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Name of Institute : KLS VDIT, Haliyal

Department of Electronics and Communication Engineering

Sem : 6

B.E. Degree Examination
Year: July / August 2022

Subject : Digital communication

Subject Code : 18EC61

Scheme and Solution

Prof. Plasin Dias / Prof. A. S. Joshi
Subject Teachers

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HOD E&C
Head of the Department
Dept. of Electronic & Communication Engg.
KLS V.D.I.T., HALIYAL (U.K.)

Modified

CBCS SCHEME

USN | | | | | | | | | | |

18EC61

Sixth Semester B.E. Degree Examination, July/August 2022
Digital Communication

Time: 3 hrs. Max. Marks: 100

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

Module-1

1 a. What are the applications of Hilbert transform? Prove that a signal g(t) and its Hilbert transform $\hat{g}(t)$ are orthogonal over the entire time interval $(-\infty, \infty)$. **(08 Marks)**

b. For a binary sequence 0 1 0 0 0 0 0 0 0 1 0 1 1 construct :
i) RZ Bipolar format ii) Manchester format iii) B3ZS format iv) B6ZS format
v) HDB3 format. **(08 Marks)**

c. Define Pre-envelope of a real valued signal. Given a band pass signal S(t) sketch the amplitude spectra of signal S(t). Pre-envelope $S_+(t)$ and Complex envelope $\tilde{S}(t)$. **(04 Marks)**

OR

2 a. Express Bandpass signal S(t) in canonical form. Also explain the scheme for deriving the in-phase and quadrature components of the band pass signal S(t). **(08 Marks)**

b. Derive the expression for the complex low pass representation of band pass systems. **(08 Marks)**

c. Write a note on HDBN signaling. **(04 Marks)**

Module-2

3 a. Explain the geometric representation of set of M energy signals as linear combination of N orthonormal basis functions. Illustrate for the case N = 2 and M = 3 with necessary diagrams and expressions. **(10 Marks)**

b. Explain the Correlation receiver using product integrator and matched filter. **(10 Marks)**

OR

4 a. Using the Gram – Schmidt Orthogonalization procedure, find a set of orthonormal basis functions to represent the three signals $S_1(t)$, $S_2(t)$ and $S_3(t)$ shown in Fig. Q4(a). Also express each of these signals in terms of the set of basis functions. **(12 Marks)**

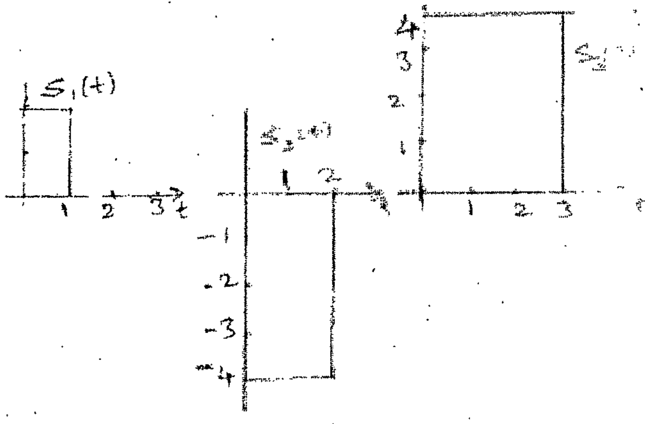

Fig. Q4(a)

b. Show that for a noisy input, the mean value of the j^{th} correlator output X_j depends only on S_{ij} and all the correlator outputs X_j, j = 1, 2, ... N have a variance equal to the PSD $\frac{N_0}{2}$ of the additive noise process W(t). **(08 Marks)**

Module-3

5 a. Derive the expression for error probability of binary PSK using coherent detection. (06 Marks)

b. Explain the generation and optimum detection of differential phase – shift keying, with neat block diagram. (08 Marks)

c. A binary data is transmitted over a microwave link at a rate of 10^6 bits/sec and the PSD of noise at the receiver is 10^{-10} watts/Hz. Find the average carrier power required to maintain an average probability of error $P_e \le 10^{-4}$ for coherent binary FSK. What is the required channel bandwidth? (Given erf (2.6) ≐ 0.9998). (06 Marks)

OR

6 a. With a neat block diagram, explain the non – coherent detection of binary frequency shift keying technique. (08 Marks)

b. In a FSK system, following data are observed. Transmitted binary data rate = 2.5×10^6 bits/second PSD of zero mean AWGN = 10^{-20} Watts/Hz. Amplitude of received signal in the absence of noise = $1\mu V$. Determine the average probability of symbol error assuming coherent detection. (Given erf (2.5) = 0.99959). (08 Marks)

c. What is the advantage of M – ary QAM over M – ary PSK system? Obtain the constellation of QAM for M = 4 and draw signal space diagram. (04 Marks)

Module-4

7 a. With a neat block diagram, explain the digital PAM technique through band limited base band channels. Also obtain the expression for inter symbol interference. (08 Marks)

b. State and prove Nyquist condition for zero ISI. (08 Marks)

c. With neat diagram and relevant expression, explain the concept of adaptive equalization. (04 Marks)

OR

8 a. For a binary data sequence $\{d_n\}$ given by 1 1 1 0 1 0 0 1. Determine the precoded sequence, transmitted sequence, received sequence and the decoded sequence. (06 Marks)

b. Draw and explain the time – domain and frequency domain of duo – binary and modified duo binary signal. (08 Marks)

c. With neat diagram, explain the timing features pertaining to eye diagram and its interpretation for base band binary data transmission system. (06 Marks)

Module-5

9 a. Explain the model of a Spread Spectrum digital Communication system. (08 Marks)

b. Explain the effect of dispreading on a narrow band interference in Direct Sequence Spread Spectrum System (DSSS). A DSSS signal is designed to have the power ratio P_R/P_N at the intended receiver is 10^{-2}. If the desired E_b/N_0 = 10 for acceptable performance determine the minimum value of processing gain. (08 Marks)

c. What is a PN sequence? Explain the generation of maximum length (ML – Sequence). What are the properties of ML sequences? (04 Marks)

OR

10 a. With a neat block diagram, explain frequency Hopped Spread Spectrum technique. Explain the terms Chip rate, Jamming Margin and Processing gain. (10 Marks)

b. With a neat block diagram, explain the CDMA System based on IS - 95. (10 Marks)

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KLS VDIT, Haliyal

Department of Electronics and Communication Engg.

Subject: Digital Communication - 18EC61 Year: 2022

1 a. What are the applications of Hilbert transform? Prove that a signal g(t) and its Hilbert transform $\hat{g}(t)$ are orthogonal over the entire time interval $(-\infty, \infty)$

The way of separation of signals is based on phase selectivity which uses phase shifts between pertinent signals to achieve the desired separation. A phase shift of special interest in this context is that of $\pm 90°$. In particular when the phase shift of all components of a given signals are shifted by $\pm 90°$ the resulting function of time is known as the "Hilbert Transform" of the signal. It is also called as quadrature filter, to emphasize its district property of providing $\pm 90°$ phase shift. Hilbert transform of signal g(t) is denoted as $\hat{g}(t)$, given by

$$\hat{g}(t) = \frac{1}{\pi}\int_{\tau=-\infty}^{\infty} \frac{g(\tau)}{t-\tau} \cdot d\tau$$

Hilbert transform is a linear operation.

Hilbert transform is used in generation of SSB signals, representation of bandpass signals and designing minimum phase type filters.

Applications of Hilbert transform

(i) Generation of single side band modulation to realize phase selectivity.

(ii) It is mathematical basis for representation of band pass signals.

A signal $g(t)$ and its Hilbert transform $\hat{g}(t)$ are orthogonal over the entire time interval $(-\infty, \infty)$.

Orthogonality of $g(t)$ & $\hat{g}(t)$ is described by

$$\int_{-\infty}^{\infty} g(t)\,\hat{g}(t)\cdot dt = 0$$

<u>Proof</u>: $\int_{-\infty}^{\infty} g(t)\cdot\hat{g}(t) = \int_{-\infty}^{\infty} G(f)\cdot\hat{G}(-f)\cdot df$

We know
$\hat{G}(f) = -j\,\mathrm{sgn}(f)\cdot G(f)$

$$= -j\int_{-\infty}^{\infty} \mathrm{sgn}(f)\cdot G(f)\,G(-f)\cdot df$$

$$= -j\int_{-\infty}^{\infty} \mathrm{sgn}(f)\cdot G(f)\cdot G^{*}(f)\cdot df$$

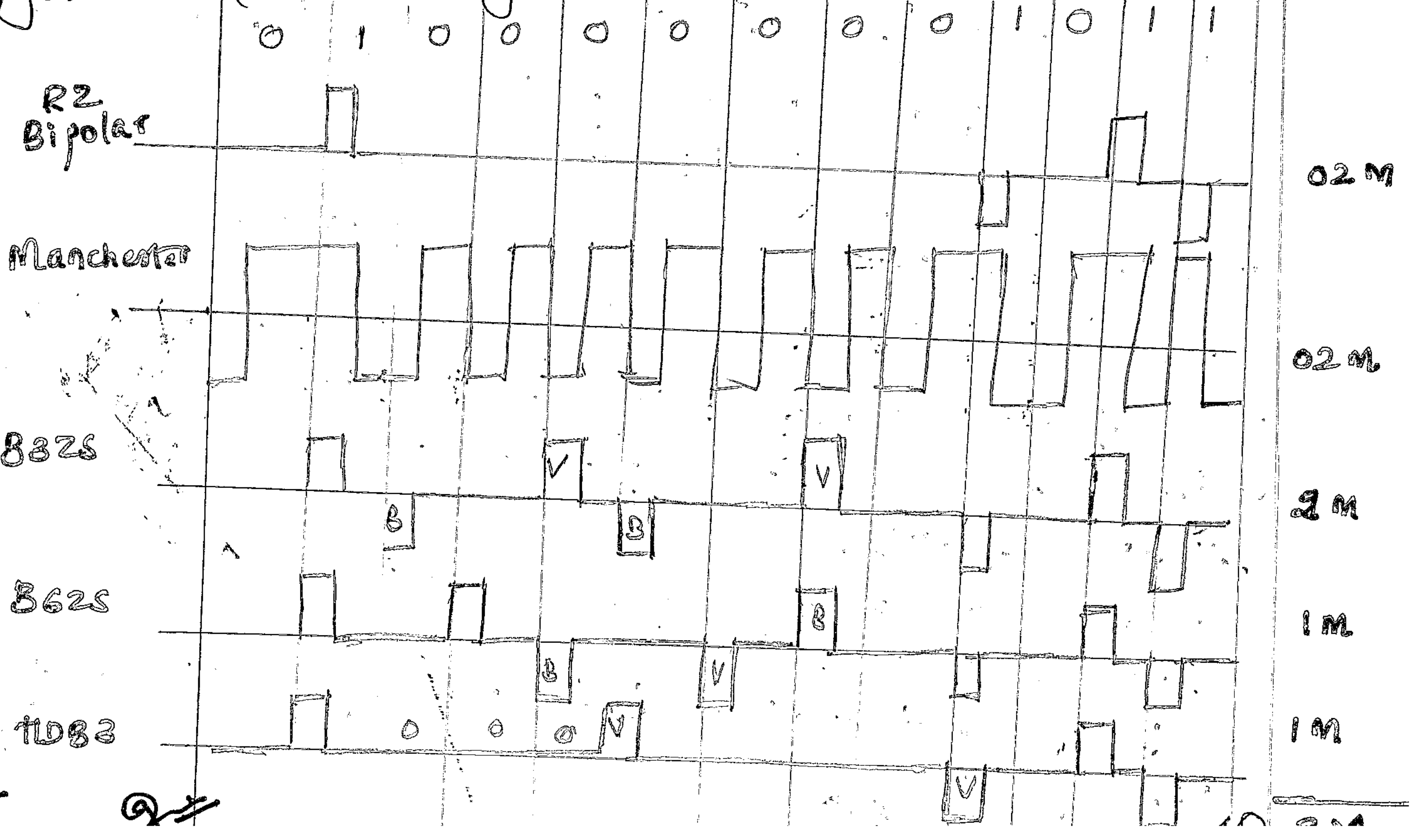
$$= -j\int_{-\infty}^{\infty} \mathrm{sgn}(f)\,|G(f)|^2\, df$$

$\int_{-\infty}^{\infty}$ product of the odd function sgn(f) over the even function $|G(f)|^2 = 0$. $\therefore \int_{-\infty}^{\infty} g(t)\cdot\hat{g}(t) = 0$

1 a)

Explanation	4
+ Proof	4
	8M.

