_i with itself. as shown by $\|s_i\|^2 = s_i^T s_i = \sum_{j=1}^{N} s_{ij}^2$, $i = 1, 2 \ldots M$.

— 2M

where s_{ij} is the j^{th} element of s_i.

The energy content of a signal and its representation as a vector can be given as

$$E_i = \int_0^T s_i^2(t)\, dt \quad , i = 1, 2 \ldots M.$$

$$E_i = \int_0^T \left[\sum_{j=1}^{N} s_{ij}\, \phi_j(t)\right]\left[\sum_{k=1}^{N} s_{ik}\, \phi_k(t)\right] dt$$

$$E_i = \sum_{j=1}^{N} \sum_{k=1}^{N} s_{ij}\, s_{ik} \int_0^T \phi_j(t)\, \phi_k(t)\, dt.$$

w.k.t. $\phi_j(t)$ & $\phi_k(t)$ form an orthonormal set

$$E_i = \sum_{j=1}^{N} s_{ij}^2 = \|s_i\|^2$$

— 2M

(RN)

(7)

3.b) Obtain the decision rule for maximum likelihood decoding & explain the correlation receiver → (8M)

→ The likelihood function for AWGN channel is given by

$$f_X(X/m_k) = (\pi N_0)^{-N/2} \exp\left[\frac{-1}{N_0}\sum_{j=1}^{N}(x_j - s_{kj})^2\right] \quad -(1)$$

$k = 1, 2 \ldots M$

The corresponding value of metric is the natural log of likelihood function and is given by.

$$\ln[f_X(X/m_k)] = -\frac{N_0}{2}\ln(\pi N_0) - \frac{1}{N_0}\sum_{j=1}^{N}(x_j - s_{kj})^2 \quad -(2)$$

From above eqn (2) it is seen that log-likelihood fun is maximum when the RHS of eqn (2) is maximum. Second term is negative and we get maximum when magnitude of summation term is minimum. — 1M

This leads to the decision rule as

Choose $\hat{m} = m_i$, if $\sum_{j=1}^{N}(x_j - s_{kj})^2$ is minimum for $k = i$

Further $\sum_{j=1}^{N}(x_j - s_{kj})^2 = \|X - S_k\|^2 \quad -(3)$. — 1M

which is the square of the distance between the received signal point X and signal point S_k.

If $\ln[f_X(X/m_i)]$ is maximum for $k = i$, then this indicates that X is nearer to "S_i" than other signal points

Hence decision rule can be simplified by expanding left hand side equation.

(RN)

$$\sum_{j=1}^{N}(x_j - s_{kj})^2 = \sum_{j=1}^{N} x_j^2 - 2\sum_{j=1}^{N} x_j s_{kj} + \sum_{j=1}^{N} s_{kj}^2$$

$k = 1, 2, \ldots M$

In RHS, 1st term is the energy of received signal and can be ignored.

3rd term gives the energy of the signal $s_k(t)$, and 2nd term is the inner product of received vector x and signal vector s_k.

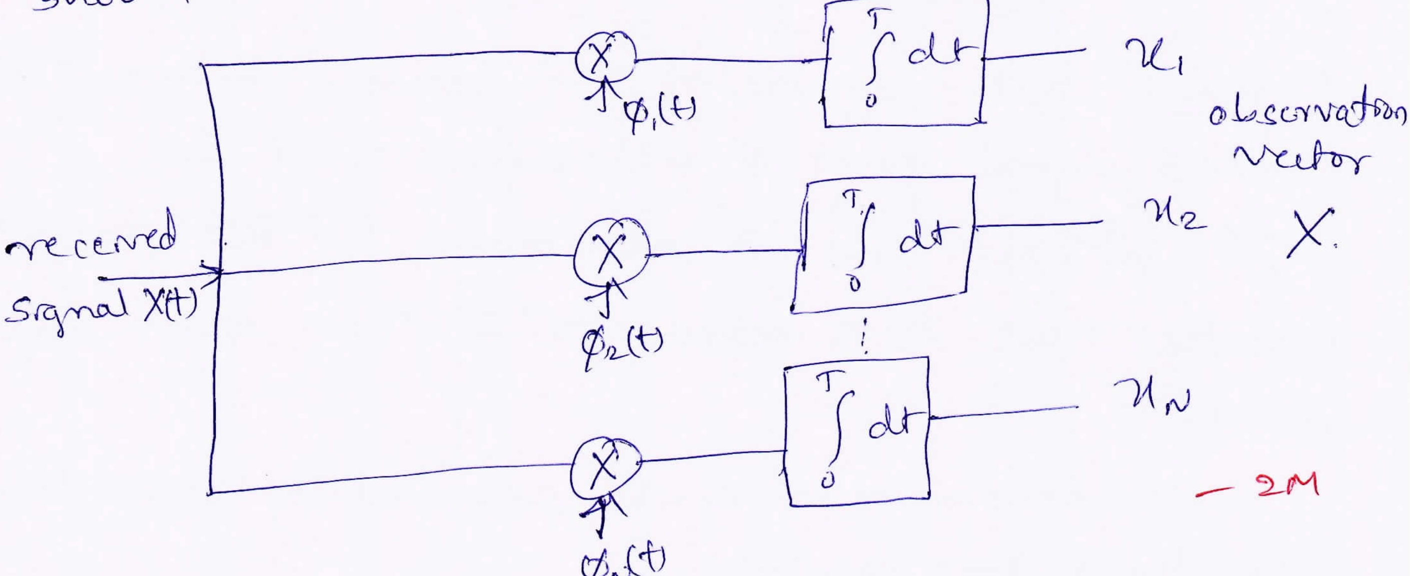
From eqn ③ X is minimum if $k=i$ & y_k is maximum

ie; $y_k = \sum_{j=1}^{N} x_j - s_{kj} - \frac{1}{2} E_k$. — 2M

Correlation Receiver

For an AWGN channel, when the transmitted signals are equally likely, the optimum receiver which minimizes average probability of error is a correlation receiver.

Correlation receiver works on maximum likelihood decision and consists of two sub-systems which are as shown below.



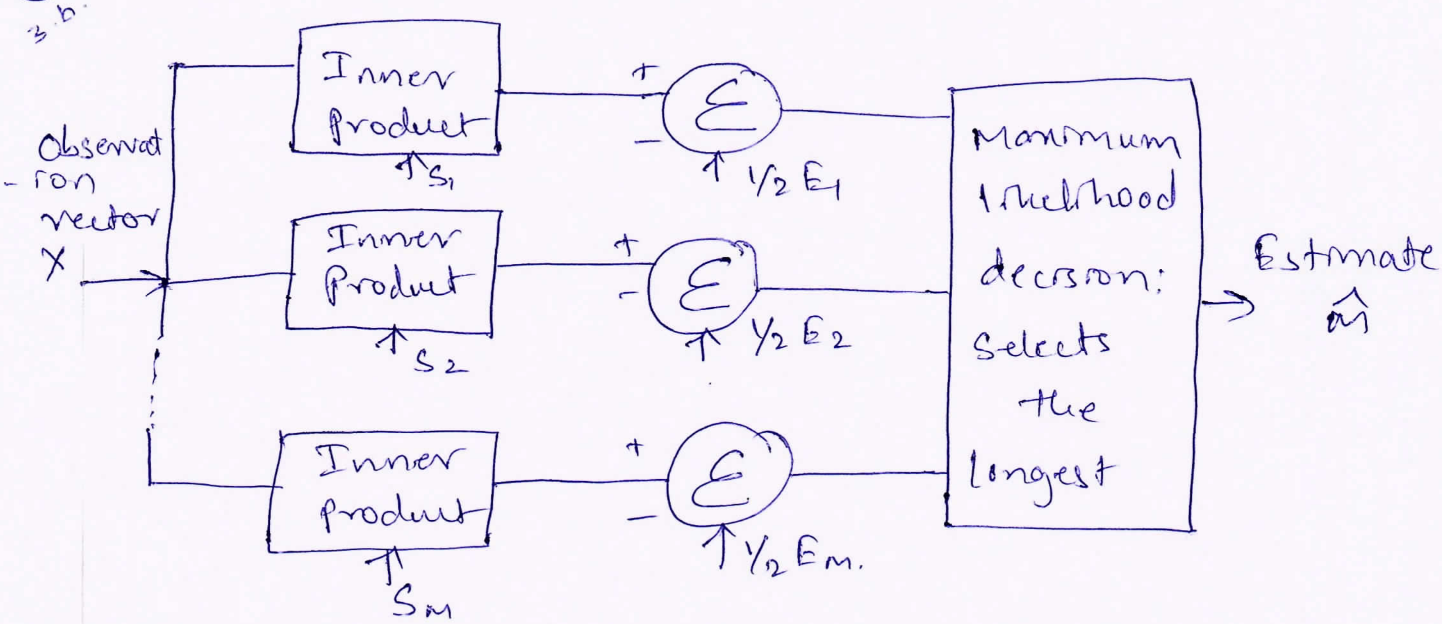
— 2M

fig ①: Bank of Correlators.

(RN)

(6)

3.b.



fig(2): Computation of $\hat{M}$

Fig(2) estimates inner product of 'x' with signal vector S_i; $i = 1, 2 \cdots M$, from which the signals half energy $1/2 E_i$ is substracted and finally maximum likelihood decision is employed to get the estimate $\hat{M}$.

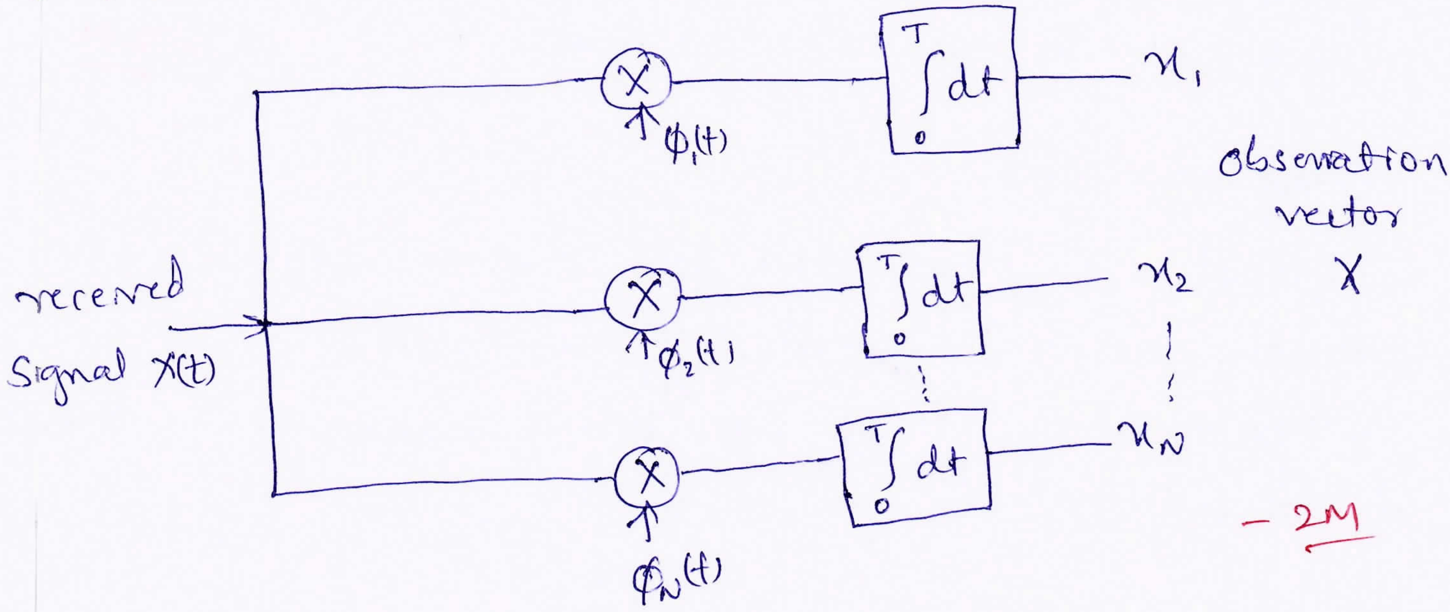
— 2M

RN

4 a). Explain the correlation receiver using product Modulator or integrator and matched filter → (8M)