1 (b) For binary sequence 010000000001011 construct (i) RZ Bipolar format (ii) Manchester format (iii) B3ZS format (iv) B6ZS format (v) HDB3 format.



page2

1 c) Define Preenvelope of real valued signal, Given a band pass signal $s(t)$, sketch the amplitude spectra of the signal $s(t)$, Preenvelope $s_+(t)$ and complex envelope $\tilde{s}(t)$

The preenvelope of $g(t)$ is defined as,

$$g_+(t) = g(t) + j\hat{g}(t)$$

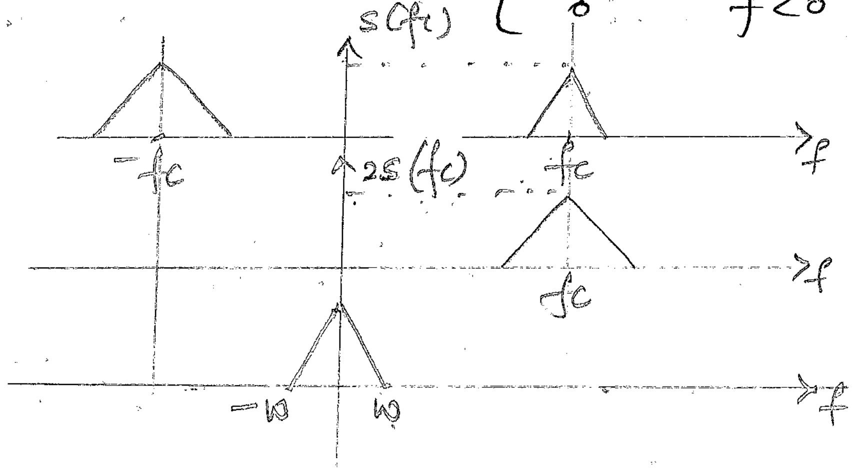
where $\hat{g}(t)$ is Hilbert transform of $g(t)$

Given a signal $g(t)$ is the real part of the preenvelope $g_+(t)$ and Hilbert transform $\hat{g}(t)$ is the imaginary part of the preenvelope.

Let $G_+(f)$ denote fourier transform of $g_+(t)$

then, $G_+(f) = G(f) + \operatorname{sgn}(f)\, G(f)$

$$G_+(f) = \begin{cases} 2G(f) & f > 0 \\ G(0) & f = 0 \\ 0 & f < 0 \end{cases}$$



Band Pass signal

Pre Envelope

Complex envelope.

(1c) Def^n + Diagram 1M + 3M = 4M

2a. Express Bandpass signal $s(t)$ in canonical form. Also explain the scheme for deriving the interphase and quadrature components of the band pass signal $s(t)$

By definition,

the real part of preenvelope $s_+(t)$ is equal to the original band pass signal

$$s_+(t) = s(t) + j\hat{s}(t) \quad \text{——— (1)}$$

$s(t) \leftarrow$ band pass signal

$\hat{s}(t) \leftarrow$ Hilbert transform of $s(t)$

$s_+(t) \leftarrow$ Pre envelope of $s(t)$

The complex envelope of band pass signal,

$$S_{+}(t) = \hat{s}(t)\, e^{j2\pi f c t} \quad \text{——— ②}$$

$\hat{s}(t)$ ← complex envelope of the band pass signal $s(t)$

The band pass signal $s(t)$ in terms of complex envelope $\tilde{s}(t)$

$$s(t) = \mathrm{Re}\left[\tilde{s}(t)\exp(j2\pi f c t)\right] \quad \text{——— ③}$$

$\tilde{s}(t)$ is a complex valued quantity and it is expressed in cartesian form as

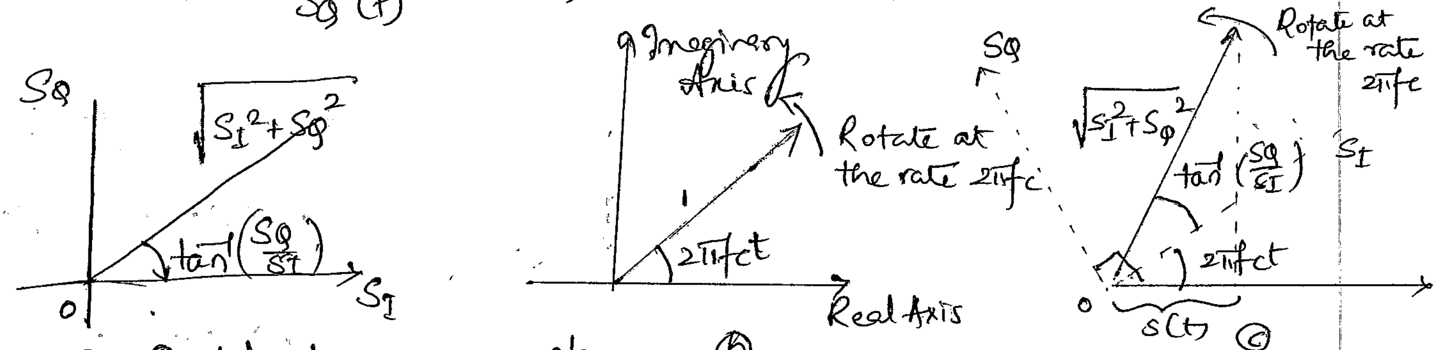
$$\tilde{s}(t) = S_I(t) + j S_Q(t) \quad \longleftarrow \text{④}$$

$S_I(t)$ and $S_Q(t)$ are both real valued low pass functions. Use Eqn ④ in Eqn ③,

$$s(t) = S_I(t)\cos(2\pi f c t) - S_Q(t)\sin(2\pi f c t) \quad \text{——— ⑤}$$

Equation ⑤ is standard form or canonical form

where $S_I(t)$ — Inphase component } of bandpass
$S_Q(t)$ — Quadrature component } signal $s(t)$



ⓐ Quadrature components of the signal ⓑ ⓒ

fig 2(a) complex Envelope $\tilde{s}(t)$ and its multiplication by $\exp(j2\pi f c t)$.

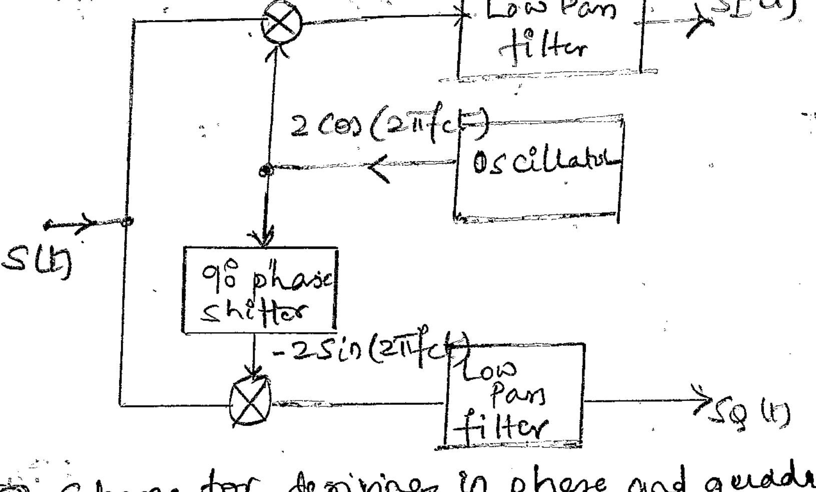


fig 2(b) Scheme for deriving in phase and quadrature components of band signal $s(t)$

$S_I(t)$ and $S_Q(t)$ are low pass signals, limited to band $-W \le f \le W$

Both the low pass filters are designed identically with bandwidth equal to W.

The fig (a) is known as Analyzer, extracts inphase ($S_I(t)$) and quadrature components ($S_Q(t)$) from band pass signal $s(t)$

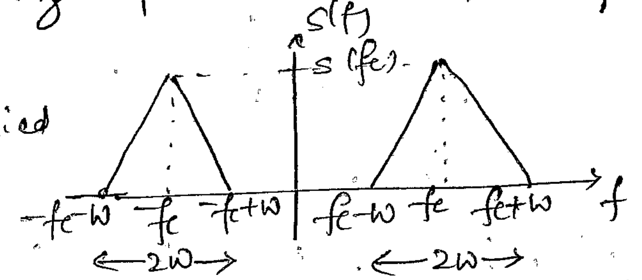
2@ Explanation (Derivation) 4
+ Diagram 4
8M

2(B) Derive the expression for the complex low pass representation of Band pass system.

The band pass signal $s(t)$ is called narrow band signal and its fourier transform is denoted by $S(f)$.

The spectrum of the signal $s(t)$ is limited to frequencies within $\pm W$ Hertz of the carrier frequency f_c

Assume $f_c > W$.



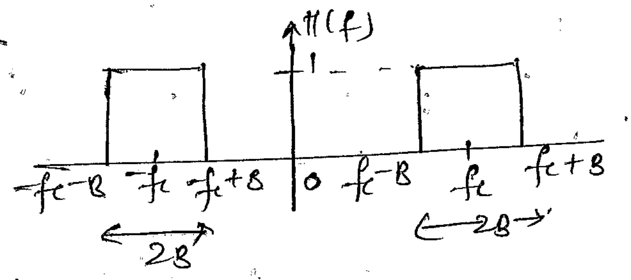
Spectrum $S(f)$ of band pass signal $s(t)$

Consider signal $s(t)$ is applied to linear time invariant (LTI) band pass system with impulse response $h(t)$

$H(f)$ is frequency response of $h(t)$

The frequency response $H(f)$ of the bandpass system is limited to frequencies within $\pm B$ Hz of the carrier frequency f_c Hz

Assume $B < f_c$



The system bandwidth is 2B and is narrower than or equal to input signal bandwidth 2W.

Represent band pass impulse response $h(t)$ in terms of $h_I(t)$ and $h_Q(t)$

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

Define complex impulse response of the bandpass system as

$$\tilde{h}(t) = h_I(t) + j h_Q(t) \quad \text{——} (2)$$

$$s(t) = Re[\tilde{s}(t) e^{j2\pi f_c t}]$$

By analogy, write,

$$h(t) = Re[\tilde{h}(t) e^{j2\pi f_c t}] \quad \text{——} (3)$$

$\tilde{h}(t)$, $h_I(t)$ & $h_Q(t)$ are all low pass functions, limited to band $-B < f < B$

Determine complex impulse response $\tilde{h}(t)$ in terms of $h_I(t)$ and $h_Q(t)$.

Complex conjugate impulse response $\tilde{h}^*(t)$ in terms of $h_I(t)$ and $h_Q(t)$

$$h(t) = Re[\tilde{h}(t) e^{j2\pi f_c t}] \quad \text{——} (a)$$

$$h(t) = Re[\tilde{h}^*(t) e^{-j2\pi f_c t}] \quad \text{——} (b)$$

Adding Eq (a) & (b)

$$h(t) + h(t) = Re[\tilde{h}(t) \cos 2\pi f_c t + j \tilde{h}(t) \sin 2\pi f_c t + \tilde{h}^*(t) \cos 2\pi f_c t - j \tilde{h}^*(t) \sin 2\pi f_c t$$

$$\therefore 2h(t) = \tilde{h}(t) e^{j2\pi f_c t} + \tilde{h}^*(t) e^{-j2\pi f_c t} \quad \text{——} (4)$$

$\tilde{h}^*(t)$ is the complex conjugate of $\tilde{h}(t)$

Apply fourier transform to both sides of Eq (4)

Use frequency shift property of fourier transform

then,

$$2H(f) = \tilde{H}(f - f_c) + \tilde{H}^*(-f - f_c) \quad \text{——} (5)$$

For a specified bandpass frequency response $H(f)$, we may determine corresponding complex low pass frequency response $\tilde{H}(f)$ by taking the part of $H(f)$ defined for the positive frequencies, shifting it to origin and scaling it by the factor 2 ie $\tilde{H}^*(-f-f_c) = 0$ for $f > 0$

2(b) Diagram 2 + Derivation 3 + Explanation 3 = 8M

$\therefore$ Eqn (5) becomes

$$\tilde{H}(f-f_c) = 2H(f), \quad f > 0 \qquad (6)$$

Having determined the complex frequency response $\tilde{H}(f)$ we decompose it into inphase and quadrature component as

$$\tilde{H}(f) = \tilde{H}_I(f) + j\tilde{H}_Q(f) \qquad (7)$$

where $$\tilde{H}_I(f) = \left| \frac{\tilde{H}(f) + \tilde{H}^*(-f)}{2} \right| \qquad (8)$$

& $$\tilde{H}_Q(f) = \left| \frac{\tilde{H}(f) - \tilde{H}^*(-f)}{2j} \right| \qquad (9)$$

finally, to determine impulse response $\tilde{h}(t)$ of the band pass system, take the IFT, (Inverse fourier Transform) of $\tilde{H}(f)$, then we get,

$$\tilde{h}(t) = F^{-1}[\tilde{H}(f)]$$

$$\tilde{h}(t) = \int_{-\infty}^{\infty} \tilde{H}(f)\, e^{j2\pi ft}\, df \qquad (10)$$

Required Equation.

2 c) Write a note on HDBN signaling

HDB scheme is an International Telecommunications Union (ITU) standard.

The problem of nontransparency in Bipolar signaling is eliminated by adding pulses, when the number of consecutive 0's exceeds N.

Such a modified coding is designated as high density bipolar coding (HDBN) where N=1,2,3

Most important of HDB code is HDB3 format, which has been adopted as International standard

The basic idea of HDBN code is that when run of N+1 zeros occurs, this group of zeros is replaced by one of special N+1 binary digit sequences.
To increase the timing content of the signal, the sequences are choosen to include some binary 1's. The 1's included deliberately violates the bipolar rule, for easy identification of the substituted sequences.

In HDB3 coding, special sequence used are 000V and B00V, where B=1 that conform to the bipolar rule and V=1 that violates the bipolar rule.

2c) Explanation 2
+ Example 2
4M

3 (a) Explain the geometric representation of set of M energy signal orthonormal basis functions. Illustrate for the case N=2 and M=3 with necessary diagrams and expressions.

Geometric representation of signals provides compact alternative characterization of signals. It can simplify analysis of their performance as modulation signals.

The essence of geometric representation is to represent any set of M energy of signals $\{S_i(t)\}$ as linear combination of N orthonormal basis functions, where $N \leq M$.

Given a set of real valued energy signals $S_1(t)$, $S_2(t)$ - - - $S_M(t)$, each of duration T seconds.

$$S_i(t) = \sum_{j=1}^{j=N} S_{ij}\, \phi_j(t) \qquad \begin{cases} i = 1, 2, \dots M \\ j = 1, 2, \dots N \end{cases} \quad \text{— Eq ①}$$

$$0 \le t \le T$$

where coefficients of the expressions are

$$S_{ij} = \int_0^T S_i(t)\, \phi_j(t)\, dt \qquad \begin{cases} i = 1, 2, \dots M \\ j = 1, 2, \dots N \end{cases} \quad \text{--- Eq ②}$$

The real valued basis functions $\phi_1(t), \phi_2(t) \dots \phi_N(t)$ form an orthonormal set, by which we mean

$$\int_0^T \phi_i(t)\, \phi_j(t)\, dt = \delta_{ij} = \begin{cases} 1 & \text{if } i=j \\ 0 & \text{if } i \neq j \end{cases} \quad \text{— ③}$$

for $i=j$. $$\int_0^T \phi_i(t)\, \phi_j(t)\, dt = \int_0^T \phi_i(t)^2\, dt = 1 \quad \text{— 3a}$$

normalized to have unit energy

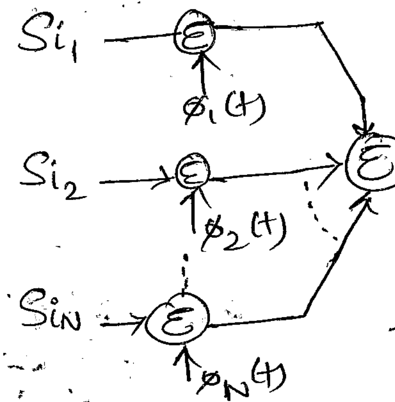
δ_{ij} = Kronecker delta.

for $i = j$

$$\int_0^{T_b} \phi_i(t)\, \phi_k(t)\, dt = 0 \quad \text{— 3b}$$

Two basis functions are orthogonal to each other over the interval $0 \le t \le T$

Synthesizer for generating the signal $S_i(t)$ is,



fig(a) Synthesizer for $S_i(t)$

The set of coefficients $\sum_{j=1}^{N} S_{ij}$ is viewed as an N dimensional signal vector denoted as S_i. Vector S_i bears one to one relationship with the transmitted signal $S_i(t)$.

Given the N elements of the vector s_i operating as input to generate the signal $s_i(t)$, the synthesizer consists of a bank of N multiplier with each multiplier having its own basis function followed by summer.

Analyzer for reconstructing the signal vector s_i

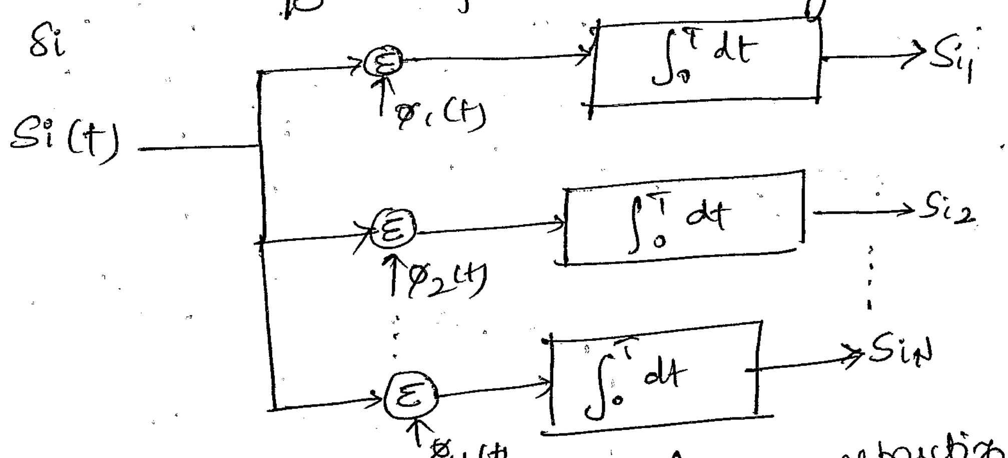


fig (2) Analyzer for reconstructing signal vector s_i.

Given the signals $s_i(t)$, $i = 1, 2, \dots M$ operating as input, calculate the coefficients $s_{i1}, s_{i2}, \dots s_{iN}$. The analyzer shows a bank of N product integrator or correlators with a common input and with each one of them supplied with its own basis function.

Each signal in the set $\{s_i(t)\}$ is completely determined by the signal vector s_i

where $s_i = \begin{bmatrix} s_{i1} \\ s_{i2} \\ s_{iN} \end{bmatrix}$ $i = 1, 2, \dots M$ ——— ④

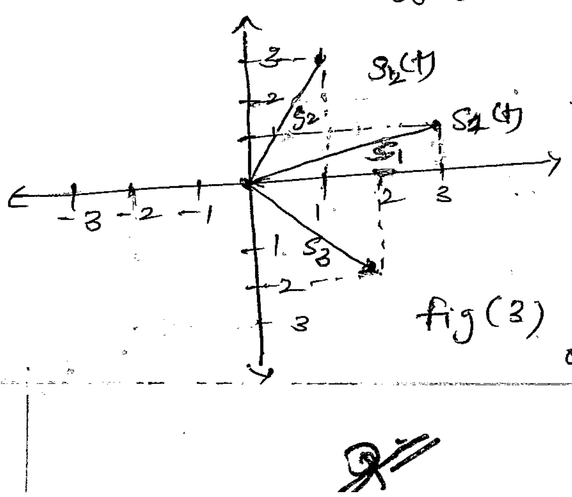


fig (3) Geometric representation of signals for case $M=3$, $N=2$

8a) Diagram 5
+ Equation + 2
+ Explanation + 3

10M

3 b) Explain the Correlation Receiver Using product integrator and matched filter.

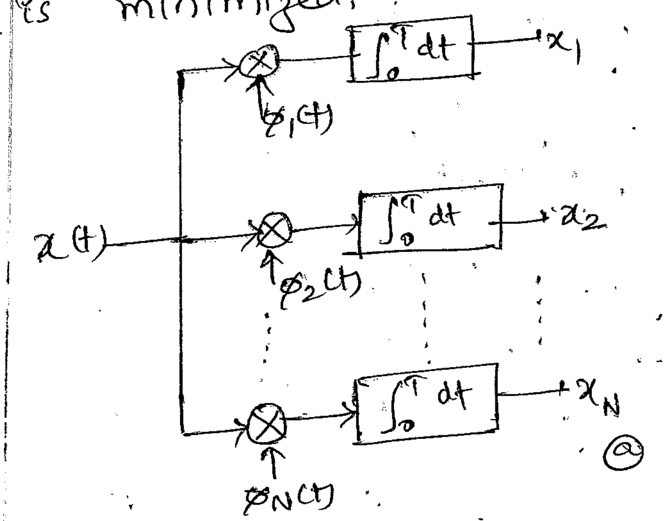
Correlation Receiver: for AWGN channel, when the transmitted signals are equally likely the optimum receiver which minimizes average probability of error is correlation receiver.

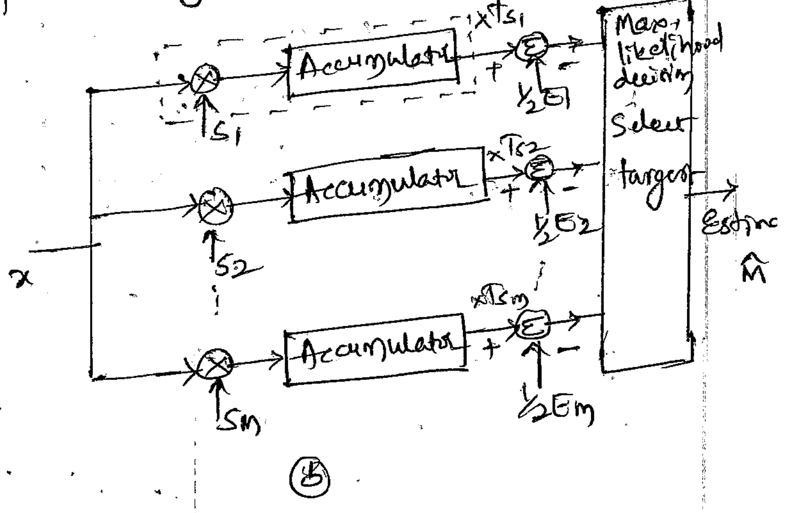
Correlation receiver works on maximum likelihood decision and consists of two subsystems which are represented in fig(a) and fig(b).

fig (a) first part of correlation receiver : bank of N correlators and fig(b) second part of correlation receiver, estimate $\hat{m}$.

1) Detector: fig (a) consists of M correlators, supplied with a set of orthonormal basis functions $\phi_1(t), \phi_2(t)$ that are generated locally. The bank of correlators operates on the received signal $x(t)$, $0 \le t \le T$ to produce the observation vector x,

2) Maximum likelihood decoder; fig(b); which operates on the observation vector x to produce an estimate $\hat{m}$ of the transmitted symbol m_i, $i = 1, 2, \ldots M$ in such a way that the average probability of symbol error is minimized.





(a) Detector or demodulator
(b) signal transmission decoder

In accordance with maximum likelihood decision rule

$$Y_k = \sum_{j=1}^{N} x_j S_{kj} - \frac{1}{2} E_k$$

the decoder multiplies N elements of the observation vector x by the corresponding N elements of each of the M signal vectors $S_1, S_2 \ldots S_M$. The resulting products are summed in accumulators. (Inner product block consists of product modulator and accumulator) to form the corresponding set of inner products. $\{x^T S_k \mid k = 1, 2, 3 \ldots M\}$ The inner products are corrected for the fact that transmitted signal energy may be unequal., finally the largest one in the resulting set of numbers is selected and using maximum likelihood decision rule or estimation of transmitted message $\hat{m}$ is done.

Matched filter Receiver: Consider a linear filter with impulse response $h_j(t)$ with the received signal $x(t)$ as the input. The resulting filter output $Y_j(t)$ for the i^{th} correlator is given by the convolution of $x(t)$ & $h_i(t)$ ie,

$$Y_i(t) = \int_{\tau=-\infty}^{\infty} x(\tau)\, h_i(t-\tau)\, d\tau \quad \text{——— (1)}$$

Evaluate this integral, over the duration of a Transmitted symbol $0 \le t \le T$

$$Y_j(T) = \int_{t=0}^{t=T} x(t)\, h_j(T-t)\, dt \quad \text{——— (2)}$$

Output of ith correlator is defined by,

$$x_i = \int_{t=0}^{t=T} x(t) \cdot \phi_j(t)\, dt \quad \text{——— (3)}$$

for $Y_i(T)$ to equal x_i, from Eq (2), this condition is satisfied provided that we choose

$$h_i(T-t) = \phi_j(t) \quad \text{for } 0 \le t \le T \;\&\; i = 1, 2, \ldots M$$

The condition imposed on the desired impulse response of filter as

$$h_j(t) = \phi_j(T-t) \quad \text{for } 0 \le t \le T, \; i = 1, 2, 3 \ldots M \quad \text{——— (5)}$$

from Equation ⑩ get the N components of the observation vector x, $x_i = 1, 2, 3, \ldots N$ that can be used to get the estimate $\hat{m}$ of the ith transmitted symbol m_i

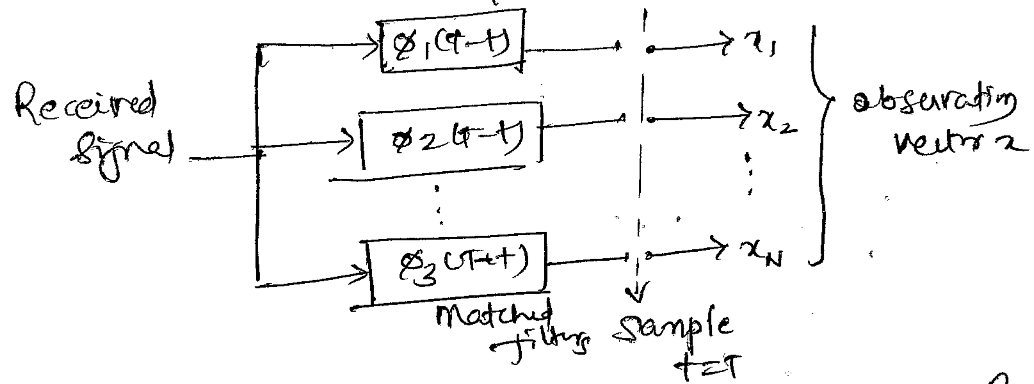
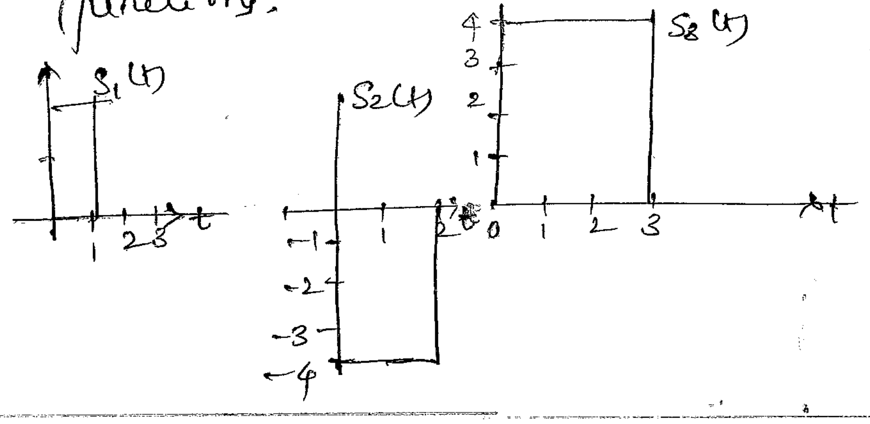


fig (c) Detector part of matched filter Receiver

A filter whose response is a time reversed and delayed version of its input signal as in Equation ⑨ is known as a filter matched to $\phi_j(t)$. optimum detector using a matched filter is called matched filter receiver. for the matched to be physically realisable, it must be causal that is its impulse response $h_i(t)=0$, for $t<0$. The causality condition is automatically satisfied since $\phi_j(t)=0$ outside $0 \le t \le T$.