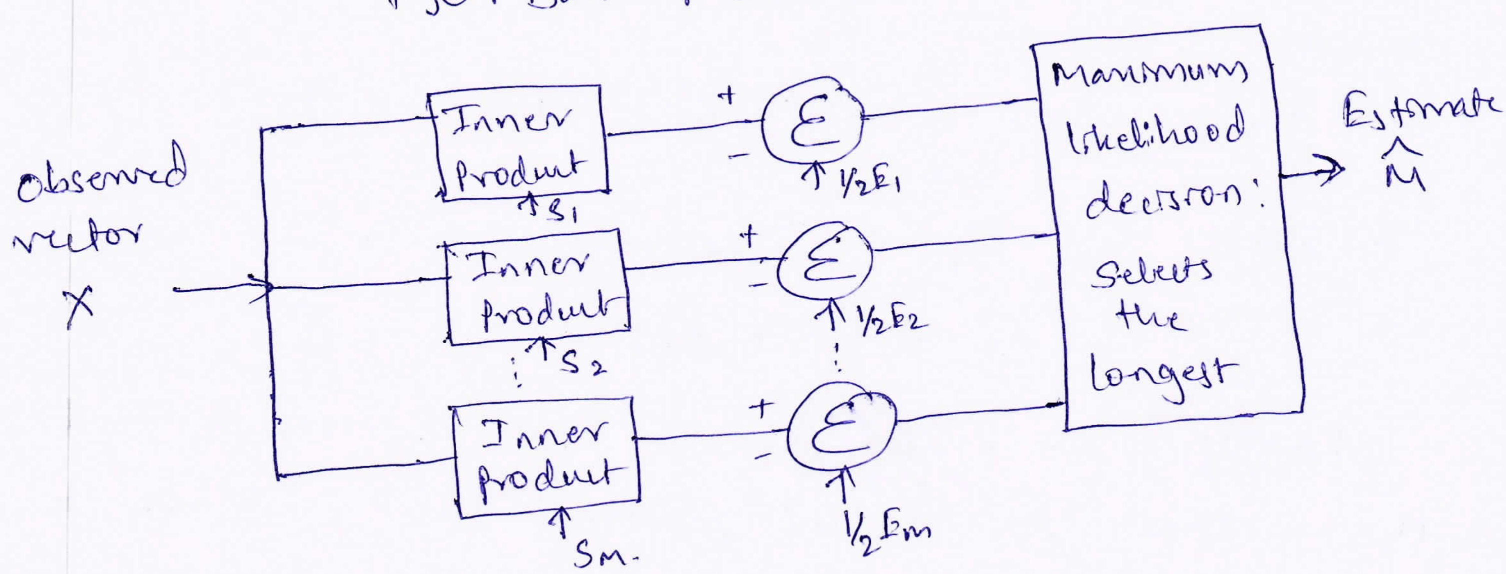
→ For an AWGN channel, when the transmitted signals are equally likely, the optimum receiver which minimizes average probability of error is a correlation receiver.

Correlation receiver works on maximum likelihood decision and consists of two sub-systems which are as shown below.



- 2M

fig①: Bank of Correlators



fig②: Computation of $\hat{M}$

Fig①, which is a detector, consists of M correlators, with a set of orthonormal basis functions $\phi_1(t)$----$\phi_N(t)$ that are generated locally. This bank of correlators operates on the received signal $x(t)$, to produce the observation vector X.

Fig② estimates inner product of 'X' with signal vector S_i, $i=1,2.--M$, from which the signals half energy $\frac{1}{2}E_i$ is substracted and finally maximum likelihood decision is employed to get the estimate $\hat{m}$.

-2M

Matched filter

Consider a linear filter with impulse response $h_i(t)$ with the received signal $x(t)$ as the i/p.

The resulting filter o/p $y_i(t)$ is given by the convolution of $x(t)$ & $h_i(t)$ ie, $y_i(t) = \int_{-\infty}^{\infty} x(\tau)\, h_i(t-\tau)\, d\tau$ —①

evaluating this integral over the duration of transmitted symbol $0 \leq t \leq T$

$$y_i(t) = \int_{t=0}^{T} x(t)\, h_i(T-t)\, dt \quad -②$$

w.k.t. the o/p of i^{th} correlator is defined by

$$x_i = \int_0^T x(t)\, \phi_i(t)\, dt \quad -③$$

For $y_i(t)$ to equal x_i, we find from eqn ② that this condition is satisfied provided that we have to choose $h_i(T-t) = \phi_i(t)$ for $0 \leq t \leq T$, $i=1,2,---M$

-2M ④

49 (10)

From eqn (4), it is possible to get the N components of the observation vector X, nearly $x_i, i = 1, 2 \ldots N$ that can be used to get the estimate $\hat{m}_i$ of the ith transmitted symbol m_i

The scheme is shown in figure below.

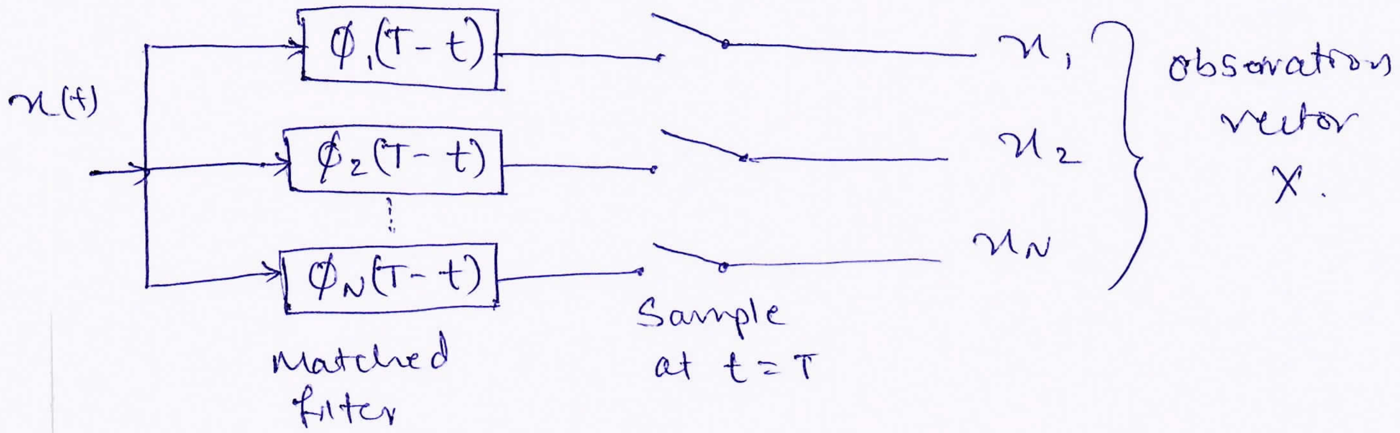


fig: Matched filter

A filter whose response is a time reversal and delayed version of its i/p signal as in eqn (4), is known as a filter matched to $\phi_i(t)$

Optimum detector using a matched filter is called Matched filter receiver.

2 M

(11)

4.b) Three signals $s_1(t)$, $s_2(t)$ & $s_3(t)$ are shown in Fig Q. 4(b). Apply Gram Schmidtt procedure to obtain an orthonormal basis for the signals. Express signals $s_1(t)$, $s_2(t)$ & $s_3(t)$ in terms of orthonormal basis functions.

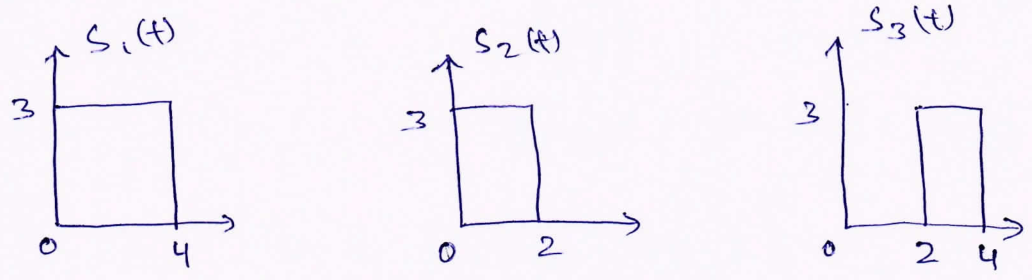


fig: Q.4(b). → (8M)

→ By observations of waveforms/signals, we can say that $s_1(t) = s_2(t) + s_3(t)$.

Therefore given set is not linearly independent.

$$\phi_i(t) = \frac{s_i(t)}{\sqrt{E_i}}, \quad E_i = \int_0^T s_i^2(t)\,dt$$

and w.k.t $$\phi_i(t) = \frac{g_i(t)}{\sqrt{\int_0^T g_i^2(t)\,dt}}$$

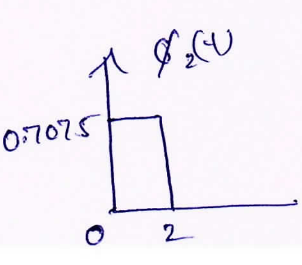
where $$g_i(t) = s_i(t) - \sum_{j=1}^{i-1} s_{ij}\,\phi_j(t)$$

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

i) $$\phi_2(t) = \frac{s_2(t)}{\sqrt{E_2}}, \quad E_2 = \int_0^T s_2^2(t)\,dt$$

$$E_2 = \int_0^2 3^2 = 18$$

$$\phi_2(t) = \frac{3}{\sqrt{18}} = 0.7075$$



— 2M

2) $\phi_3(t) = \dfrac{g_3(t)}{\sqrt{\int_0^T g_3^3(t)\, dt}}$

$$g_3(t) = S_3(t) - \sum_{j=1}^{i-1} S_{ij}\, \phi_j(t)$$

$$g_3(t) = S_3(t) - S_{32}\phi_2(t)$$

$$S_{32} = \int_0^2 S_3(t)\, \phi_2(t)\, dt = 0$$

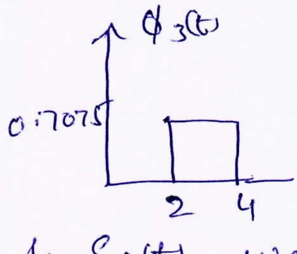
$$\therefore \quad g_3(t) = S_3(t).$$

$$E_3 = \int_2^4 S_3^2(t)\, dt = \int_2^4 3^2\, dt$$

$$E_3 = 18$$

$$\therefore \quad \phi_3(t) = \frac{g_3(t)}{\sqrt{E_3}} = \frac{3}{\sqrt{18}} = 0.7075$$

— 2M



Since $S_1(t)$ is sum of $S_2(t)$ & $S_3(t)$. we donot need to calculate $\phi_1(t)$ related to $S_1(t)$. $\therefore$ $\phi_1(t) = 0$.

$$\therefore \quad S_1(t) = (4.24)\,\phi_2(t) + (4.24)\,\phi_3(t)$$

$$S_2(t) = (4.24)\,\phi_2(t)$$

$$S_3(t) = (4.24)\,\phi_3(t).$$

— 2M

Module - 3

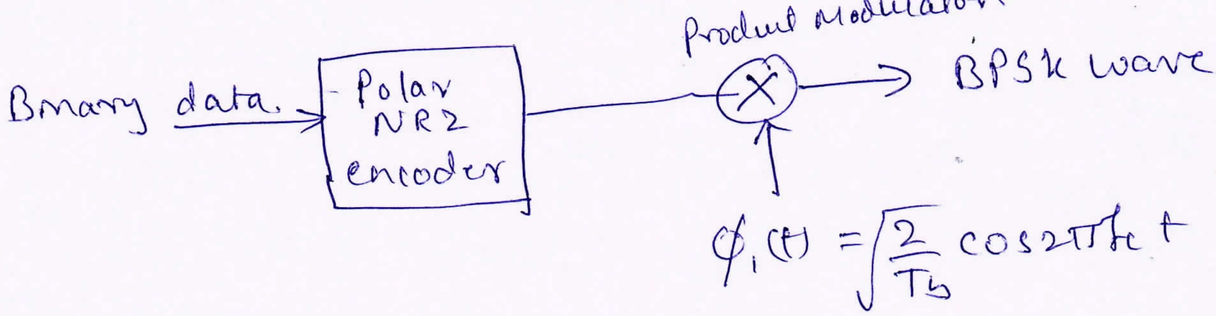
5 a) With necessary diagrams, explain the generation and reception of BPSK signal → (10M)

→ In BPSK, the signals $S_1(t)$ and $S_2(t)$ that are used to represent binary symbols 1 and 0 respectively are defined as.

$$S_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_c t) \qquad 0 \le t \le T_b.$$

$$S_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_c t + \pi) \qquad 0 \le t \le T_b.$$ — 1M

Binary PSK can be generated using product modulator as shown in figure below.



— 2M

fig ①. BPSK generator

The input binary sequence is in NRZ - polar form with symbol 1 represented by $+\sqrt{E_b}$ and symbol 0 represented by $-\sqrt{E_b}$ — 1M

The other input to the product modulator is the carrier $\phi_1(t)$

The output of the product modulator is the required binary PSK waveform.

Decision Rule for Receiver.

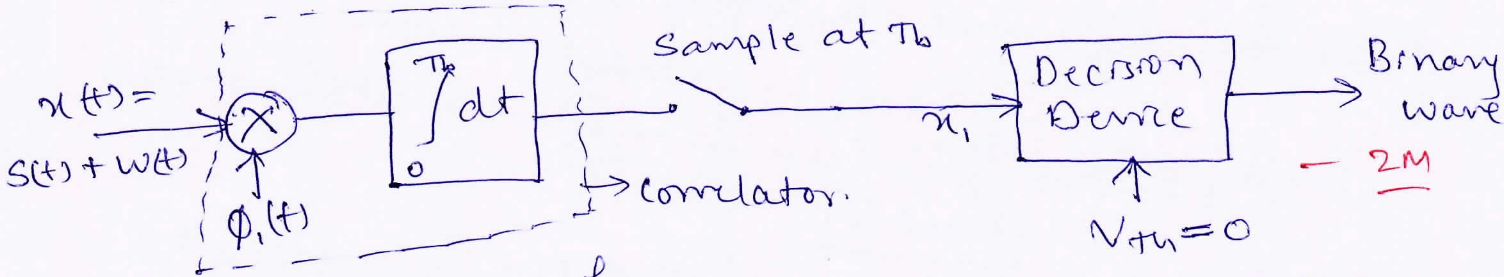
choose signal $s_1(t)$ or binary symbol 1, if the received signal point x_1 falls in the region Z_1 of Signal constellation diagram. That is, choose symbol 1 if $x_1 > 0$, and choose $S_2(t)$ or Symbol '0' if $x_1 < 0$. - 1M

But if $x_1 = 0$, the decision is arbitrary.

Detection of Binary PSK

If phase reference is available at the receiver, the detection process is called as coherent detection.

The coherent BPSK receiver is shown in fig(2).



- 2M

fig(2): Coherent BPSK receiver

Let $x(t)$ be received BPSK signal corrupted by AWGN

Also, let $\phi_1(t)$ be the local coherent reference signal which is synchronized with respect to the freq and phase of the carrier used at the transmitter.

The correlator output is sampled every Tb seconds, then the sample value is compared with V_{Th} (threshold) = 0 volts

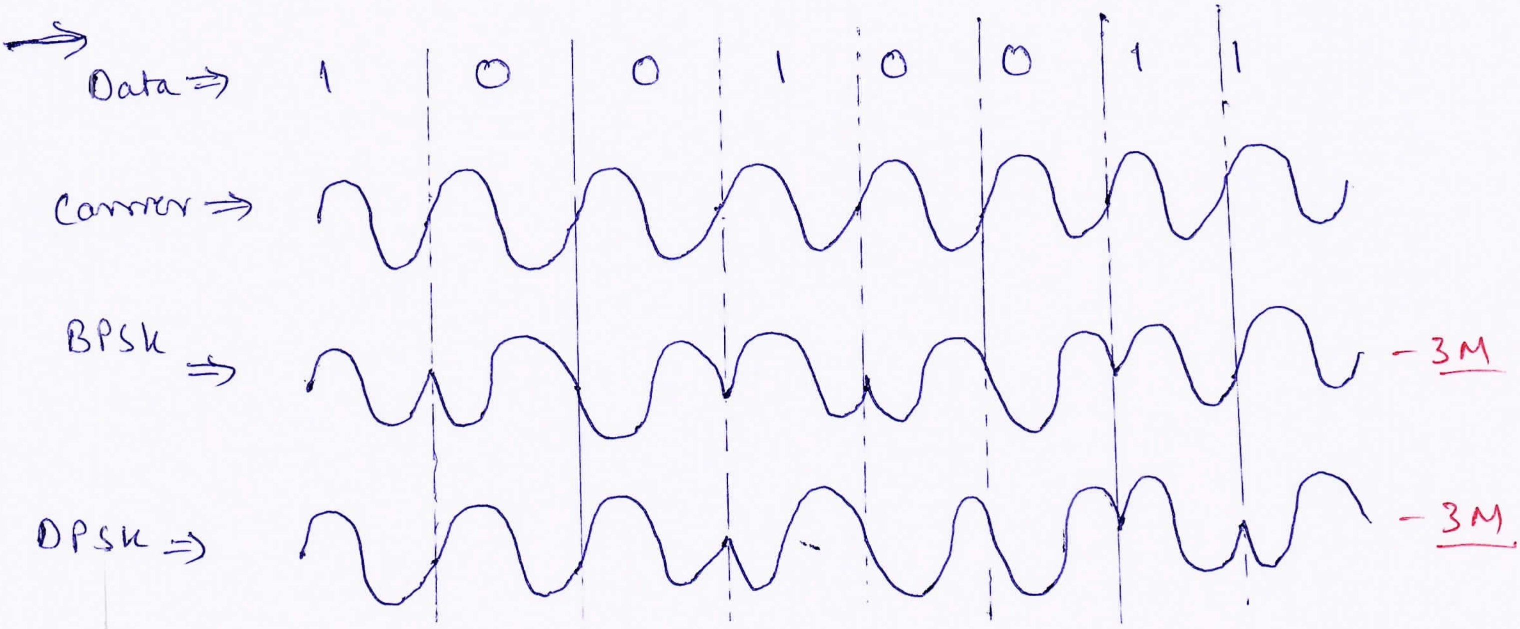
If $x_1 > 0$ then Rx decides in favour of '1'

$x_1 < 0$ then Rx decides in favour of '0'.

$x_1 = 0$ then the decision is arbitrary. - 3M

(13)

5 b). Given the binary data 10010011, draw the BPSK and DPSK waveforms. → 6 M



RN

(24)

6. a) Derive the expression for error probability of BFSK. → (8 M)

→ we receive signals at receiver

$x_1(t) = s_1(t) + w_1(t)$ and

$x_2(t) = s_2(t) + w_2(t)$

let us assume here y as new variable, such that

$$y = x_1 - x_2$$

The conditional mean value of y depends on which binary symbol was transmitted.

For symbol '1' $E[Y/1] = E[X_1/1] - E[X_2/1]$

$= E[\sqrt{E_b}] - E[0]$

$= \sqrt{E_b}.$

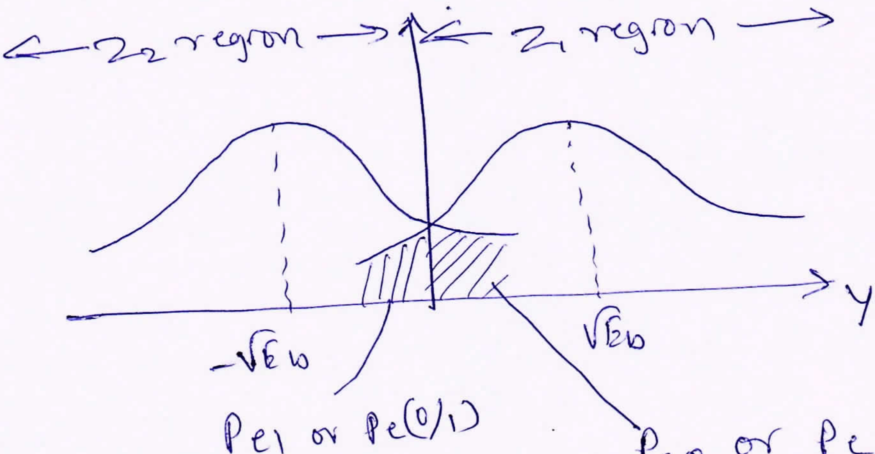
For symbol '0' $E[Y/0] = E[X_1/0] - E[X_2/0]$

$= -\sqrt{E_b}$ — 2M

Variance of y is independent of which symbol was sent

$\therefore Var[Y] = Var[X_1] + Var[X_2]$

$$= \frac{N_0}{2} + \frac{N_0}{2} = N_0$$



— 2M

The observation vector x has two elements x_1 & x_2 given by $x_1 = \int_0^{T_b} x(t)\,\phi_1(t)\,dt$ and

$$x_2 = \int_0^{T_b} x(t)\, \phi_2(t)\, dt$$

w.k.t gaussian function is given by

$$f(x) = \frac{1}{\sqrt{2\pi\sigma^2}} \int \exp\left[\frac{-(x-\mu_x)^2}{2\sigma^2}\right] dx$$

$$\therefore P_e(1/0) = \frac{1}{\sqrt{2\pi N_0}} \int_0^{\infty} \exp\left[\frac{-(y-(-\sqrt{E_b}))^2}{2N_0}\right] dy.$$

$$= \frac{1}{\sqrt{2\pi N_0}} \int_0^{\infty} \exp\left[\frac{-(y+\sqrt{E_b})^2}{2N_0}\right] dy. \quad - 2M$$

w.k.t. $\frac{1}{2} \operatorname{erfc}(u) = \frac{1}{\sqrt{\pi}} \int_u^{\infty} \exp(-z)^2\, dz$ — ①

$$z = \frac{y + \sqrt{E_b}}{\sqrt{2N_0}}$$

for $y_0 = 0$, $z = \sqrt{\frac{E_b}{2N_0}}$ $\quad y = \infty, \ z = \infty.$

$$dy = \sqrt{2N_0}\, .dz.$$

$$\therefore P_e(1/0) = P_{e0} = \frac{1}{\sqrt{\pi}} \int_{\sqrt{\frac{E_b}{2N_0}}}^{\infty} \exp(-z)^2\, dz. \quad — ②.$$

Comparing ① & ②.

$$P_{e0} = \frac{1}{2} \operatorname{erfc} \sqrt{\frac{E_b}{2N_0}}$$

Similarly $P_{e1} = \frac{1}{2} \operatorname{erfc} \sqrt{\frac{E_b}{2N_0}}$

$\therefore$ Total probability of error $= P_e = \frac{1}{2}\left[P_{e0} + P_{e1}\right]$

$$P_e = \frac{1}{2} \operatorname{erfc} \sqrt{\frac{E_b}{2N_0}} \quad - 2M$$

6. b) With block diagram explain generation and detection of DPSK. → (8M)

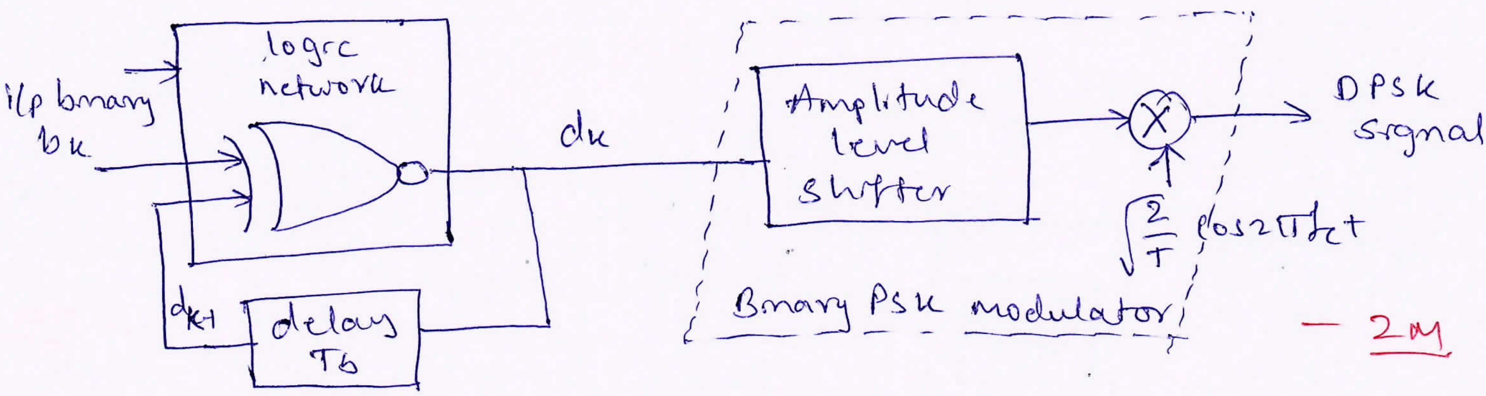
→ In DPSK two basic operations are carried out
1) Differential encoding of input binary sequence
2) phase shift keying of encoded sequence

The differential encoded sequence d_k is generated by using logic equation. $d_k = \overline{b_k \oplus d_{k-1}}$ — ①

$\therefore d_k = b_k d_{k-1} + \bar{b}_k \bar{d}_{k-1}$ — ②

$\therefore$ from eqn ② we conclude that
if $b_k = d_{k-1}$; we get $d_k = 1$
$b_k \neq d_{k-1}$; we get $d_k = 0$. — 1M

Generation of DPSK signal



— 2M

fig ①: DPSK transmitter.