3(b) Diagram 4
+ Explanation 4
+ Equations 2
10M

4(a) Using the Gram Schmidt Orthogonalization procedure find a set of orthonormal basis functions to represent the three signals $S_1(t)$, $S_2(t)$ and $S_3(t)$ shown in fig Q.4(a). Also express each of these signals interms of the set of basis functions.



$$E_1 = \int_0^1 2^2\,dt = 4$$

$$\phi_1(t) = \frac{s_1(t)}{\sqrt{E_1}} = \frac{s_1(t)}{\sqrt{4}} = \frac{2}{2} = 1$$

$$s_{21}(t) = \int_0^T s_2(t)\,\phi_1(t)\cdot dt$$

$$= \int_0^1 (-4)(1)\cdot dt = 4$$

$$g_2 = s_2(t) - s_{21}(t)\,\phi_1(t)$$ 02M

$$= \begin{cases} -4 & \text{for } 1 \le t \le 2 \\ 0 & \text{otherwise} \end{cases}$$

$$E_{g_2} = \int_1^2 (-4)^2\cdot dt = 16$$

$$\phi_2(t) = \frac{g_2(t)}{\sqrt{E_{g_2}}} = \frac{g_2(t)}{\sqrt{16}} = \frac{-4}{4} = \begin{cases} -1 & 1 \le t \le 2 \\ 0 & \text{otherwise} \end{cases}$$

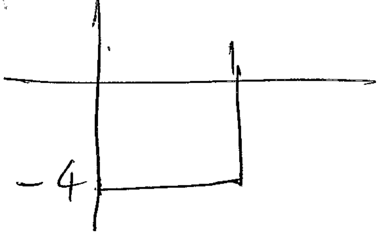
$$s_{31} = \int_0^T s_3(t)\cdot\phi_1(t) = \int_0^1 3(1)\,dt = 3$$ + 2M

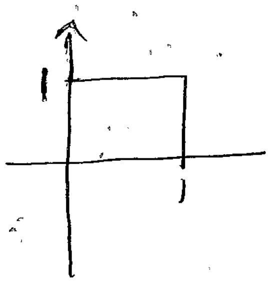
$$s_{32} = \int_0^T s_3(t)\cdot\phi_2(t)$$

$$= \int_0^2 3(-1)\cdot dt = -3$$ + 2m

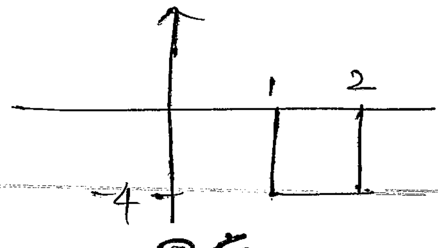
$$g_3(t) = g_3(t) - s_3\,\phi_1'(t) - s_{32}'\,\phi_2(t) = \begin{cases} 3 & \text{for } 2 \le t \le 3 \\ 0 & \text{otherwise} \end{cases}$$

$s_{21}\phi_1(t)$ + 2M



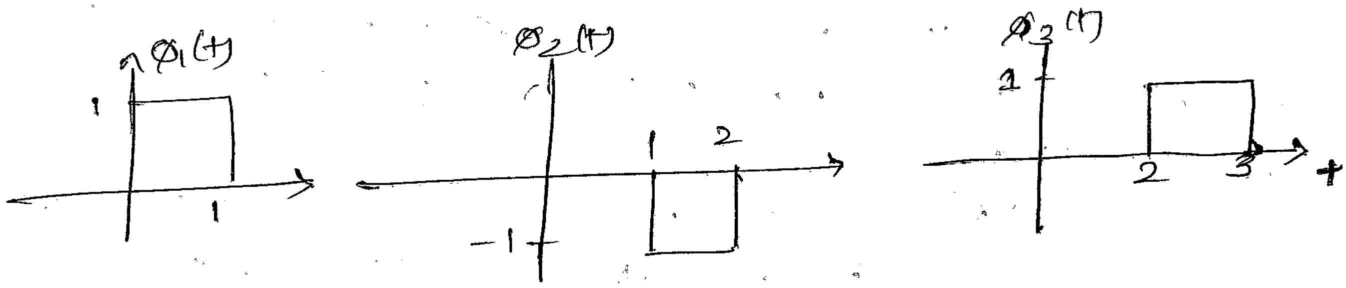


$s_2(t) - s_{21}(t)\,\phi_1(t)$



$$E_{g_3}{}^2 = \int_2^3 3^2 \, dt = 9$$

$$\phi_3(t) = \frac{g_3(t)}{\sqrt{E_{g_3}}} = \frac{g_3(t)}{\sqrt{9}} = \frac{3}{3} \begin{cases} 1 & \text{for } 2 \le t \le 3 \\ 0 & \text{otherwise} \end{cases}$$



+ 2m

$$S_1(t) = \sqrt{E_1}\,\phi_1(t) = 2\phi_1(t)$$

$$S_2(t) = S_{21}\phi_1(t) + \sqrt{E_{g_2}}\,\phi_2(t) = -4\phi_1(t) + 4\phi_2(t)$$

$$S_3(t) = S_{31}\phi_1(t) + S_{32}\phi_2(t) + \sqrt{E_{g_2}}\,\phi_3(t)$$ + 2m

$$= 3\phi_1(t) - 3\phi_2(t) + 3\phi_3(t)$$

12 m

4 (b) Show that for a noisy input, the mean value of the j^{th} correlator output x depends on S_i, and all the correlator output x_j, $j=1,2,\dots N$ have a variance equal to the PSD $N_0/2$ of the additive process $w(t)$

Let $X(t)$ denote the stochastic process and let $x(t)$ is sample function of stochastic process $X(t)$.

Let X_j denote the random variable whose sample value is represented by the correlator outputs x_j ∀ $j=1,2,3 \dots N$.

According to AWGN model, the stochastic process $X(t)$ is a Gaussian process.

X_j is a Gaussian random variable for all j in accordance with property 1, If a Gaussian process $X(t)$ is applied to a stable linear filter then the stochastic process $Y(t)$ developed at the output of the filter is also Gaussian,

X_j is characterized completely by its mean & variance

Let W_j denote the random variable represented by the sample value w_j produced by the j^{th} correlator in response to white Gaussian noise component $w(t)$. The random variable W_j has zero mean because the channel noise process $W(t)$ represented by $w(t)$ in the AWGN model has zero mean by definition.

Mean X_j depends only on S_{ij} and shown by

$$\begin{aligned}\mu_{x_j} &= E[X_j] \\ &= E[S_{ij} + W_j] \\ &= S_{ij} + E[W_j] \quad \text{——— (1)} \\ &= S_{ij}\end{aligned}$$

4(b) Explanation 3 + Equation + 5 Derivation = 8M

To find variance of X_j:

$$\sigma_{x_j}^2 = Var[X_j]$$
$$= E[(x_j - s_{ij})^2]$$
$$= E[w_j^2] \quad \text{—— (2)} \qquad \because x_j = s_{ij} + w_j$$

x_j & w_j are replaced by X_j & W_j

Random variable W_j is defined by,

$$W_j = \int_0^T w(t)\,\phi_j(t)\,dt \quad \text{—— (3)}$$

Use Equation (3) in Equation (2) then

$$\sigma_{x_j}^2 = E \int_0^T w(t)\cdot\phi_j(t)\,dt \int_0^T w(u)\,\phi_j(u)\,du$$
$$= E\left[\int_0^T \int_0^T \phi_j(t)\,\phi_j(u)\; w(t)\,w(u)\,dt\,du\right] \quad \text{—— (4)}$$

Interchange the order of summation & expectation

$$\sigma_{x_j}^2 = \int_0^T \int_0^T \phi_j(t)\cdot\phi_j(u)\,E[w(t)\cdot w(u)]\,dt\,du$$
$$= \int_0^T \int_0^T \phi_j(t)\cdot\phi_j(u)\cdot R_w(t,u)\,dt\,du \quad \text{—— (5)}$$

where $R_w(t,u)$ auto correlation function of the noise process $w(t)$.

Since noise is stationary, $R_w(t,u)$ depends only on the time difference $t-u$.

Gaussian noise with $PSD = N_0/2$

$$R_w(t,u) = \left(\frac{N_0}{2}\right)\delta(t-u) \quad \text{—— (6)}$$

Put Eq (6) in (5), using shifting property of delta function $\delta(t)$

$$\sigma_{x_j}^2 = \frac{N_0}{2}\int_0^T \int_0^T \phi_j(t)\cdot\phi_j(u)\,\delta(t-u)\,dt\,du$$
$$= \frac{N_0}{2}\int_0^T \phi_j^2(t)\,dt$$

$\phi_j(t)$ has unit energy, $\sigma_{x_j}^2 = \frac{N_0}{2}$ for all j —— (7)

All the correlator outputs denoted by x_j with $j = 1, 2, \ldots N$, have a variance equal to the $PSD = N_0/2$ of the noise process $w(t)$.

5 (a) Derive the expression for error probability of binary PSK using coherent detection.

The two basic components in the binary PSK receiver are correlator and Decision Device.

Correlator correlates the received signal $x(t)$ with the basis function $\phi_1(t)$ on a bit by bit basis. The decision device compares the output against a zero threshold assuming that binary symbols 1 and 0 are equiprobable. If the threshold is exceeded a decision is made in favour of symbol 1, if not decision is made in favour of symbol 0.

Partition the signal space into two regions.

- set of points closest to message point 1 at $+\sqrt{E_b}$
- set of points closest to message point 2 at $-\sqrt{E_b}$.

Two kinds of decisions,

1. Error of the first kind: Signal $s_2(t)$ is transmitted but the noise is such that the received signal point falls inside region Z_1, so the receiver decides in favour of signal $s_1(t)$.
2. Error of the second kind: Signal $s_1(t)$ is transmitted but the noise is such that the received signal point falls inside region Z_2, so the receiver decides in favour of signal $s_2(t)$.

To calculate the probability of making an error of the first kind, the decision region associated with symbol 1 or signal $s_1(t)$ described by,

$x_1 : 0 < x_1 < \infty$

observable element x_1 is related to the received signal $x(t)$ by

$$x_1 = \int_0^{T_b} x(t) \cdot \phi_1(t) \cdot dt \quad \text{——— (1)}$$

Conditional probability density function of random variable x_1, given that symbol 0 (ie signal $S_2(t)$) was transmitted,

$$f_{x_1}(x_1|0) = \frac{1}{\sqrt{\pi N_0}} \exp\left[-\frac{1}{N_0}(x_1 - S_{21})^2\right] \quad \text{——— (A)}$$

We know,

$$S_{21} = \int_0^{T_b} S_2(t) \cdot \phi_1(t) \cdot dt$$

$$= -\sqrt{E_b} \quad \text{——— (2)}$$

Use this Equation (2) in (A)

$$f_{x_1}(x_1|0) = \frac{1}{\sqrt{\pi N_0}} \exp\left[-\frac{1}{N_0}(x_1 + \sqrt{E_b})^2\right]$$

Conditional probability of the receiver deciding in favour of symbol 1, given that symbol 0 was transmitted.

$$P_{10} = \frac{1}{\sqrt{\pi N_0}} \int_0^{\infty} \exp\left[-\frac{1}{N_0}(x_1 + \sqrt{E_b})^2\right] \cdot dx_1 \quad \text{——— (B)}$$

Putting,

$$z = \sqrt{\frac{2}{N_0}}(x_1 + \sqrt{E_b})$$

and changing the variable of integration from x_1 to z

Rewrite eq. (B) in terms of Q function

$$P_{10} = \frac{1}{\sqrt{2\pi}} \int_{\sqrt{2E_b/N_0}}^{\infty} \exp\left(-\frac{z^2}{2}\right) \cdot dz \quad \text{——— (C)}$$

We know, $Q(x) = 1 - F_x(x)$

Q function

$$= \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp\left(-\frac{t^2}{2}\right) \cdot dt \quad \text{——— (a)}$$

Using ⓐ & ⓒ

$$P_{10} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

P_{01}, the conditional probability of the receiver deciding in favour of symbol 0, given that symbol 1 was transmitted.