Differential encoding starts with reference bit
1) If the new bit at the i/p is 1, leave the encoded symbol unchanged w.r.t. the current bit
2) If, on the other hand, the input is 0, change the encoded symbol w.r.t. the current bit.

The differentially encoded sequence, denoted by $\{d_k\}$ is used to shift sinusoidal carrier phase by $0°$ and $180°$, representing '1' and '0' respectively.

Thus ① To send symbol '1', the phase of DPSK signal remains unchanged

② To send symbol '0', the phase of DPSK signal is shifted by $180°$.

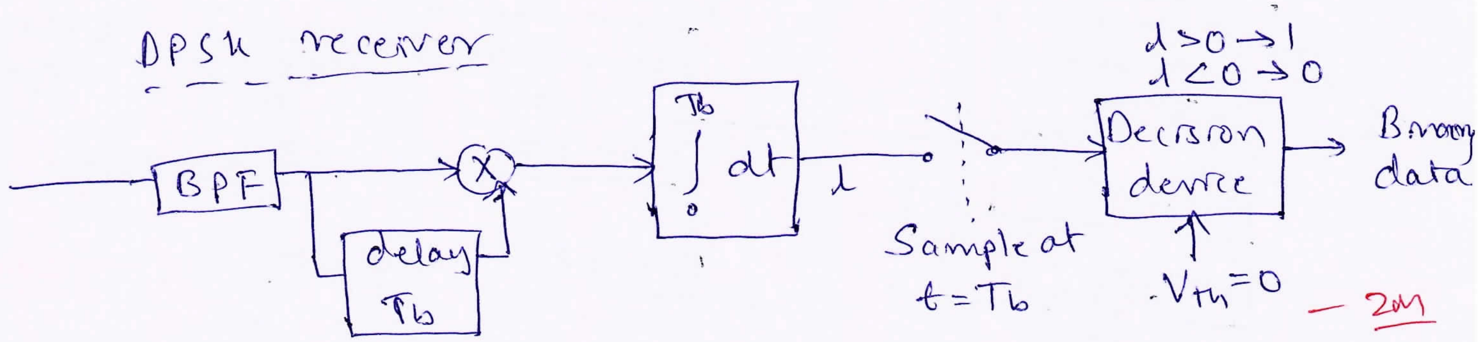
The amplitude level shifter produces amplitudes $+\sqrt{E_b}$ if $d_k = 1$ and $-\sqrt{E_b}$ for $d_k = 0$.

The product modulator o/p is

$$\sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_c t \quad \text{for } d_k = 1 \text{ and}$$

$$\sqrt{-\frac{2E_b}{T_b}} \cos 2\pi f_c t \quad \text{for } d_k = 0.$$ — 2M.

DPSK receiver



— 2M

fig ②: DPSK receiver

The receiver is having storage capability, to measure relative phase difference between waveforms received during two ~~corre~~ successive bit intervals.

The BPF passes only the spectrum of the DPSK signal

If $d_k = d_{k-1}$, then integrator o/p is positive and thus $l > 0$. and decision is taken in favour of symbol '1'

6.b (16)

If $d_k \neq d_{k-1}$, then the integrator o/p is negative, that is $l < 0$, then the decision is taken in favour of symbol '0'.

This leads to the decision rules for the k^{th} bit as,

If $l > 0$, $b_k = 1$

$l < 0$, $b_k = 0$.

— 3M.

(RN)

Module - 4

7a) What is ISI? Obtain the expression of output of a filter with intersymbol interference. → (8M)

→ Intersymbol interference is a form of distortion of a signal in which one symbol interferes with subsequent symbols.

When a pulse of short duration T_b is transmitted on band limited channel, freq components of the pulse are differentially attenuated due to the freq response of channel causing dispersion of pulse over the interval greater than T_b.

ISI causes a degradation in the performance of the digital comm. system.

By appropriately designing the transmitting & receiving filters, we can satisfy the condition $x_n = 0$ for $n \neq 0$, so that the ISI term vanishes. — 2M.

Consider the problem of filter design when the channel distorts the transmitted signal.

W.k.t $\quad H(f) = C(f)\, G_T(f)$

$C(f)$ is the freq. response of channel

$G_T(f)$ is the ~~freq.~~ response of transmitter filter.

The condition for distortion-free transmission is that the freq. response $C(f)$ of the channel must have a constant magnitude and a linear phase over the BW of the transmitting signal.

$$C(f) = \begin{cases} C_0\, e^{-j2\pi f t_0} & |f| \leq W \\ 0 & |f| > W. \end{cases}$$ — 3M.

C_0 - Constant gain factor, set to unity

Thus under the condition that the channel is distortion-free and BW of $g_T(t)$ is limited to W, we have $H(f) = G_T(f)$.

As a result the matched filter at the receiver has a freq. response $G_R(f) = G_T^*(f)$ and its o/p at the periodic sampling times $t = mT$ has the form.

$$y(mT) = x(0)\, a_m + \sum_{n \neq m} a_n\, x(mT - nT) + w(mT) \quad — ②$$

or simply

$$y_m = x_0 a_m + \sum_{n \neq m} a_n\, x_{n-m} + w_m \quad — ③ \quad - 2M$$

where $x(t) = g_T(t) * g_R(t)$ and $w(t)$ is the o/p response of the matched filter to the i/p AWGN process $n(t)$.

The middle term $\sum_{n \neq m} a_n x_{m-n}$ represents the ISI. — 1M

7.b). Explain the Nyquist criterion for distortionless baseband binary transmission and obtain the ideal solution for zero ISI. → 8M

→. The pulse shaping function $p(t)$ with FT given by $P(f)$. that satisfies

$$\sum_{n=-\infty}^{\infty} P(f - nR_b) = T_b$$

has

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

where $p(0)$ is set to unity in accordance with normalization. — 1M

This condition is known as Nyquist criterion for zero-ISI.

(Raghavendra. N)

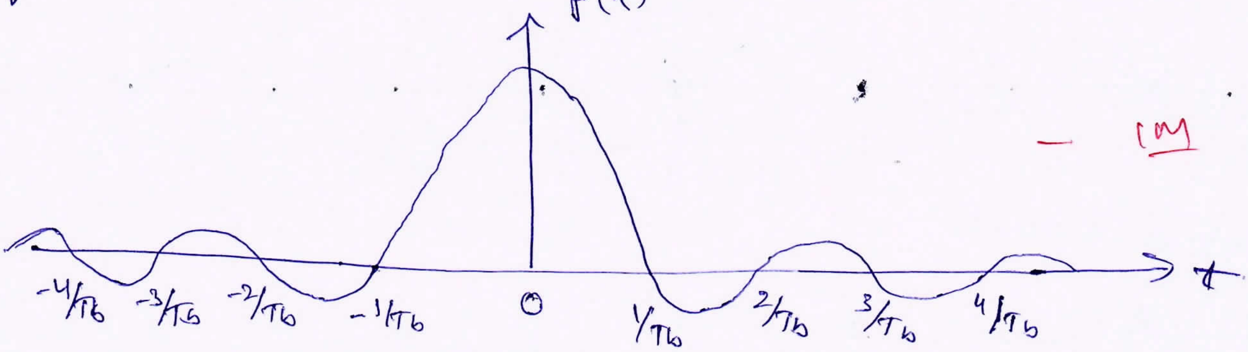
7.B (18)

Ideal solution or Nyquist solution for zero ISI.

The ISI can be minimized by controlling $p(t)$ in time domain and $P(f)$ in freq. domain.

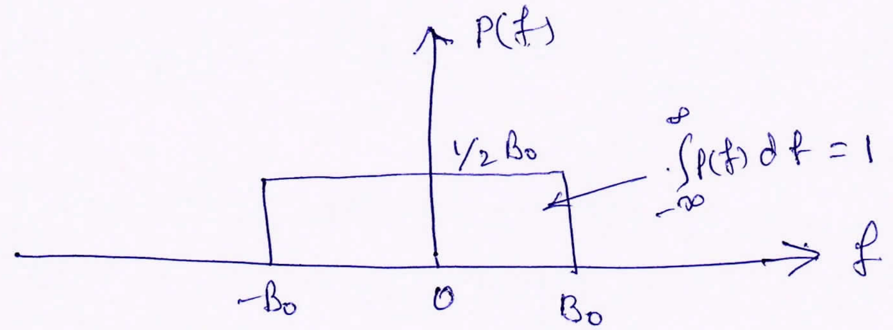
One of the functions that gives zero-ISI is

$p(t) = \text{Sinc}(2B_0 t)$ shown in fig① below and the spectrum shown in fig②, where $B_0 = \frac{1}{2T_b}$ is called Nyquist BW.



— 1M

fig① Impulse response of ideal filter.



— 1M

fig② freq. response of ideal filter.

Nyquist BW is defined as the minimum transmission BW for zero ISI.

The FT of $p(t)$ gives $$P(f) = \begin{cases} 1/2B_0 & |f| < B_0 \\ 0 & |f| > B_0 \end{cases} \quad -①$$

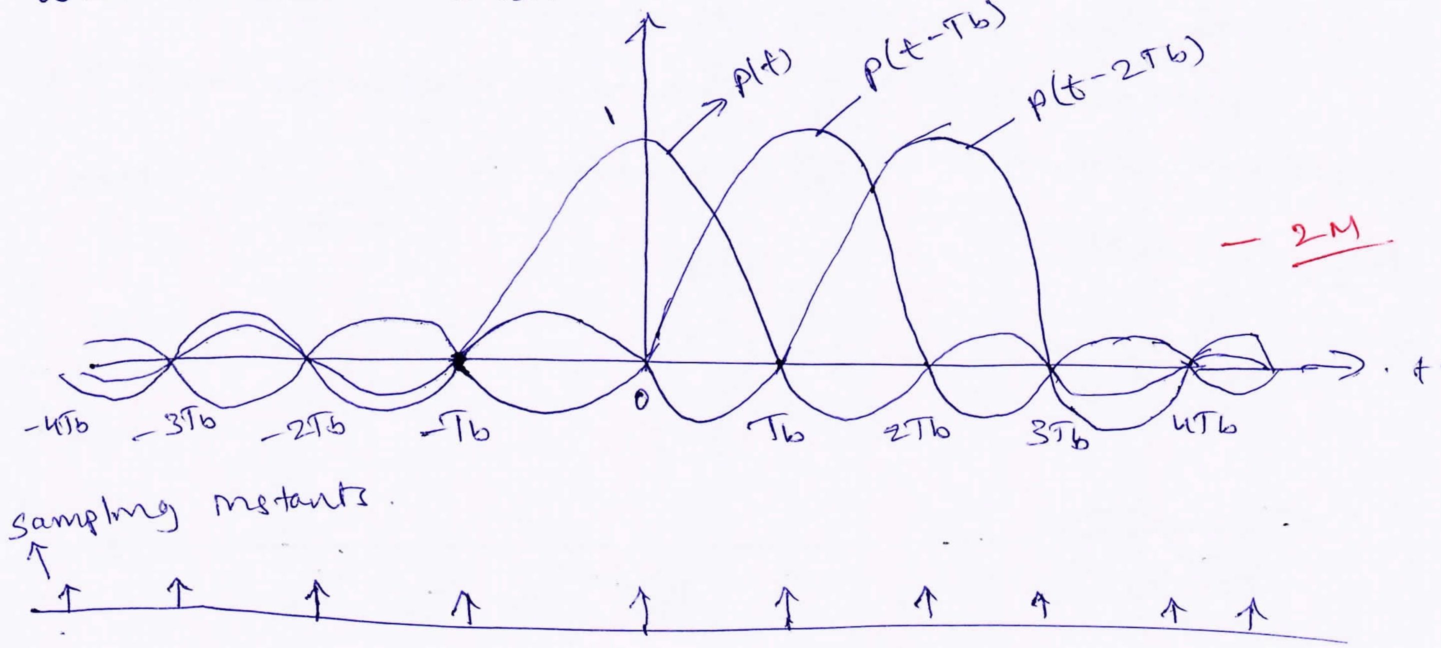
Since $p(t)$ is a sinc function it goes through zero at integer multiples of T_b.