Averaging conditional error probabilities P_{10} and P_{01}, average probability of symbol error or equivalently, the BER for binary PSK using coherent detection & assuming equiprobable symbols is given by

$$P_e = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

5ⓐ Equations/Derivation 3 + Explanation + 3 = 6 m

5.ⓑ Explain the generation and optimum detection of differential phase shift keying with neat diagram

Generation of DPSK signal

Transmitter consists of two functional blocks

* logic network and one bit delay (storage) element, which are interconnected so as to convert the raw input binary sequence $\{b_k\}$ into the differentially encoded sequence $\{d_k\}$
* Binary PSK modulator, the output of which is the desired DPSK signal.

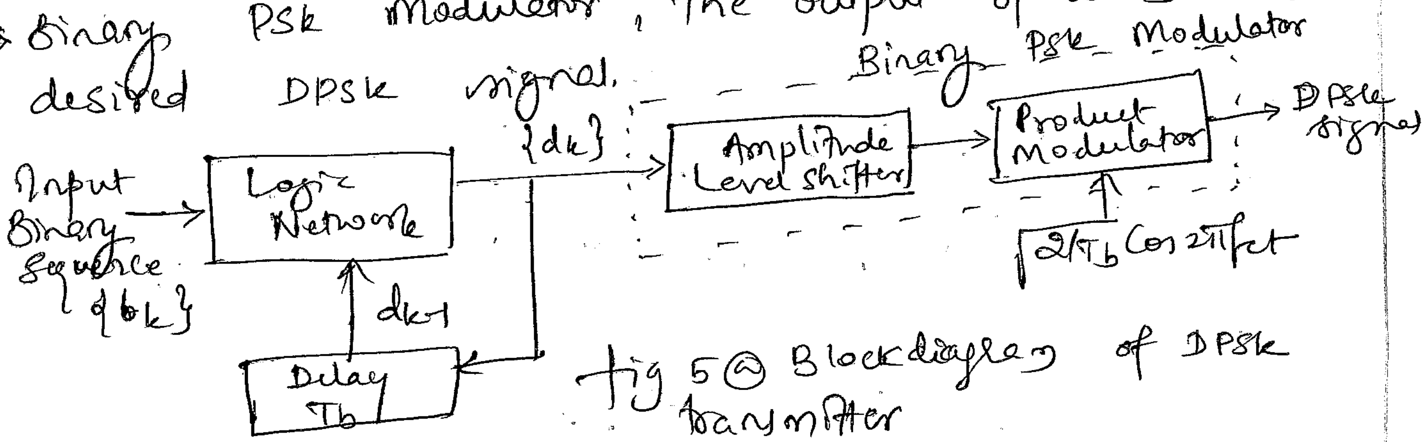


fig 5ⓐ Block diagram of DPSK transmitter

DPSK Receiver :-

Receiver of DPSK has in phase and quadrature paths. In signal space diagram the received signal points over the two bit interval $0 \le t \le 2T_b$ are defined by $(A\cos\theta, A\sin\theta)$ and $(-A\cos\theta, -A\sin\theta)$, A denotes the carrier amplitude.

For two bit interval $0 \le t \le 2T_b$, the receiver measures the co-ordinates x_{I_0}, x_{Q_0} first at time $t = T_b$ and then measures x_{I_1}, x_{Q_1} at time $t = 2T_b$

$$x_{I_0} x_{I_1} + x_{Q_0} x_{Q_1} \underset{\text{Say } 0}{\overset{\text{Say } 1}{\gtrless}} 0 \quad \text{——— (1)}$$

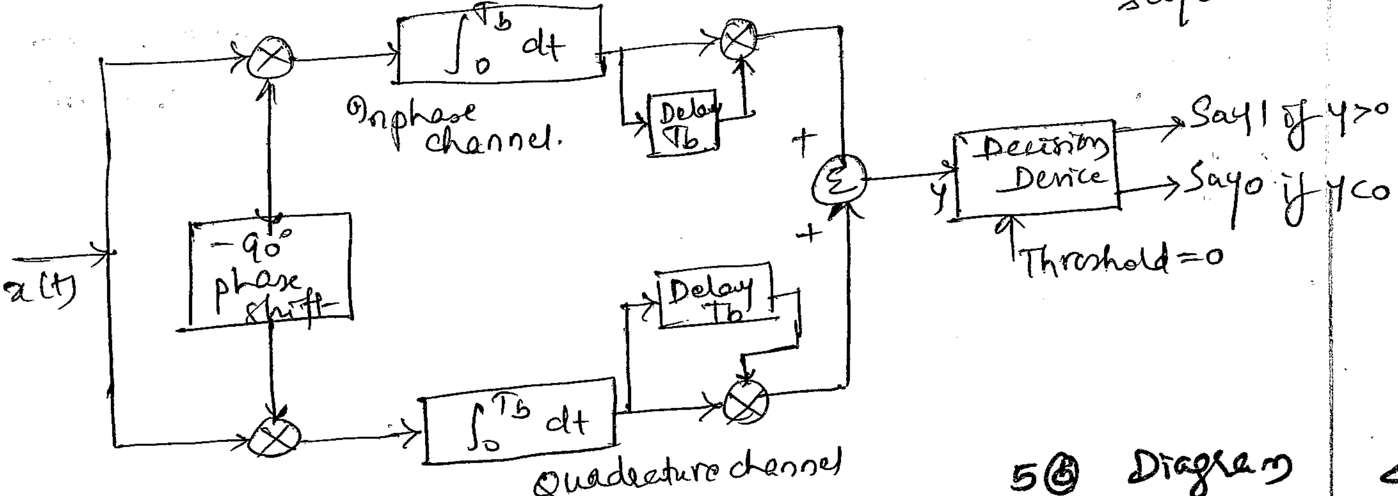
where the threshold is zero for equiprobable symbols.

$$x_{I_0} x_{I_1} + x_{Q_0} x_{Q_1} = \frac{1}{4}\left((x_{I_0} + x_{I_1})^2 - (x_{I_0} - x_{I_1})^2 + (x_{Q_0} + x_{Q_1})^2 - (x_{Q_0} - x_{Q_1})^2\right) \quad \text{——— (2)}$$

substitute (2) in (1)

Equivalent test is

$$(x_{I_0} + x_{I_1})^2 + (x_{Q_0} + x_{Q_1})^2 - (x_{I_0} - x_{I_1})^2 - (x_{Q_0} - x_{Q_1})^2 \underset{\text{Say } 0}{\overset{\text{Say } 1}{\gtrless}} 0$$



5(b) Diagram 4
+ Explanation 4

8M

5(c) A binary data is transmitted over a microwave link at a rate of 10^6 bits/sec and the PSD of noise at the receiver is 10^{-10} watts/Hz. Find the average carrier power required to maintain an average probability of error $P_e \leq 10^{-4}$ for coherent binary FSK. What is the required channel bandwidth? Given ($erf(2.6) = 0.9998$)

Data Rate $= \frac{1}{T_b} = 10^6$ bits/sec

$$P_e = \frac{1}{2} erfc\sqrt{\frac{0.6E}{N_0}}$$

$$10^{-4} = \frac{1}{2} \cdot erfc\sqrt{\frac{0.6E}{N_0}}$$

$$\therefore erfc\sqrt{\frac{0.6E}{N_0}} = 2\times10^{-4}$$

$$1-erf\sqrt{\frac{0.6E}{N_0}} = 1-2*10^{-4} = 0.9998$$

$$erf\sqrt{\frac{0.6E}{N_0}} = 0.9998$$

PSD of Noise $= N_0/2$

$= 10^{-10}$ watts/Hz

$P_e \leq 10^{-4}$

$1-erfc(4) = erfc(4)$

$$\sqrt{\frac{0.6E}{N_0}} = 2.8$$

$$\frac{0.6E}{N_0} = (2.8)^2 = 7.84$$

$$\frac{N_0}{2} = 10^{-10} = N_0 = 2*10^{-10}$$

$$E = \frac{7.84 * 2*10^{-10}}{0.6} = 2.61 * 10^{-9} \text{ Joules}$$

$E = PT_b$ $\qquad P = \frac{E}{T_b} = 2.61 * 10^{-9} * 10^6$

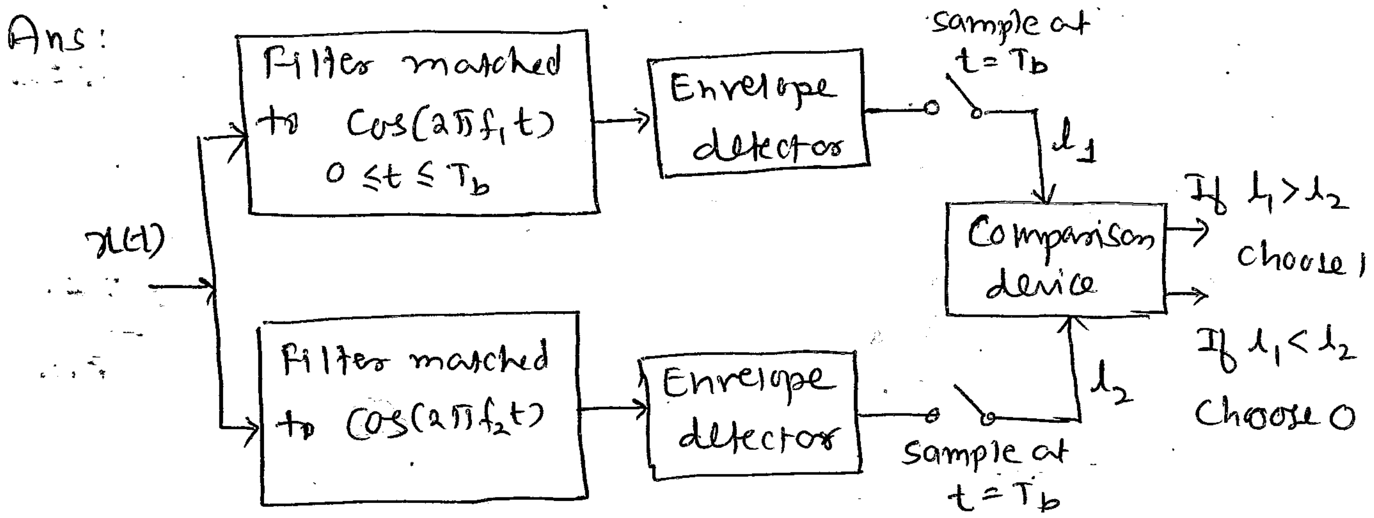
$= 2.61$ mW.

$B_T = \frac{1}{T_b} = 10^6$ bits/sec $= 1$ MHz

Data	1
+ Formula	2
+ Solution	3
	6M

Q.6 a) With neat diagram explain non coherent detection FSK.

Ans:



→ In Binary FSK transmitted signal is defined as

$$S_i(t) = \begin{cases} \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_i t) & 0 \le t \le T_b \quad i = 1, 2 \\ 0 & \text{elsewhere} \end{cases}$$

→ T_b - bit duration, carrier freq $f_i = f_1$ and f_2 for $i = 1, 2$.

→ To ensure f_1 & f_2 are orthogonal, choose $f_i = n_i/T_b$

n_i - is an integer

→ f_1 represents symbol '1' & f_2 represents symbol '0'.

→ Here the receiver consists of a pair of matched filters followed by envelope detectors as shown in above fig

→ The filter in upper path of the receiver is matched to $\cos 2\pi f_1 t$ & similarly in lower path it matches to $\cos 2\pi f_2 t$ for the signaling interval of $0 \le t \le T_b$.

→ The o/p of envelope detectors are sampled at $t = T_b$ & their values are compared.

→ The receiver decides in favour of symbol 1 if $l_1 > l_2$ and in favour of symbol 0 if $l_1 < l_2$.

Diagram 3 + Explanation +5 = 8M

Q.6 b) $A_c = \sqrt{2P}$ $\therefore \frac{A_c^2}{2} = P$ $\therefore P = \frac{(1\times10^{-6})^2}{2} = 0.5\times10^{-12}\,W$ 02

$T = \frac{1}{\text{datarate}} = \frac{1}{2.5\times10^{+6}} = 4\times10^{-7}$ sec + 02

$\frac{N_0}{2} = 10^{-20}$ w/sec $\therefore N_0 = 2\times10^{-20}$ + 02

$P_e = \frac{1}{2}\,\text{erfc}\sqrt{\frac{E_b}{2N_0}}$ $E_b = PT_b$
$= 0.5\times10^{-12}\times4\times10^{-7} = 2\times10^{-18}$ Joules

$\therefore P_e = \frac{1}{2}\,\text{erfc}\sqrt{\frac{0.2\times10^{-18}}{4\times10^{-20}}}$ + 02

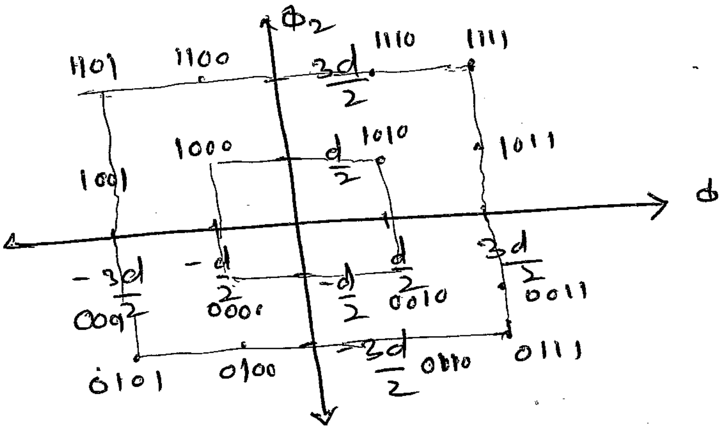
$= \frac{1}{2}\,\text{erfc}\,(2.236)$ 8M

$P_e = 2.05\times10^{-4}$

Q.6 c) → M-ary QAM does phase modulation and amplitude modulation of the carrier signal.

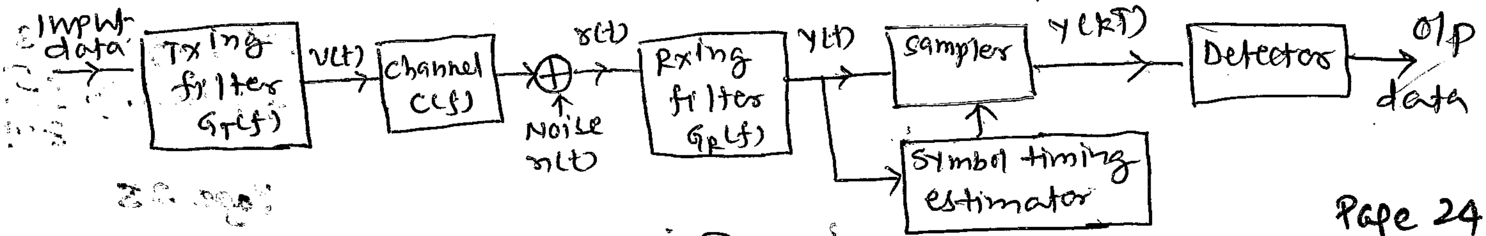
→ The amplitude constraint of M-ary PSK is removed in M-ary QAM.

→ signal constellation diagram of M-ary PAM is one dimensional where as in M-ary QAM it is two dimensional



6C Diagram 02
+ Explanation + 02
4M

Q.7 a) Digital PAM transmission through band limited ~~Channel~~ base band channels.



7a) The system consisting of transmitting filter having an impulse response of $g_T(t)$, the linear filter channel with AWGN, a receiving filter with an impulse response of $g_R(t)$, a sampler that periodically samples the output of the receiving filter, and a symbol detector.

→ The sampler requires the extraction of a timing signal from the received signal

→ The timing signal serves as a clock that specifies the appropriate time instants for sampling the output of the receiving filter

→ Here the digital communication takes place by means of M-ary PAM

→ Hence the input binary data sequence is subdivided into k-bit symbols, and each symbol is mapped into corresponding amplitude level that amplitude modulates the output of the transmitting filter

→ The baseband signal at the o/p of Txing filter is

$$v(t) = \sum_{n=-\infty}^{\infty} a_n g_T(t-nT)$$

where $T = \frac{k}{R_b}$ is the symbol interval & $R_b = \frac{k}{T_b}$ is the symbol rate or bit rate

a_n - sequence of amplitude levels corresponding to the sequence of k-bit blocks of information bits.

→ The channel output

$$r(t) = \sum_{n=-\infty}^{\infty} a_n h(t-nT) + n(t)$$

where $h(t)$ – impulse response of the cascade of Txing filter & channel

7a) Diagram 4
+ Explanation 4
8m

Joshua

$\therefore h(t) = c(t) * g_T(t)$

→ The received Signal is passed through a linear receiving filter with the impulse response $g_R(t)$ & frequency response of $G_R(f)$.

→ If $g_R(t)$ matched to $h(t)$, then its SNR is a maximum at the proper Sampling instants.

⇒ The output of the receiving filter is

$$y(t) = \sum_{n=-\infty}^{\infty} a_n x(t - nT) + w(t)$$

where $x(t) = h(t) * g_R(t)$

$= g_T(t) * c(t) * g_R(t)$

$w(t) = n(t) * g_R(t)$

⇒ To recover original signal (information symbols) $\{a_n\}$, the output of the Rxing filter is sampled periodically, every T seconds.

$\therefore$ The sampler produces

$$y(mT) = \sum_{n=-\infty}^{\infty} a_n x(mT - nT) + w(mT)$$

or equivalently

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

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

where $x_m = x(mT)$, $w_m = w(mT)$ & $m = 0, \pm 1, \pm 2 \ldots$

→ The first term $x_0 a_m$ is the desired symbol scaled by the gain parameter x_0

→ Second term $\sum_{n \neq m} a_n x_{m-n}$ represents the effect of the other symbols at the sampling instants $t = mT$, called ISI

7 b) State Nyquist Condition for Zero I.S.I

A necessary & sufficient condition for signal $x(t)$ i.e. output of the receiving filter to satisfy

$$x(nT) = \begin{cases} 1 & \text{for } n = 0 \\ 0 & \text{for } n \neq 0 \end{cases}$$ is that its FT

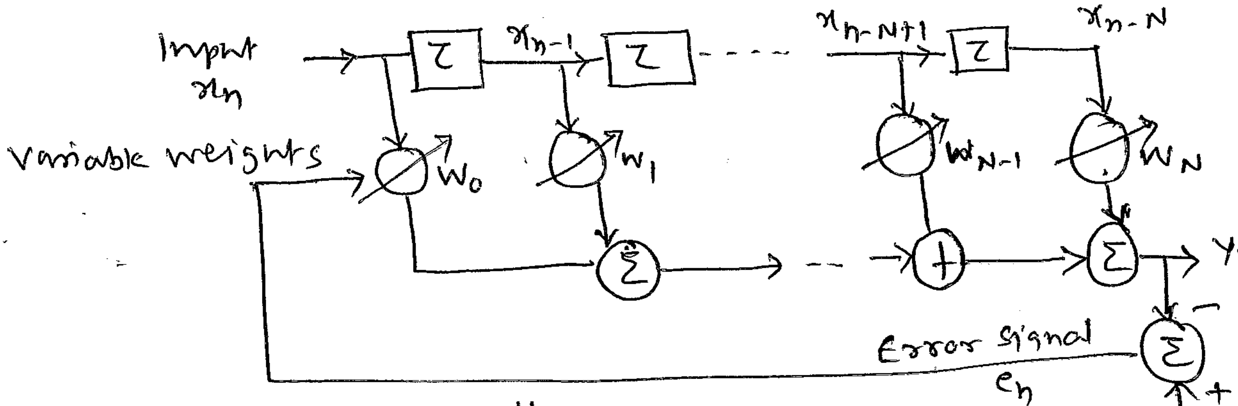
7b) 8M

$\therefore X(f)$ must satisfy

$$\sum_{m=-\infty}^{m=\infty} X\left(f + \frac{m}{T}\right) = T$$

7c) Adaptive equalization

7c) Diagram + Explanation 4M



$$y_n = \sum_{k=0}^{N} w_k x_{n-k}$$

$$e_n = a_n - y_n$$
$$= a_n - \sum_{k=0}^{N} \hat{w}_{k,n} x_{n-k}$$

→ The adaptive equalizer works in two modes

i) Training mode – a known PN sequence is Txed and a synchronized version of it is generated in the receiver, where it is applied to the adaptive equalizer, as the desired response, the tap weights are thereby adjusted in accordance with LMS algorithm.

ii) After completion of training mode equalizer is switched to its second mode of operation the decision-directed mode.

7c) In this mode of operation

$$e_n = \hat{a}_n - y_n$$

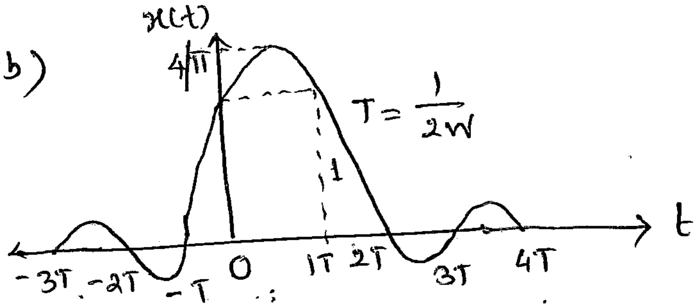
where y_n - equalizer o/p at time $t = nT$ & $\hat{a}_n$ is the final correct estimate of the transmitted symbol $\hat{a}_n$.

8a) steps $\frac{8 \times 2m}{6m}$

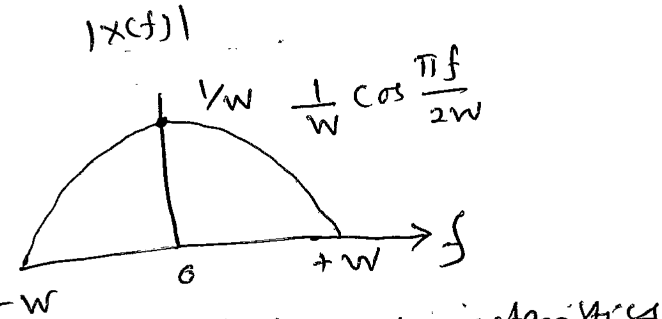
8a) $\{d_n\}$		1	1	1	0	1	0	0	1
precoded sequence $\{P_n\}$	0	1	0	1	1	0	0	0	1
Transmitted sequence $\{a_n\}$	-1	1	-1	1	1	-1	-1	-1	1
Received seq $\{b_n\}$		0	0	0	2	0	-2	-2	0
Decoded seq $\{d_n\}$		1	1	1	0	1	0	0	1

$$\left.\begin{array}{l} d_n = 1 \text{ if } b_n = 0 \\ d_n = 0 \text{ if } b_n = \pm 2 \end{array}\right\}$$

8b)



Time domain characteristics of duobinary signal



frequency domain characteristics of duobinary signal

one special case that leads to (approximately) physically realizable transmitting and receiving filters is specified by the samples

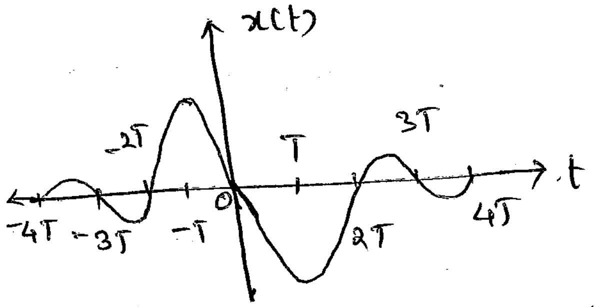
$$x(nT) = \begin{cases} 1 & n = 0, 1 \\ 0 & \text{elsewhere} \end{cases}$$

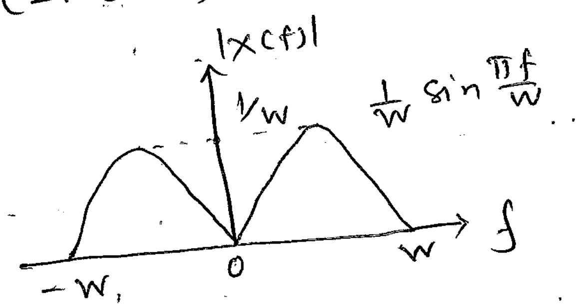
8 b)

$$x(f) = \begin{cases} \frac{1}{2W}\left[1+e^{-j\pi f/W}\right] & |f| < W \\ 0 & \text{otherwise} \end{cases}$$

$$= \begin{cases} \frac{1}{W} e^{-j\pi f/2W} \cos\left(\frac{\pi f}{2W}\right) & |f| < W \\ 0 & \text{otherwise} \end{cases}$$

$\therefore\ x(t) = \text{sinc}(2Wt) + \text{sinc}(2Wt - 1)$





Time domain characteristics of a modified duobinary signal

freq domain chara.