— 1M

Thus if $T_b = \frac{1}{2B_0}$, it is clear that

RN

$p(t - kT_b) = \text{Sinc}[2B_o(t - kT_b)]$ for integer values of k will appear as shown in fig ③. Also fig ③ implies that if $y(t)$ is sampled at instants of time $t = 0, \pm T_b, \pm 2T_b$ ---- will have zero ISI.



— 2M

fig ③ : Demonstration of sampling instants

However, the $\frac{\sin \pi x}{\pi x}$ type of pulse shaping function has two difficulties.

① The freq. response $P(f)$ has to be flat over the range $-B_o < f < B_o$ and zero elsewhere. This is physically unrealizable.

② The synchronization of the clock in the receiver sampling ckt has to be almost perfect, since $\frac{\sin \pi x}{\pi x}$ decays only as $1/x$ and is zero in adjacent time slots only when it is at the exactly correct sampling time.

Thus inaccurate synchronization cause ISI.

W.k.t. $y(t) = \sum_{k=-\infty}^{\infty} \mu A_k\, p(t - kT_b)$

$p(t) = \text{Sinc}(2B_o t)$ we get — 1M

$y(t) = \sum_{k=-\infty}^{\infty} \mu A_k \,\text{Sinc}(2B_o t - 2B_o kT_b)$

(RN)

Since $B_0 = 1/2T_b$ or $B_0 2T_b = 1$

$$y(t) = \sum_{k=-\infty}^{\infty} \mu A_k \operatorname{sinc}[2B_0 t - k]$$

letting $t = \Delta t$, where Δt is timing error.

$$y(\Delta t) = \sum_{k=-\infty}^{\infty} \mu A_k \operatorname{sinc}[2B_0 \Delta t - k]$$

$$= \sum_{k=-\infty}^{\infty} \frac{\mu A_k}{\pi[2B_0\Delta t - k]} [\sin(2B_0\pi\Delta t)\cos\pi k - \cos(2B_0\pi\Delta t)\sin\pi k]$$

$$= \sum_{k=-\infty}^{\infty} \frac{\mu A_k}{\pi(2B_0\Delta t - k)} (-1)^k \sin(2B_0\pi\Delta t)$$

$$y(\Delta t) = \mu A_0 \operatorname{sinc}(2B_0\Delta t) + \frac{\mu}{\pi} \sum_{\substack{k=-\infty \\ \neq 0}}^{\infty} \frac{A_k(-1)^k \sin(2B_0\pi\Delta t)}{(2B_0\Delta t - k)}.$$ — 1M.

1st term gives desired symbol

2nd term represents ISI caused by timing error.

8 a) Draw and explain the time-domain and freq-domain of duobinary and modified duobinary signal. → (8M)

→ Based on correlative coding, to have controlled ISI & BW, Duobinary coding is used.

Time domain condition for Zero ISI

$$P(nT_b) = \begin{cases} 1, & n = 0 \\ 0, & n \neq 0 \end{cases}$$

Here in duobinary this condition is now relaxed to

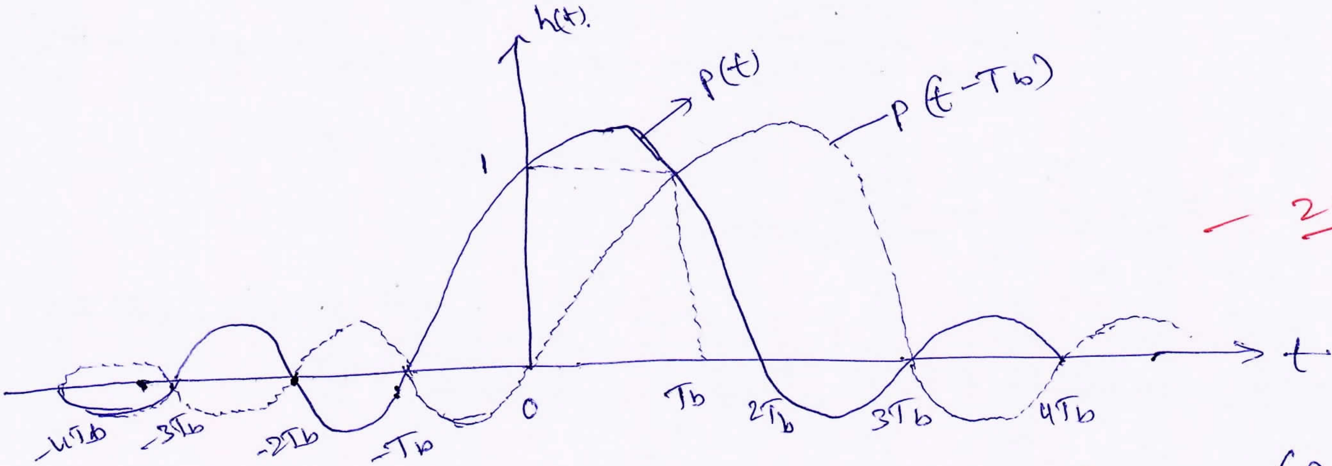
$$P(nT_b) = \begin{cases} 1, & n = 0, 1 \\ 0, & \text{for all other } n \end{cases}$$ — 1M

Allowing ISI between successive symbols, but not all

(RN)

In duobinary coding some controlled amount of ISI is introduced into the data stream to be transmitted at fixed number of adjacent sampling instants rather than eliminating it.

Considering Duobinary filter whose impulse response $h(t)$ has a shape as shown below.



— 2M

fig(1): Demonstration of controlled ISI. (Duobinary).

The overall transfer function of duobinary filter is

$$H(f) = H_c(f) \cdot H_I(f)$$

where $H_c(f)$ - freq. response of ideal channel filter

$H_I(f)$ - Freq. response of digital filter.

$$H_I(f) = 1 + e^{-j2\pi f T_b}.$$

$$H_I(f) = 2\, e^{-j\pi f T_b} \cos(\pi f T_b)$$

w.k.t $H_c(f) = \begin{cases} 1 & ;\ f \le W \\ 0 & ;\ \text{elsewhere} \end{cases}$

$$\therefore\ H(f) = H_c(f) \cdot H_I(f) = 2 H_c(f) \cos(\pi f T_b) e^{-j\pi f T_b}$$

$$= \begin{cases} 2\, e^{-j\pi f T_b} \cos \pi f T_b & ;\ |f| \le W \\ 0 & ;\ \text{elsewhere} \end{cases}$$

and magnitude response.

$$|H(f)| = 2\, |\cos(\pi f T_b)|$$

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(20)

put $f=0$ $|H(0)| = 2(1) = 2$

at $f = Tb/2$, i.e., the Nyquist BW,

$W = 1/2Tb = Rb/2$.

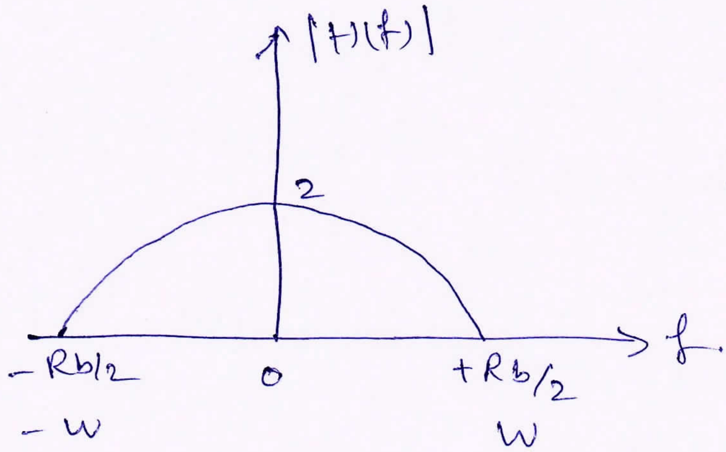
$$|H(Rb/2)| = 2\cos\left(\pi . \frac{Rb}{2} . Tb\right)$$

$$= 2\cos(\pi/2) = 0$$

and at $f = -Rb/2$

$$|H(-Rb/2)| = 2\cos\left(-\pi \frac{Rb}{2} . Tb\right)$$ — 1M

$$|H(-Rb/2)| = 0.$$



fig(2): freq. response of duobinary — 1M

Impulse response of duobinary filter is obtained by taking FT of H(f)

$$h(t) = \int_{-\infty}^{\infty} H(f) e^{j2\pi ft} df.$$

$$= \int_{-1/2Tb}^{1/2Tb} 2\cos(\pi f Tb) e^{-j\pi f Tb} \times e^{j2\pi ft} df$$

$$= \frac{\sin\left(\frac{\pi t}{Tb}\right)}{\left(\frac{\pi t}{Tb}\right)} + \frac{\sin\left[\frac{\pi(t-Tb)}{Tb}\right]}{\left[\frac{\pi(t-Tb)}{Tb}\right]}$$

$$h(t) = \frac{\sin\left(\frac{\pi t}{Tb}\right)}{\left(\frac{\pi t}{Tb}\right)} - \frac{\sin\left(\frac{\pi t}{Tb}\right)}{\left[\frac{\pi(t-Tb)}{Tb}\right]} \quad \text{— (1)}$$

1^{st} term in eqn ① represents sinc function at $t=0$

2^{nd} term in eqn ① represents sinc function at $t = T_b$.

The condition for controlled ISI in modified duobinary filter is

$$P(nT_b) = h(nT_b) = \begin{cases} 1 & , n=0 \\ -1 & , n=2 \\ 0 & , \text{elsewhere} \end{cases}$$

w.k.t. $H(f) = H_I(f) \cdot H_c(f)$

$$H_I(f) = 1 - e^{-j2\pi f(2T_b)}$$
$$= 1 - e^{-j4\pi f T_b}.$$
$$= 2j e^{-j2\pi f T_b} \cdot \sin(2\pi f T_b)$$

$$\therefore H(f) = \begin{cases} 2j e^{-j2\pi f T_b} \cdot \sin(2\pi f T_b) & ; |f| \leq R_b/2 \\ 0 & ; \text{else} \end{cases}$$

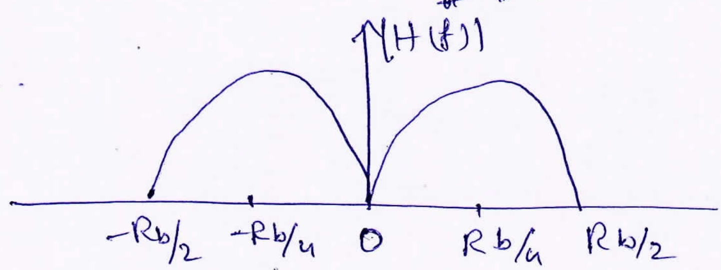
Magnitude response is $|H(f)| = \begin{cases} 2|\sin(2\pi f T_b)| \\ 0 \end{cases}$

at $f=0$ $\quad \sin(0) = 0 \quad \therefore |H(f)| = 0$ [$\because$ no dc component]

at $f = R_b/2$ $\quad H(R_b/2) = 2\sin(2\pi R_b/2 \cdot T_b)$ — 1M
$$= 0$$

at $f = -R_b/2$ $\quad H(-R_b/2) = 2\sin(2\pi(-R_b/2) \cdot T_b) = 0$

$\therefore$



— 1M

fig ③ : Freq. response of Modified duobinary filter.