8 b) Diagram + 4
+ Explanation 4
8M

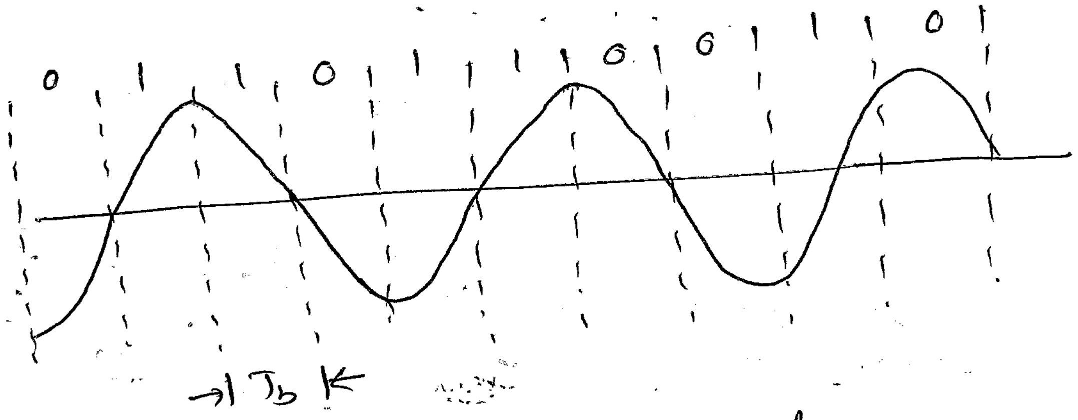
$$x(nT) = \begin{cases} 1 & n = -1 \\ -1 & n = 1 \\ 0 & \text{otherwise} \end{cases}$$

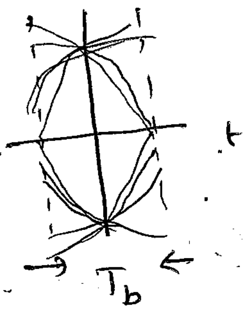
$x(t) = \text{sinc}(t+T)/T - \text{sinc}(t-T)/T$

and its spectrum

$$x(f) = \begin{cases} \frac{1}{2W}\left[e^{j\pi f/W} - e^{-j\pi f/W}\right] = \frac{1}{W}\sin\frac{\pi f}{W} & |f| \leq W \\ 0 & |f| > W \end{cases}$$

8c)

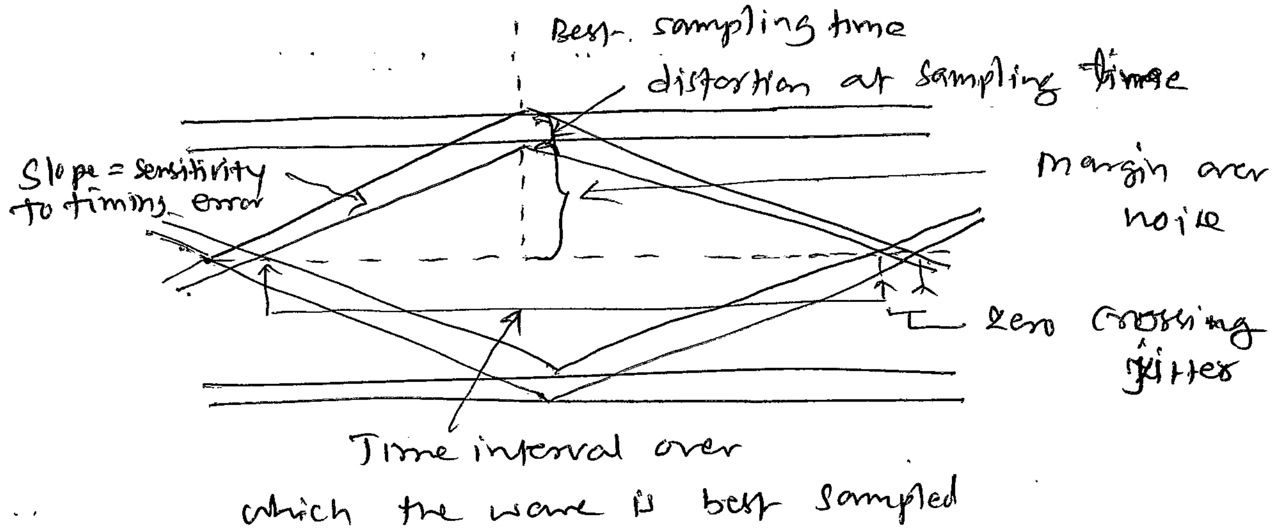




Binary data sequence & its waveform

Eye pattern

8c) Diagram + 3
+ Explanation + 3
6M



Timing features.

i) optimum sampling time - It is the time at which the eye opening is at its widest.

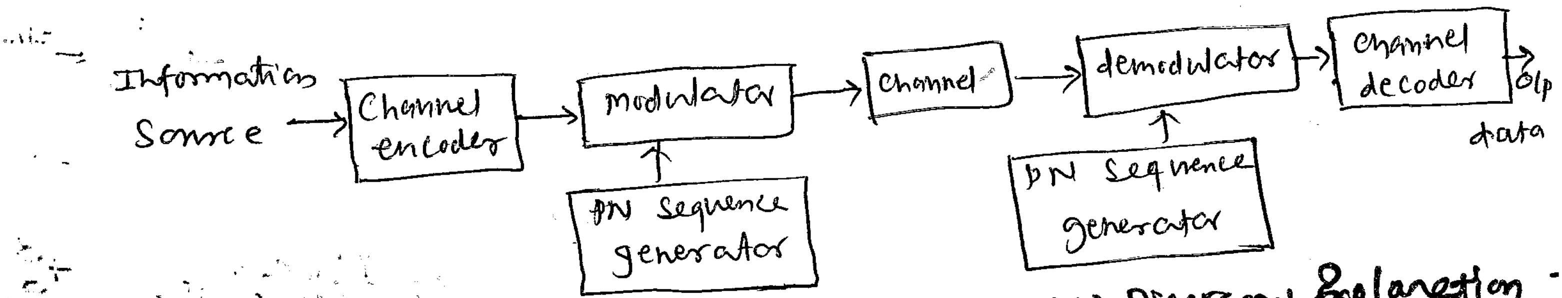
ii) Zero-crossing jitter - In practice, the timing signal is extracted from the zero-crossings of the waveform that appears at the Rx filter output. This is kind of irregularities

iii) Timing sensitivity - This is time related feature of the PAM sys. to timing errors. This is determined by the rate at which the eye pattern is closed as the sampling time is varied.

→ An eye opening of unity which corresponds to zero ISI

→ An eye opening of zero corresponds to a completely closed eye pattern : Here ISI effect is severe.

9 a) model of Spread spectrum digital Communication sys.



9a) Diagram + Explanation
3+5= 8 M

Page 30

9 a) The channel encoder & decoder & modulator & demodulator are the basic elements of a conventional digital comm. sys.

→ In addition spread spectrum system employs two identical PN sequence generators, one that interfaces with the modulator at the Tx-ing end and one that interfaces with the demodulator at the Rx-ing end.

→ The two generators produce a pseudorandom (PN) binary-valued sequence, which is used to spread the transmitted signal at the modulator & to despread the received signal at the demodulator.

→ Time synchronization of the PN sequence generated at the Rx with the PN sequence contained in the Rxed signal is reqd. ~~for~~ to properly despread the Rxed spread-spectrum signal.

9 b) Effect of despreading

Received signal $r(t) = A_c v(t) c(t) \cos 2\pi f_c t + i(t)$

Despreading $r(t)c(t) = A_c v(t) \cos 2\pi f_c t + i(t) c(t)$

The total power interfering the signal at the output of the demodulator is

9b. Equations + Explanation = 8M

$$\frac{P_I}{(W/R_b)} = \frac{P_I}{(T_b/T_c)} = \frac{P_I}{L_c}$$

∴ power is reduced by an amount L_c. L_c is called processing gain.

$$\frac{E_b}{N_0} = \frac{P_R T_b}{N_0} = \frac{P_R L_c T_c}{N_0} = \left(\frac{P_R}{W N_0}\right) L_c = \left(\frac{P_R}{P_N}\right) L_c$$

Joshy

$$\therefore \left(\frac{E_b}{N_o}\right) = \left(\frac{P_R}{P_N}\right) L_C$$

$$10 = 10^{-2} \times L_C \qquad \therefore L_C = 1000$$

9c) Def^n of PN sequence

Are class of sequences of 1^s and 0^s which are periodic and possesses autocorrelation property

→ Such PN sequences are generated by using shift registers having feedback connections.

→ Using an n-stage shift register having an appropriate linear feedback signals, it is possible to generate a periodic sequence with a period equal to $2^n - 1$ bits.

→ Such sequences are also called ML sequences.

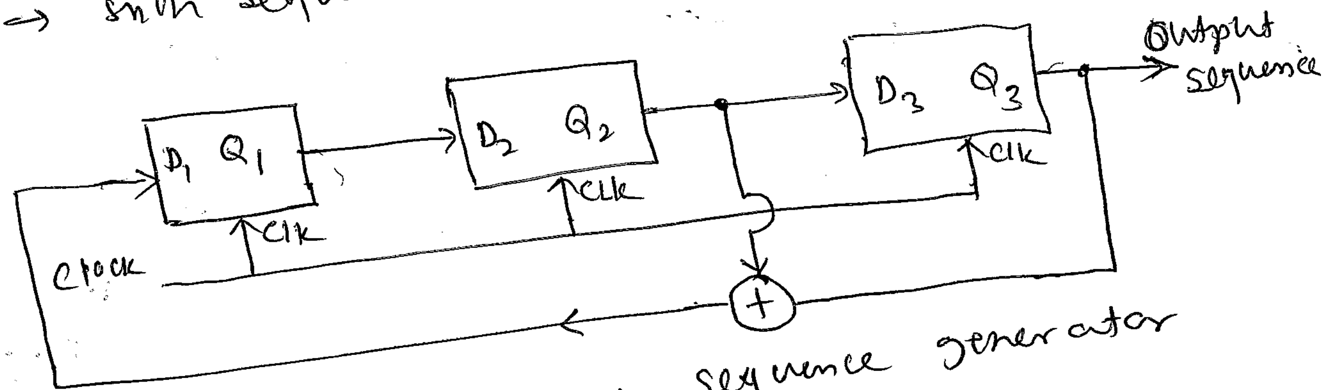


fig 3-stage ML sequence generator

→ An n-stage shift register consists of n clocked D-type F/F's

→ The o/p of the F/F after a clock pulse is equal to the input D of the F/F just before the clock pulse.

Def^n	1
9c) Diagram	2
+ Explanation	1
	4M

Joshi

9c) Linear feedback path performs the task of Ex-ORing at the output of the last stage of shift register and other selected outputs from intermediate stages and feeding the resultant o/p to the input of the first stage.

→ For every clock the contents of each stage of the register is shifted by one position to the right

→ Also, for every clock pulse the contents of stages 2 and 3 are modulo-2 added & the result is feedback to stage 1.

→ The shift register sequence is defined to be the output of the last stage - stage 3 in this context.

Properties of PN sequences (ML sequences)

I) Balance property
II) Run property
III) Auto correlation property

10 a) FHSS

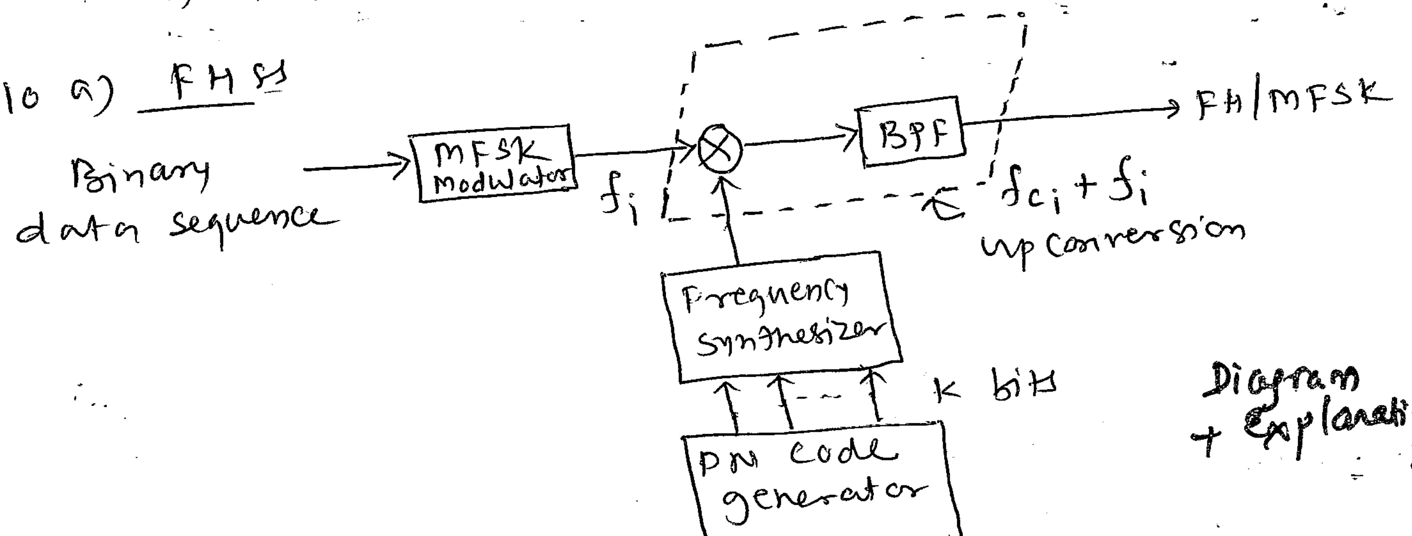


fig: FH/MFSK transmitter

Diagram 6
+ Explanation 4
10M

→ The first stage of an FH/MFSK transmitter is the frequency modulator and second stage is frequency mixer

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

→ Depending on k-bits binary pattern any one of the 2^k discrete amplitude levels of M-ary PAM signal is obtained & This signal is applied to VCO.

→ Therefore s(t) is an MFSK wave.

→ The o/p of the MFSK modulator is then mixed with o/p of the frequency synthesizer.

→ The frequency synthesizer's o/p is one of $Y = 2^k$ values, where k - equals number of bits of the PN sequence generator output. As a result, freq. hops over 2^k distinct values.

→ The BPF passes the sum freq for the transmission & rejects the difference frequency components.