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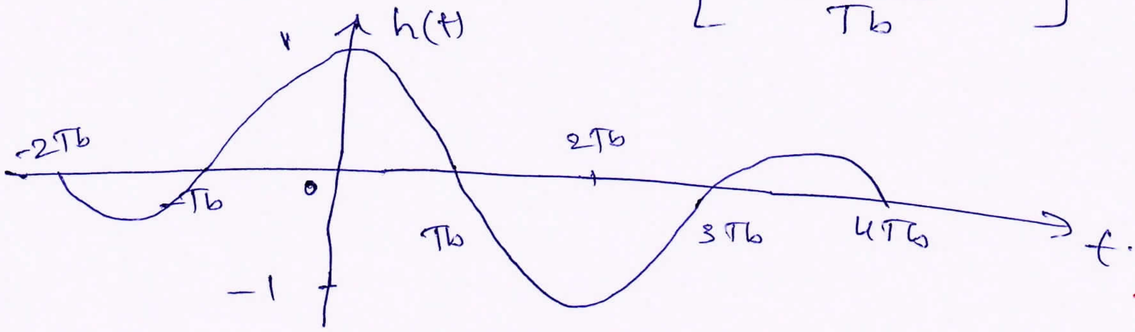
(21)

Impulse response $h(t)$ is obtained by.

$$h(t) = \int_{-\infty}^{\infty} H(f)\, e^{j2\pi f t}\, df$$

$$= \int_{-1/2T_b}^{1/2T_b} 2j \sin(2\pi f T_b)\, e^{-j2\pi f T_b} \cdot e^{j2\pi f t}\, df.$$

$$h(t) = \frac{\sin\left(\frac{\pi t}{T_b}\right)}{\left(\frac{\pi t}{T_b}\right)} - \frac{\sin\left[\frac{\pi(t-2T_b)}{T_b}\right]}{\left[\frac{\pi(t-2T_b)}{T_b}\right]}$$



— 1M.

fig(4): Time domain characteristics of modified duobinary signalling.

8P- (28)

(12)

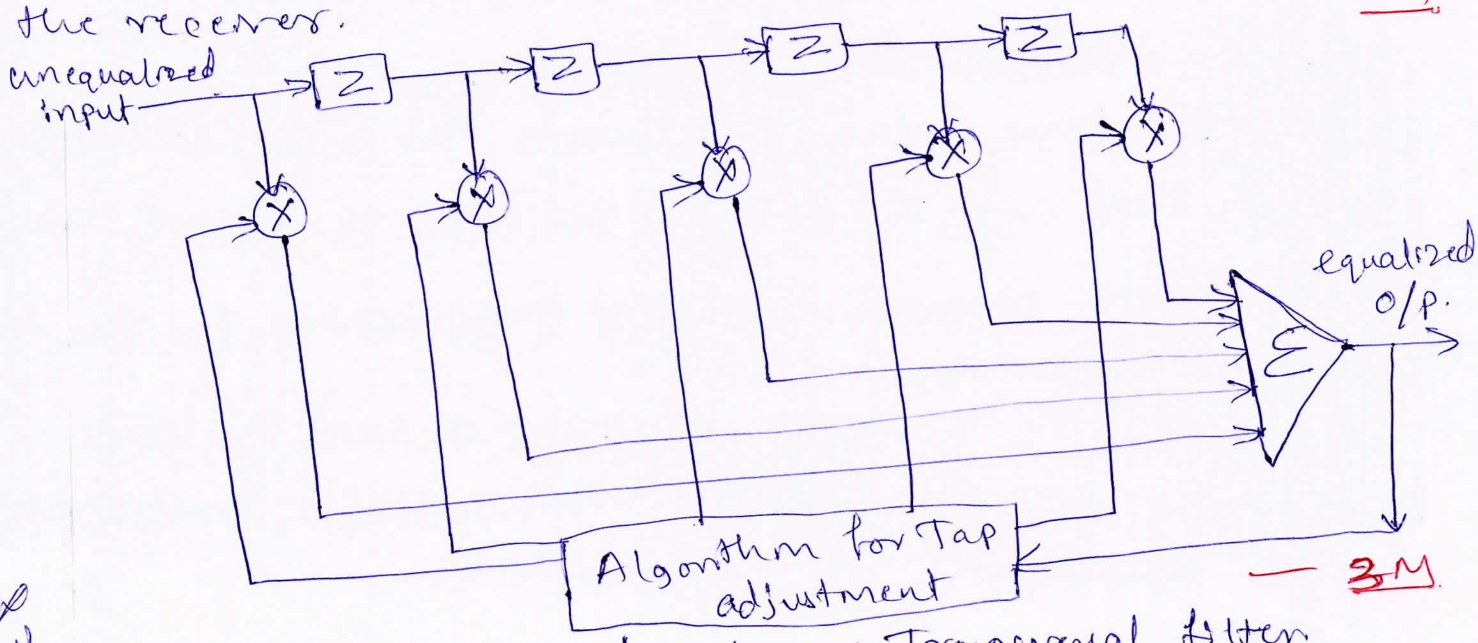
30

8 b) what is channel equalization? with a neat diagram explain the concept of equalization using a linear transversal filter. —(8 M).

→ In digital communication systems, data transmitted from the transmitter to the receiver over a channel is distorted by the addition of some random noise in the channel. If the freq. response of the channel is not flat and its phase response is not linear the signal is distorted. To make the amplitude response of the channel flat the channel equalization technique is used. The device used at the receiver is called as equalizer. It is also called as Inverse channel or Inverse filter.

whatever the effect of channel introduces on the signal, we can nullify the effect using equalizer.

To ~~po~~ compensate the effect of transmitting filter, channel and receiving filter we can use equalizer in the receiver. — 2 M.



— 3 M.

Fig: Linear Transversal filter

(RN)

we know that in real channels, the ISI is limited to a finite no. of samples.

As a result, in practice for example, the channel equalizer is approximated by a finite duration impulse response (FIR) filter or transversal filter, with adjustable tap co-efficients $\{C_n\}$. —1M

The time delay τ between adjacent taps may be selected as large as T, the symbol interval in which case the FIR equalizer is called a symbol spaced equalizer.

In this case the input to the equalizer is the sampled sequence given by

$$y_m = x_0 a_m + \sum_{\substack{n=-\infty \\ n \neq m}}^{\infty} a_n x_{m-n} + w_m$$

—1M

However, we note that when $1/T < 2W$, frequencies in the received signal that are above the folding freq. $1/T$ are aliased into freq. below $1/T$

In this case the equalizer compensates for the aliased channel-distorted signal.

when the time delay τ between the adjacent taps is such that $\frac{1}{\tau} \geq 2W \geq \frac{1}{T}$, no aliasing occurs. hence the inverse channel equalizer compensates for the true channel distortion

since $\tau < T$, channel equalizer is said to have fractionally spaced taps, and it is called a fractionally spaced equalizer.

(KRS)

8b
(23) In practice z is often selected as $z = \frac{T}{2}$, In 27
this case, the sampling rate at the o/p of the
filter $G_R(t)$ is $\frac{2}{T}$.

— 1M

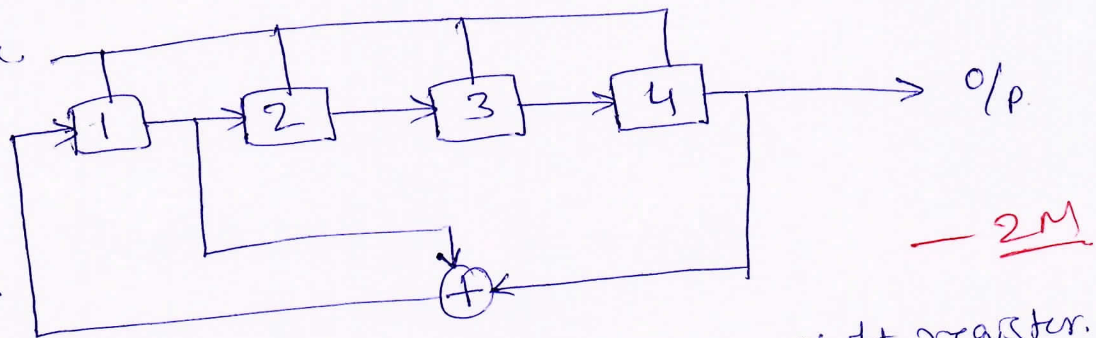
RN

Q1 (24)

27

9 a) Draw the 4 stage linear feedback shift register with 1st & 4th stage is connected to modulo-2 adder. Output of Modulo-2 is connected to 1st stage input. Find the output PN sequence and obtain the autocorrelation Sequence. —(6M).

→ Let the initial condition of shift registers as 1 0 0 0



— 2M

Fig: 4 stage linear Feedback shift register.

Clk	$S_1 \oplus S_4$	S_2	S_3	S_4	O/p
1	1	0	0	0	0 ← initial state.
1	1	1	0	0	0
1	1	1	1	0	0
1	1	1	1	1	1
1	0	1	1	1	1
1	1	0	1	1	1
1	0	1	0	1	1
1	1	0	1	0	0
1	1	1	0	1	0
1	0	1	1	0	0
1	0	0	1	1	1
1	1	0	0	1	1
1	0	1	0	0	0
1	0	0	1	0	0
1	0	0	0	1	1
1	1	0	0	0	0 ← Initial sequence.

$N = 2^M - 1$

$N = 2^4 - 1 = 15.$

— 1M

RN

$\therefore$ ~~ben~~ PN sequence is

PN $\Rightarrow$ 0 0 0 1 1 1 1 0 1 0 1 1 0 0 1 — 1M

Autocorrelation sequence.

PN $\Rightarrow$ 0 0 0 1 1 1 1 0 1 0 1 1 0 0 1

$c(n) \Rightarrow$ 1 1 1 -1 -1 -1 -1 1 -1 1 -1 -1 ~~+~~ 1 1 -1

mapping $0 \to 1$

$1 \to -1$

$$R(d) = \frac{1}{N} \sum_{n=1}^{N} c(n)\, c(n-d)$$

for $d=0$ $\quad R(0) = \frac{1}{15} \sum_{n=1}^{15} c^2(n) = \frac{1}{15} \times 15 = 1$

for $d=1$ $\quad R(1) = \frac{1}{15} \sum_{n=1}^{15} c(n)\, c(n-1)$

$\therefore$ $c(n)$ = 1 1 1 -1 -1 -1 -1 1 -1 1 -1 -1 1 1 -1

$c(n-1)$ = -1 1 1 1 -1 -1 -1 -1 1 -1 1 -1 -1 1 1

$c(n)\,c(n-1)$ = -1 1 1 -1 1 1 1 -1 -1 -1 -1 1 -1 1 -1

$\therefore$ $R(1) = \frac{1}{15} \sum_{n=1}^{15} c(n)\, c(n-1) = \frac{1}{15}(-1)$

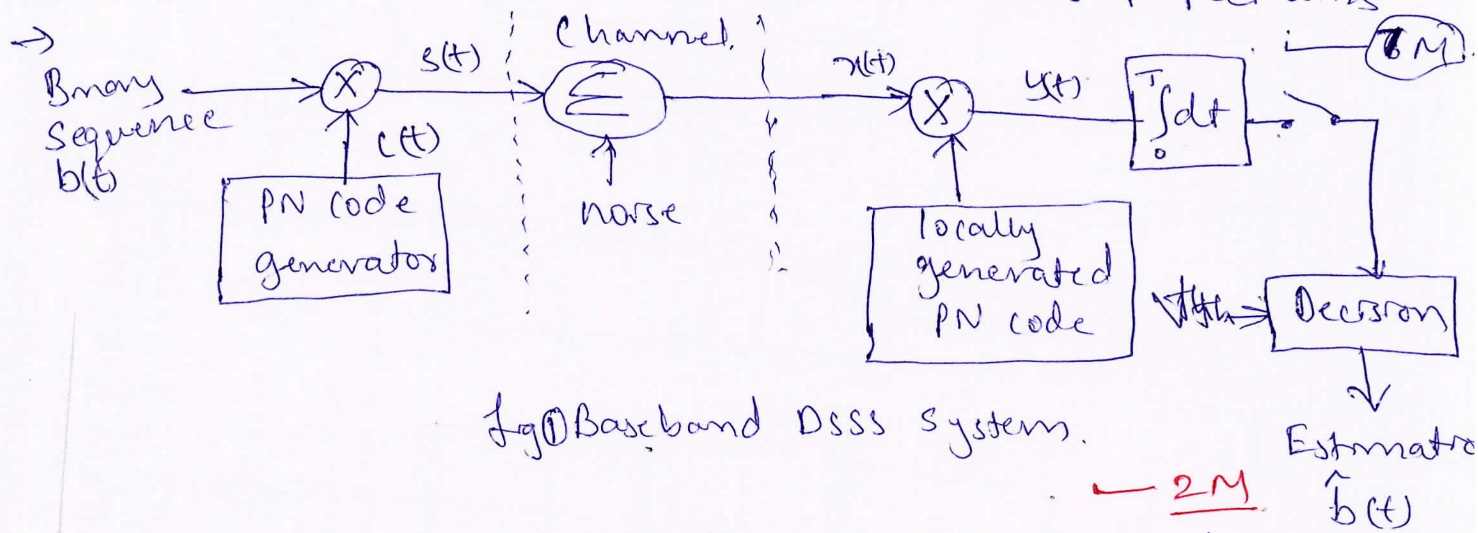
Similarly $\quad R(2) = R(3) = R(4) = \ldots = R(14) = \frac{-1}{15}$

$R(15) = R(0) = 1.$

$$\therefore \quad R(d) = \begin{cases} 1, & d = 0, \pm N, \pm 2N, \ldots \\ -1/15, & \text{others} \end{cases}$$

— 2M

Q9 b) Explain the generation of direct sequence spread Spectrum signal with the relevant waveforms & spectrums 31



fig① Baseband DSSS system. — 2M

The i/p binary sequence b(t) is the baseband signal in NRZ polar format. with each bit occupying a time of Tb seconds.