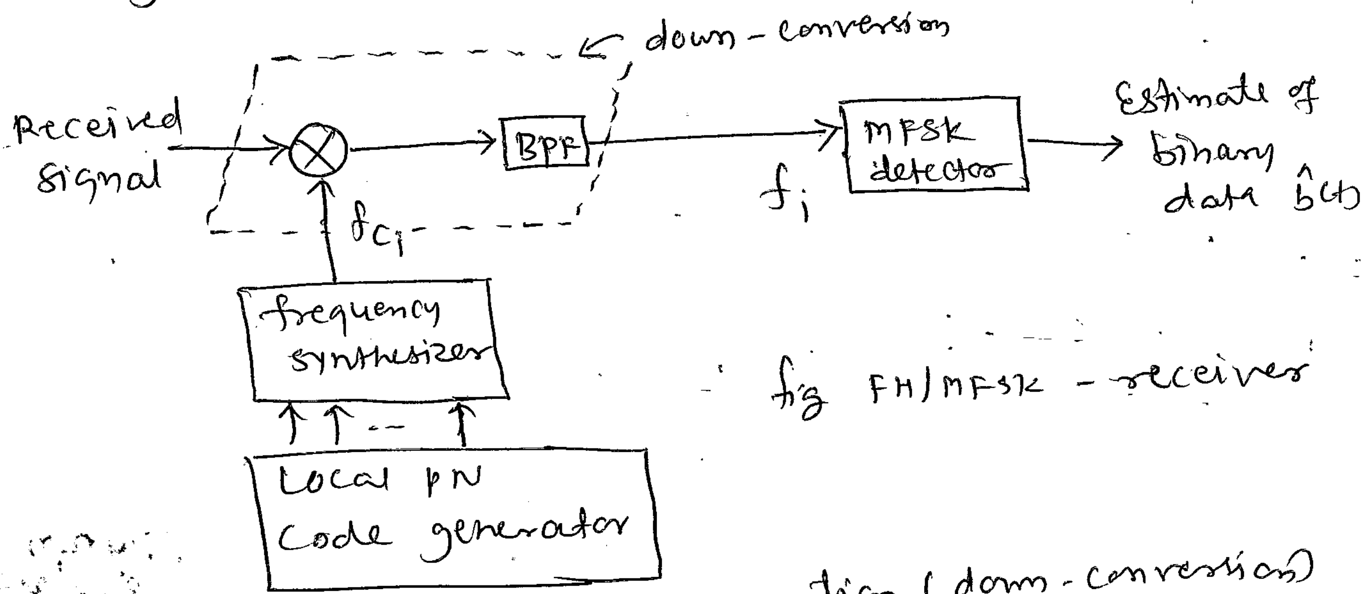


fig FH/MFSK - receiver

→ In the first-stage mixing operation (down-conversion) removes the frequency hopping

10 a) The mixer i/p's are the received signal & the output of a local frequency synthesizer that is in synchronization with that of transmitter.

→ The o/p of mixer is passed through a BPF which selects the difference frequency component from the mixer

→ The o/p of BPF is the MFSK signal, which is demodulated using non coherent MFSK detector.

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

→ The estimate of the original symbol transmitted is obtained by selecting the largest filter output

Define i) Chip rate $(R_c) = \frac{1}{T_c}$

where R_c – chip rate
T_c – chip interval

ii) Processing Gain

$$\frac{W}{R} = \frac{T_b}{T_c} = L_c$$

$$PG = \frac{\text{Bandwidth of Spread Signal}}{\text{Bandwidth of baseband Signal}}$$

iii) Jamming margin $\left(\frac{P_I}{P_s}\right)_{dB} = \left(\begin{matrix}\text{Processing}\\\text{Gain}\end{matrix}\right)_{dB} - \left(\frac{\text{Energy of signal/bit}}{\text{Noise}}\right)_{dB}$

$$\text{Jamming Margin} = \left(\begin{matrix}\text{Processing}\\\text{Gain}\end{matrix}\right)_{dB} - \left(\frac{E_b}{N_o}\right)_{dB}$$

Joshi

10b) CDMA tech. overcomes the problem of FDMA & TDMA means in situations where all the users are required to transmit simultaneously but also occupy the same RF bandwidth of the channel, CDMA can be used

→ The enhancement in performance obtained from a DSSS signal through the processing gain & coding gain can enable many DSSS signal simultaneously occupy the same channel BW provided that each signal has its own distinct PN sequence.

Forward link

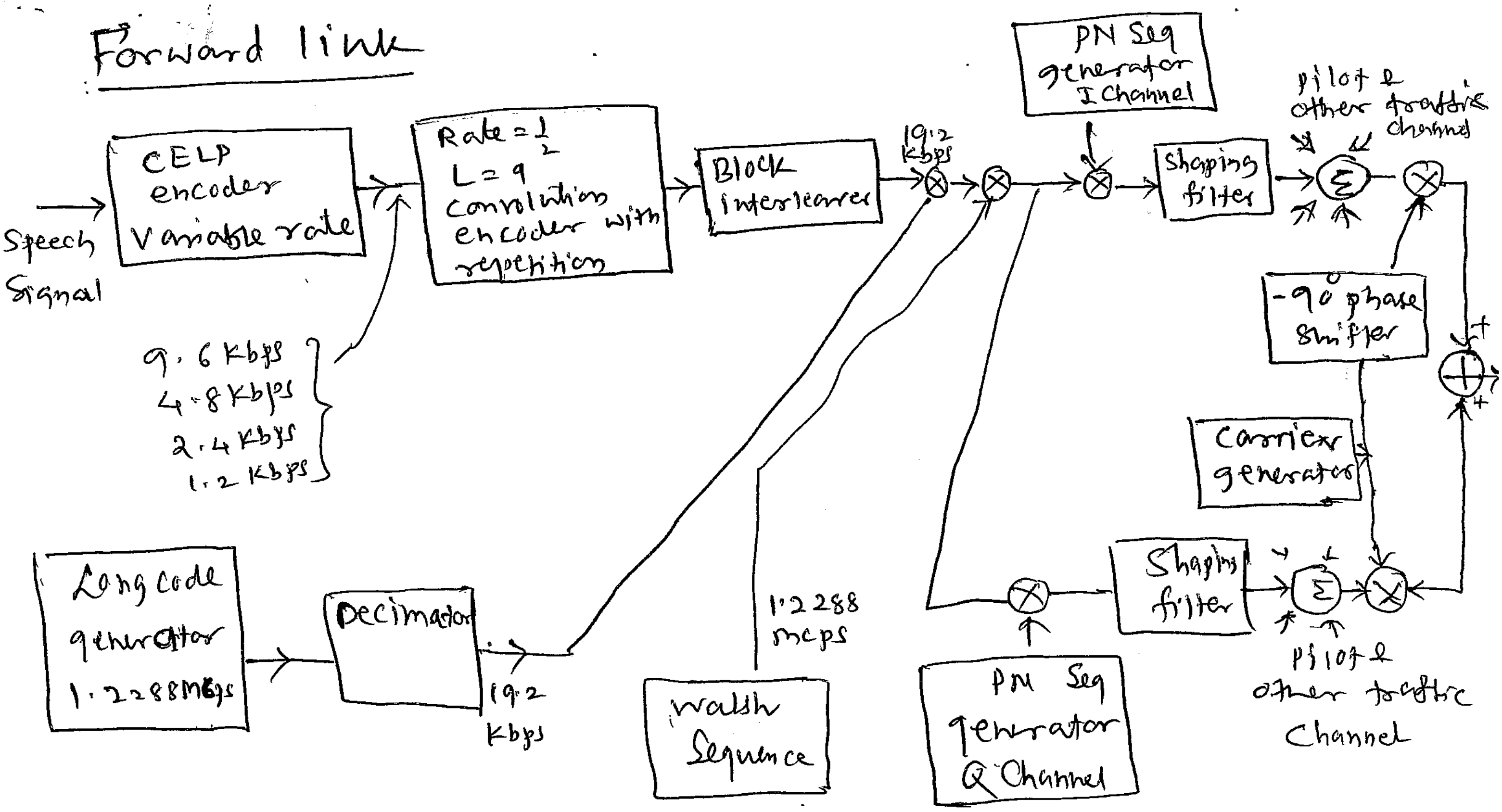


fig block diagram IS 95 forward link

→ Speech encoder code excited linear predictor generates data at variable rates of (9.6, 4.8, 2.4, 1.2) kbps in frame interval of 20msec

10b) Diagram + Explanation $\frac{5 + 5}{10M}$

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10 b)

→ Encoded speech is passed through rate $\frac{1}{2}$ constraint length of convolutional encoder and block interleaver

For lower speech data rates of (4.8, 2.4, 1.2) kbps

→ the o/p symbols of convolutional encoder is separated 2 times, 4 times, 8 times. Thus bit rate is fixed at 9.6 kbps

→ The o/p of block interleaver is at 19.2 kbps and is multiplied by long code of chip rate 1.2288×10^6 cps & decimated by factor 64

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

→ Each channel 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 chara. like signal strength, carrier phase & used in coherent detection.

→ one sequence is used for time synchronization & same sequences for paging. About remaining 64 channels are assigned to different users. Each user multiplies the data sequence by 64 bit walsh sequence

→ The resulting binary sequence is spread by multiplication with two PN sequences of length $N \cong 2^{15}$ creating inphase & quadrature signal components, resulting in 4 phase signal. Different base stations are identified by offsets of these PN sequences. The signal in all the 64 channels are time synchronized & are immune to interference from signals originating from other channels due to orthogonality of walsh sequences.

→ At the Rx a RAKE demodulator is used to separate the multipath components & combined before detection using Viterbi soft Decision.

Reverse link

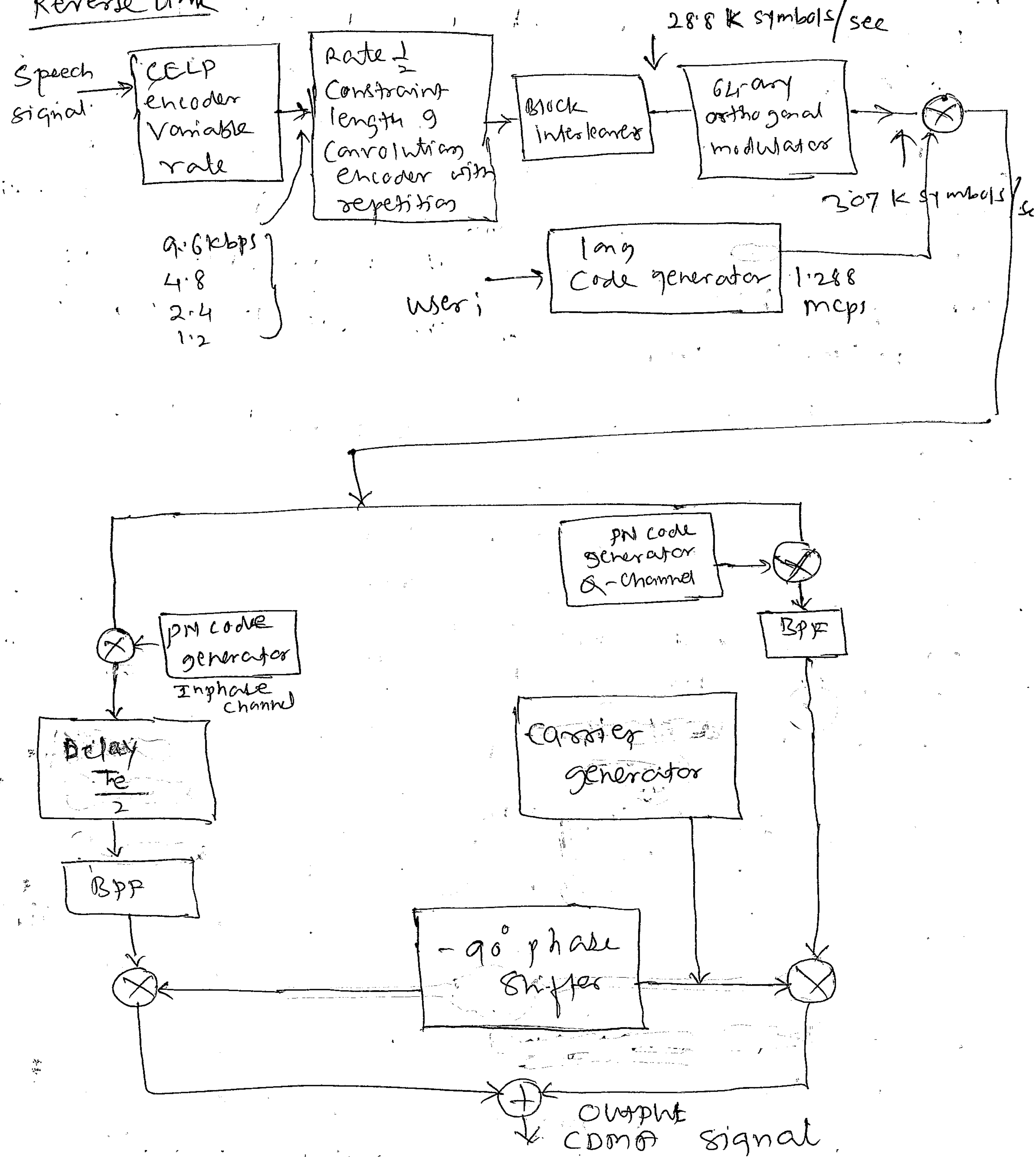

Fig: Block diagram of the IS 95: reverse link

→ The reverse link modulator from a mobile transmitter to a base station is different from the forward link modulator

10b)

→ The signals are Txed from various sources to the base stations are asynchronous resulting in interference from other users.

→ Battery operated mobile users (Tx^{ers}) result in a power limited transmissions

→ Effect of channel noise is reduced by using $\frac{1}{3}$ convolutional code in the constraint length 9 reverse link. Coded bit rate is 28.8 kbps.

→ For low rate speech output, bits from convolutional encoder are repeated approximately by 2/4/8 times.

→ As shown in fig of reverse link the Q channel is delayed by $\frac{T_c}{2}$ relative to I-channel signal producing an offset QPSK signal.

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