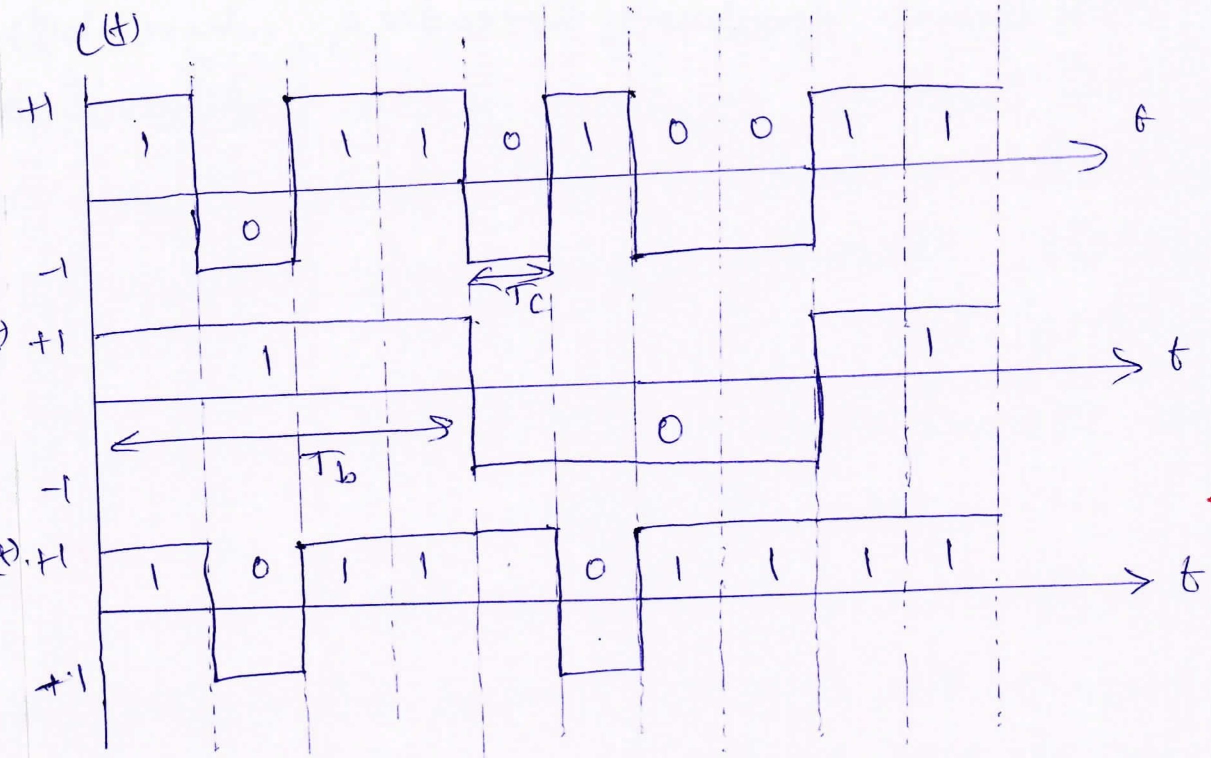
c(t) is a wideband code sequence which has noise like properties. This sequence is generated by a feedback shift registers. which is also expressed in NRZ polar format. But the duration of each bit (T_c) in c(t) is the period of the clock pulse that is applied to the generator.

Normally $T_c << T_b$ and hence bit rate of c(t) i.e., R_c also called as chip rate is much greater than $R_b = \frac{1}{T_b}$

The info-bearing signal b(t) is multiplied by the signal from the PN sequence generator i.e., c(t)

Q.9b

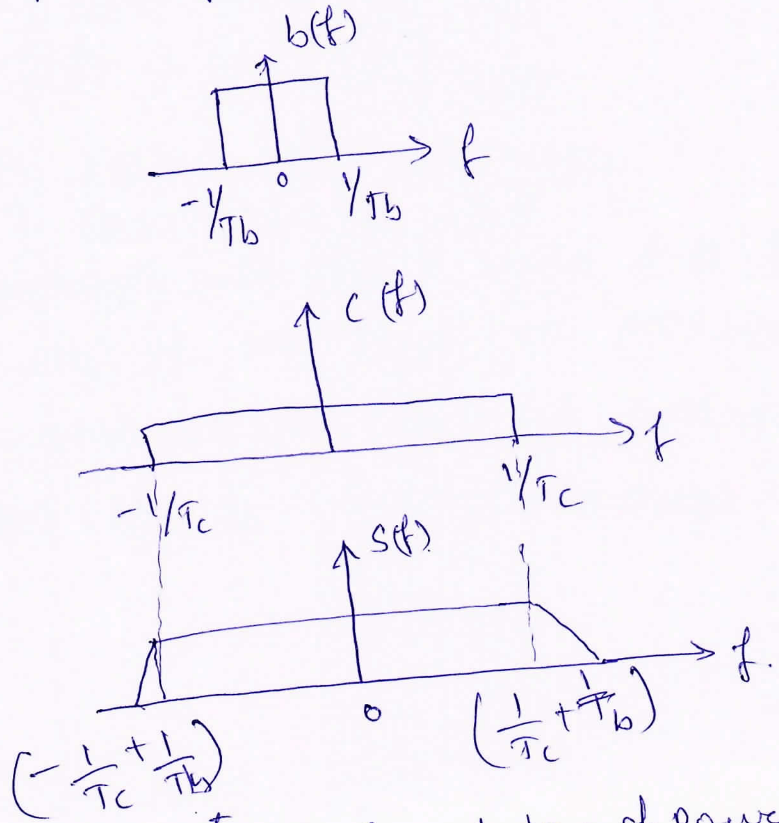
26



— 2M

Fig ②. generation of DSSS signal

This multiplication operation serves to spread the BW of the info-bearing signal into the wider BW occupied by PN sequence C(t).



— 1M

Fig ③. Convolution of power spectra.

38 The fig (3) shows spectrum spreading, where $b(f)$ is narrowband and PN sequence $c(f)$ is wideband, and the BW of the spectrum $S(f)$ is nearly equals to that of $c(f)$. Therefore PN sequence performs the role of a spreading code.

Let the received signal be $x(t)$, then.

$$x(t) = s(t) + i(t).$$

— 1M

where $i(t)$ is the interference signal,

The 1st step in the detection process is to multiply the received signal $x(t)$ by PN sequence $c(t)$. performed by product modulator.

$\therefore$ O/p of product modulator is

$$y(t) = x(t).\ c(t)$$

$$y(t) = [s(t) + i(t)]\ c(t).$$

$$y(t) = [b(t).\ c(t) + i(t)]\ c(t) \qquad \text{— 1M}$$

$$y(t) = b(t)\ c^2(t) + i(t).\ c(t). \quad \text{— (1)}$$

$$y(t) = b(t) + i(t).\ c(t) \quad [\because c^2(t) = 1]$$

Thus eqn (1) it is clear that the spectrums of $s(t)$ is despread, resulting in spectrum of $b(t)$

On multiplication by $c(t)$, the spectrum of $i(t)$ is spread but spectral height of $i(t).\ c(t)$ is small.

(RN)

Q1. (28)

Q9 b). Write a note on application of spread spectrum in wireless LAN's. → (3M)

→ There are many applications of spread spectrum, some of them are.

1) Low-detectability signal Transmission

In this application, the information-bearing signal is tmed at a very low. power level relative to the background channel noise and thermal noise that is generated in the front end of a receiver.

The intended receiver can recover the weak info-bearing signal from the background noise with the help of processing gain and coding gain. The other receiver having lack of info. abount PN code sequence is unable to take the advantage of processing gain & coding gain.

2) Code division Multiple Access (CDMA)

The spread spectrum permits many DSSS signals to occupy the same channel BW. Here all the users has its own pseudorandom sequence.

Therefore it is possible to have several users' transmit messages simultaneously over the same channel BW.

3) Wireless LAN's — 1M

Spread spectrum signals have been used in the IEEE wireless LAN standards 802.11 & 802.11b, which operate in the 2.4GHz ISM unlicensed freq. band.

The available BW is subdivided into 14 overlapping 22MHz channels, although not all channels are used in all countries. In USA only channels 1 ~~to~~ 11

40

are used.

In the 802.11 standard an 11-chip barker sequence is modulated and transmitted at a chip rate of 11 MHz.

The 11-chip Barker sequence is

$$\{1, -1, 1, 1, -1, 1, 1, 1, -1, -1, -1\}$$

This sequence is desirable because its autocorrelation has sidelobes of less than or equal to 1, compared with the peak autocorrelation value of 11.

The Barker sequence is modulated either with BPSK or QPSK.

DSSS is also used in the higher speed (2^{nd} generation) IEEE 802.11b wireless LAN standard, which operates in the same 2.4 GHz ISM band.

— 2M

10. a) With necessary block diagram, explain the transmitter and receiver of frequency hop spread spectrum → (8 M)

→ In FHSS technique, the carrier hops randomly from one freq. to another. By choosing a large number of randomly hoping discrete carrier frequencies, it is possible to have a modulated signal of wide BW, which results in a large processing gain.

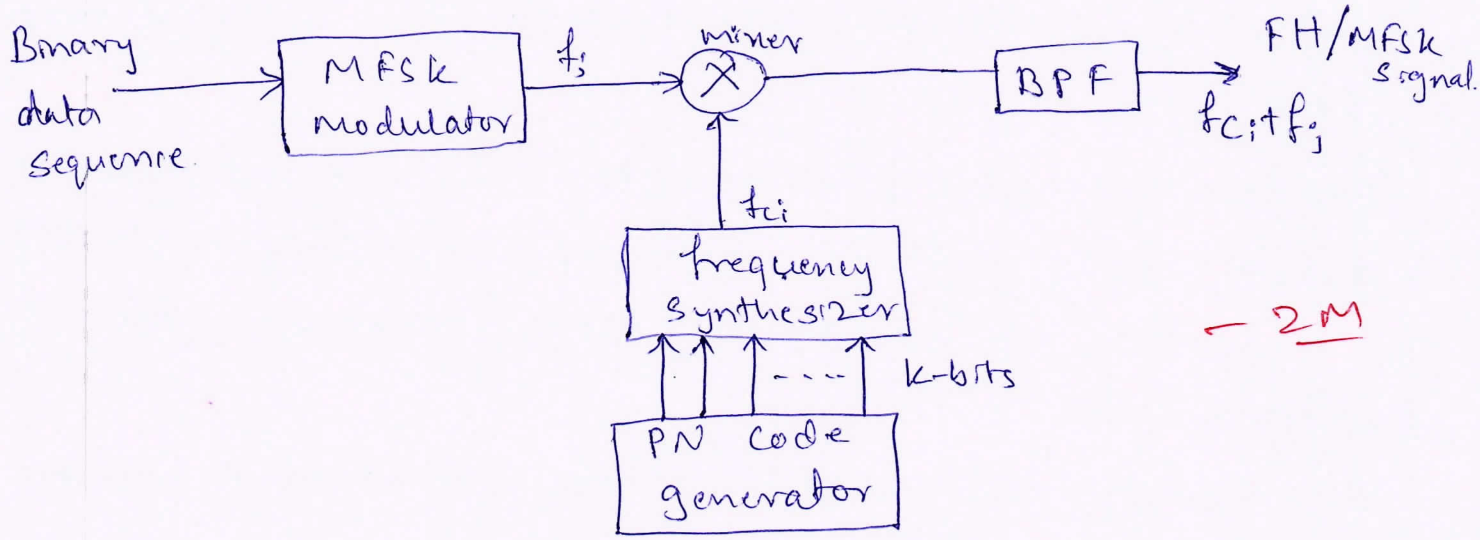


fig: (1): Block diagram of FH/MFSK transmitter

← 2M

In FHSS, spectrum spreading is sequential.

1^{st} stage : Frequency modulator (M-FSK)

2^{nd} stage : Frequency mixer (freq. adder)

Incoming binary sequence is applied to a serial to parallel converter to get blocks of k-bits.

Depending on k-bit binary pattern, any one of the 2^k discrete amplitude levels of m-ary PAM signal is obtained.

This M-ary PAM signal is applied to a voltage controlled oscillator (VCO).

← 1M

For every amplitude level in the m-ary PAM signal, the VCO produces one of the 2^k discrete freq.

The o/p of the MFSK modulator is then mixed with the o/p of a freq synthesizer.

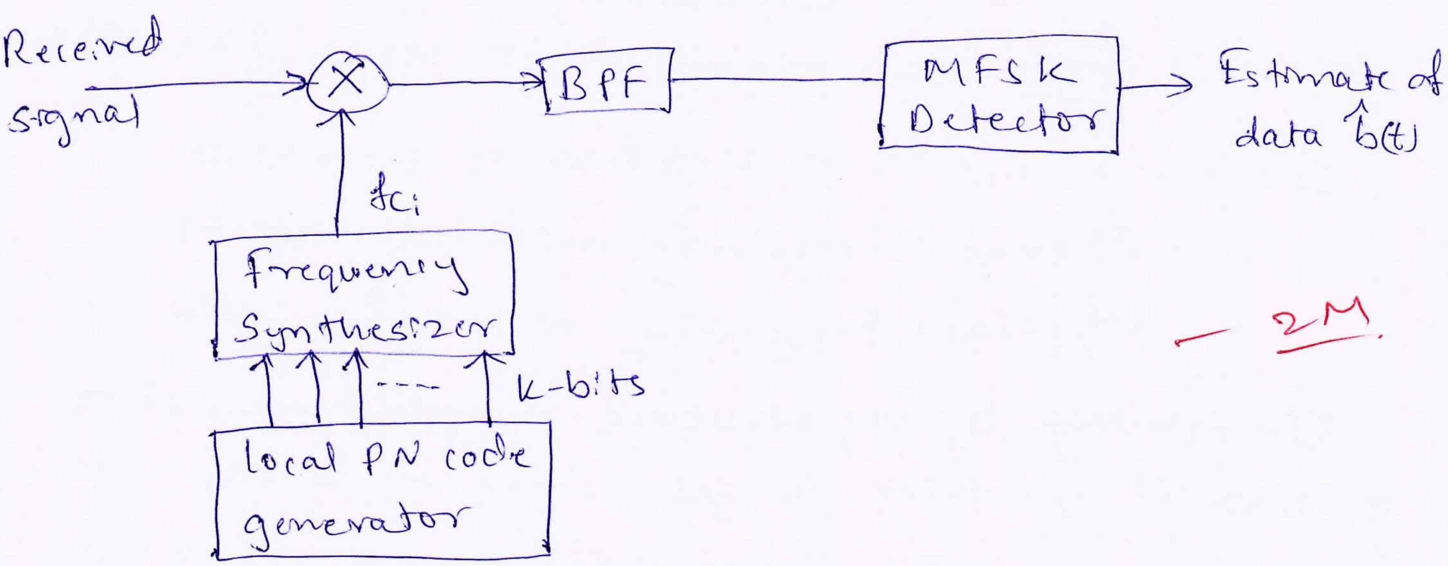
As a result freq. hops over 2^k distinct values.

The BPF passes the sum freq. for the transmission and rejects the difference freq. components.

FHSS signal occupies much larger BW ranging over a few GHz. — 1M

Therefore because of large FH BW, coherent detection is possible only within each hop, since freq. synthesizer are incapable of maintaining phase coherence over successive hops.

As a result, noncoherent MFSK detection is need.



— 2M

fig. ② : Block diagram of FH/MFSK Receiver.

In the 1st stage, mixing operation (down-conversion) removes the freq. hopping.

(AN)

10.a

(29)

The O/P of mixer is passed through a BPF which selects the difference freq. component from the mixer.

The O/P of the BPF is the MFSK signal, which is demodulated using non-coherent MFSK detector.

The non-coherent MFSK detector consists of a bank of M matched filters each of which is matched to one of the $M = 2^k$ MFSK tones.

An estimate of the original symbol transmitted is obtained by selecting the largest filter O/P.

– 2M.

(RN)

Q1 (30)

10 b). With a neat block diagram explain the CDMA systems based on IS-95 — (8M). 27

→ CDMA technique overcomes the problem of FDMA and TDMA, In situations where all the users are required to transmit simultaneously. but also occupy the same RF BW of the channel, CDMA can be used.

DS-CDMA has been adopted as a multiple access method for a digital cellular voice communication in North America and was developed by Qualcomm as a standard. TIA designated the standard as IS-95 for use in the 800MHz and 1900MHz bands.

A major advantage of CDMA over other multiple access methods is that entire freq. band is available at each base station, The BW used for transmission from a base station to the mobile receiver is 1.25 MHz.

The signals transmitted in both forward & reverse links are DSSS signals having a chip rate of 1.2288×10^6 chips/s. — 1M

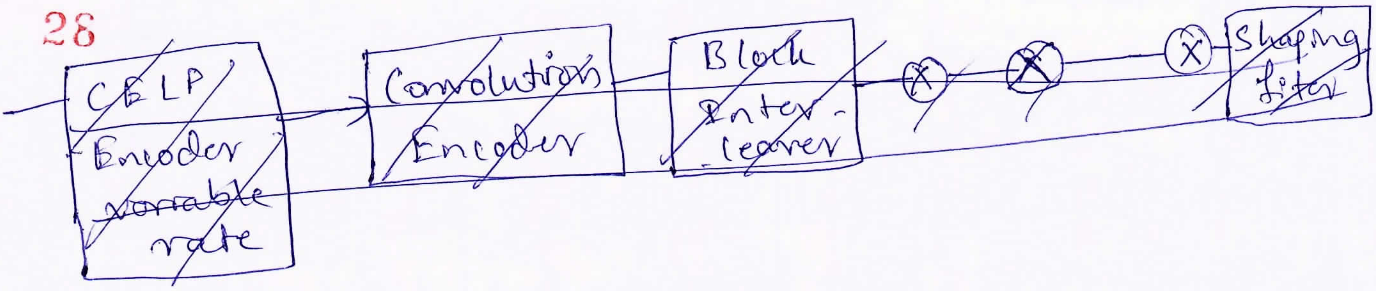
* the Block diagram of IS-95 Forward link is given below

Speech encoder Code excited linear Predictor (CELP). generates data at variable rates of 9600, 4800, 2400, 1200 bps in frame intervals of 20msec.

Encoded Speech is passed through convolutional encoder and block interleaver. — 1M

Forward Link

28



For lower speech data rates of 4800, 2400, or 1200 bps, the O/p symbols of convolutional encoder is seperated twice, 4 times or eight times. Thus the bit rate is fixed at 9600 bps.

The O/p of block interleaver is at 19.2bps and is multiplied by long code (of period $N = 2^{42} - 1$) of chip rate 1.2288×10^6 chips per second & decimated by a factor 64. — 1M

The long code identifies the mobile station in forward and reverse links uniquely.

Each channel user is assigned a walsh sequence of 64 bits uniquely from a set of 64 sequences.

One sequence is used to transmit pilot signal.

Pilot signal is used to estimate channel characteristics like signal strength and carrier phase and employed in coherent detection.

One sequence is used for time synchronization and some sequences for paging.

About remaining 61 channels are assigned to different users.

Each user multiplies the data sequences by 64 bit walsh sequence. — 1M

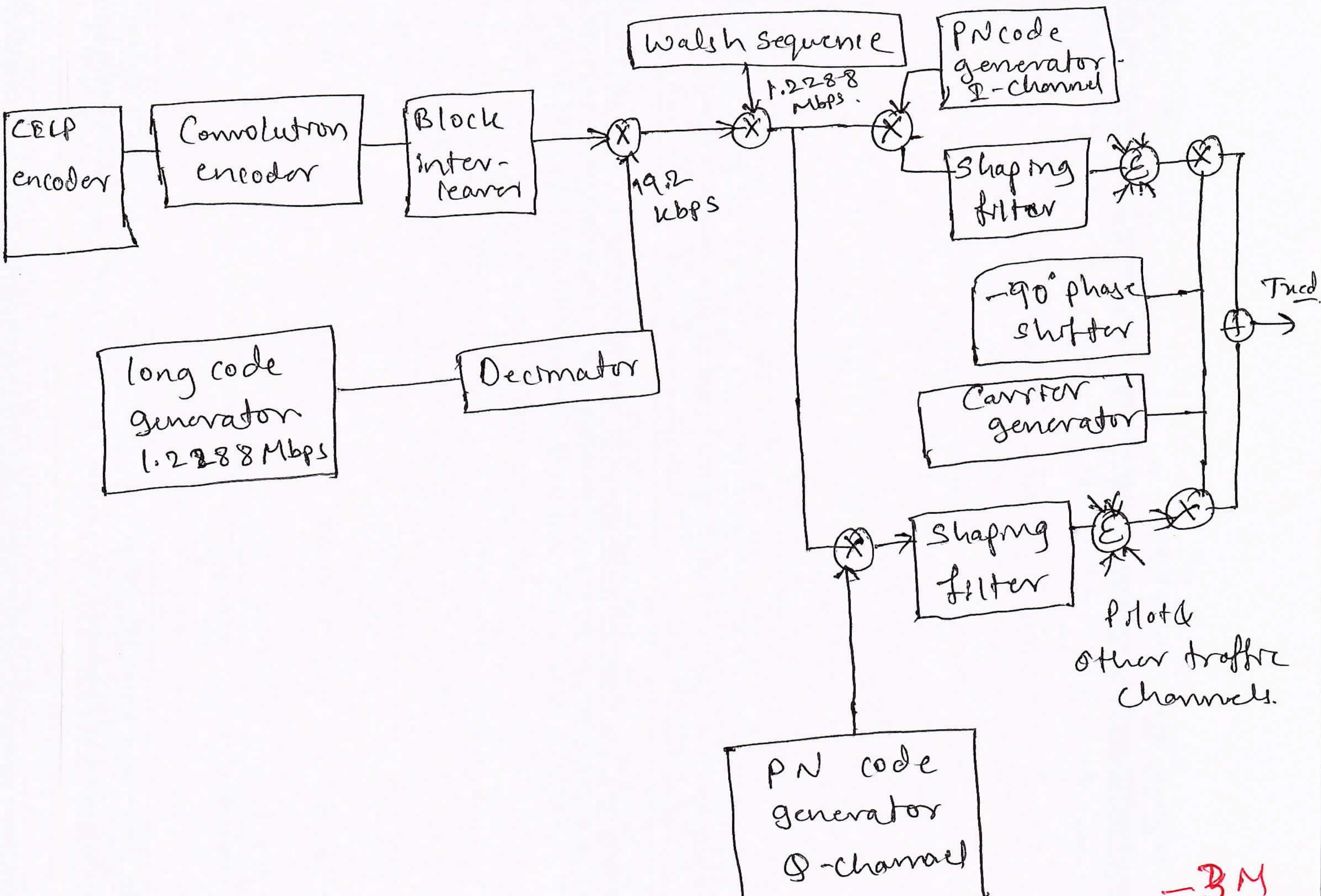
The resulting ~~binary~~ sequence is spread by multiplication with two PN sequences of length $N \cong 2^{15}$ creating I and Q signal components, resulting in 4 phase signal.

Different base stations are identified by the offsets of these PN sequences.

At the Rx a RAKE demodulator is used to seperate the multipath components and combined before detection using Viterbi soft decision.

— 1M

Demodulator uses non-coherent demodulation of the 64 orthogonal walsh sequences to recover the encoded data bits.



— 3M

Fig: Block diagram of IS-95 Forward Link