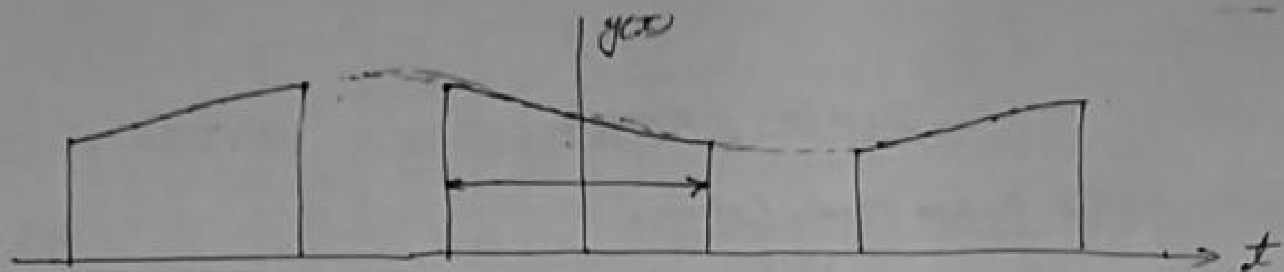
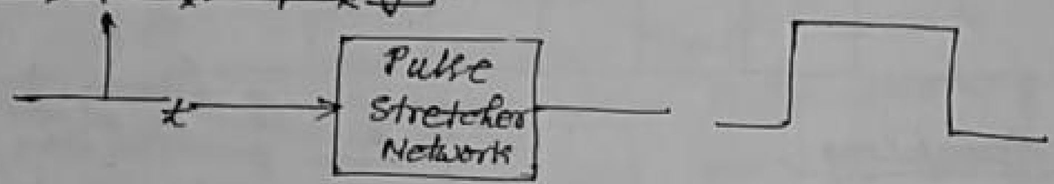
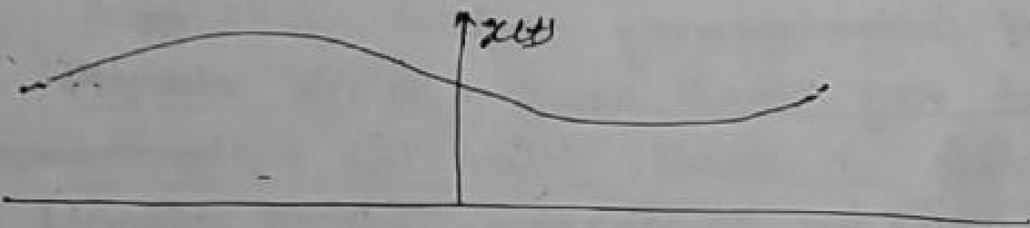


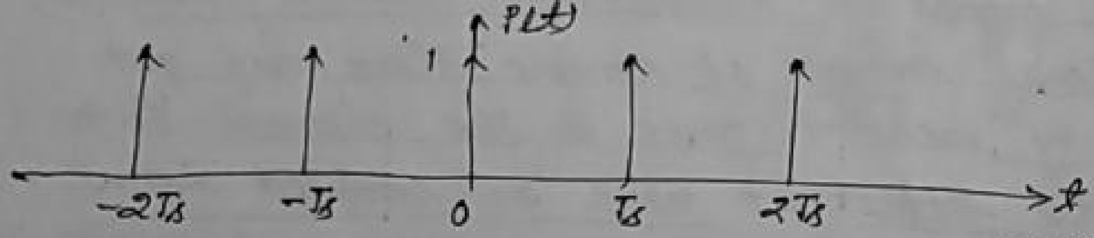
Disadvantage of Natural Sampling:

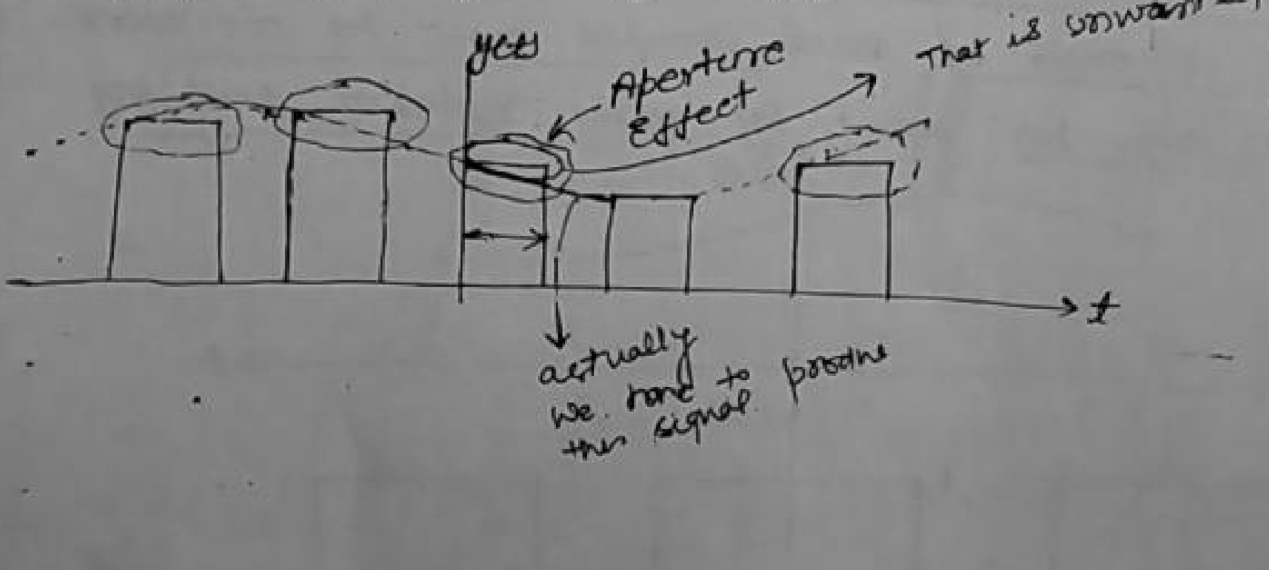
→ In natural sampling, the output amplitude is not constant throughout the pulse width but varies according to modulating signal. So the Tx has to provide variable amount of power throughout the pulse width which is not possible for any practical transmitter. That is why flat-top sampling is used for the transmission of signal.

// **Flat-Top Sampling:**



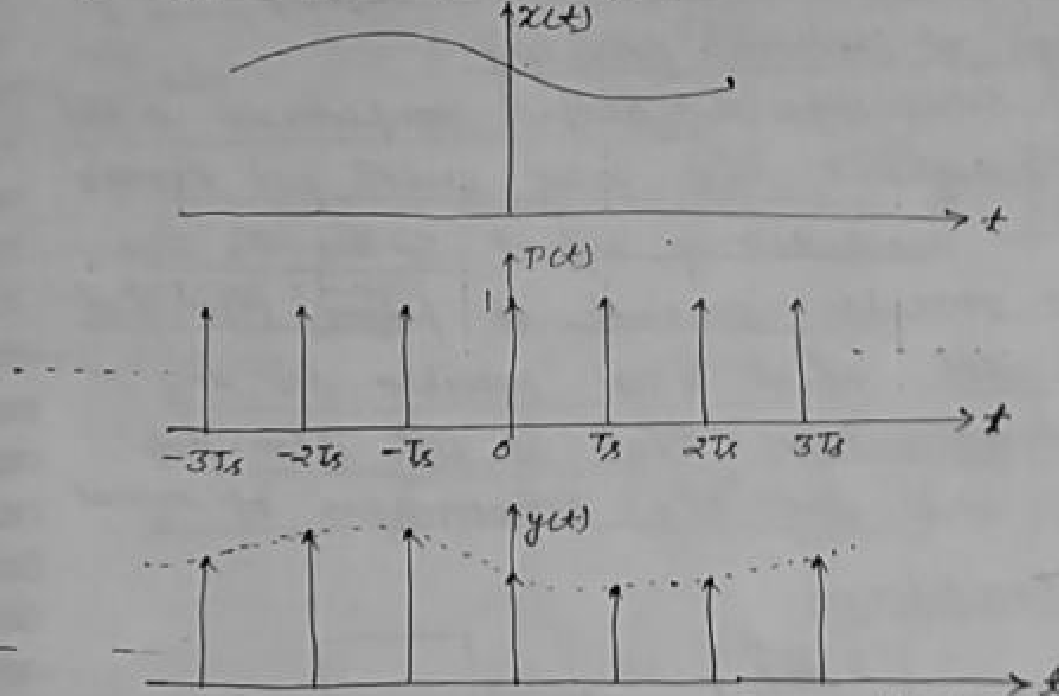






Digital Communication: 31/5/10

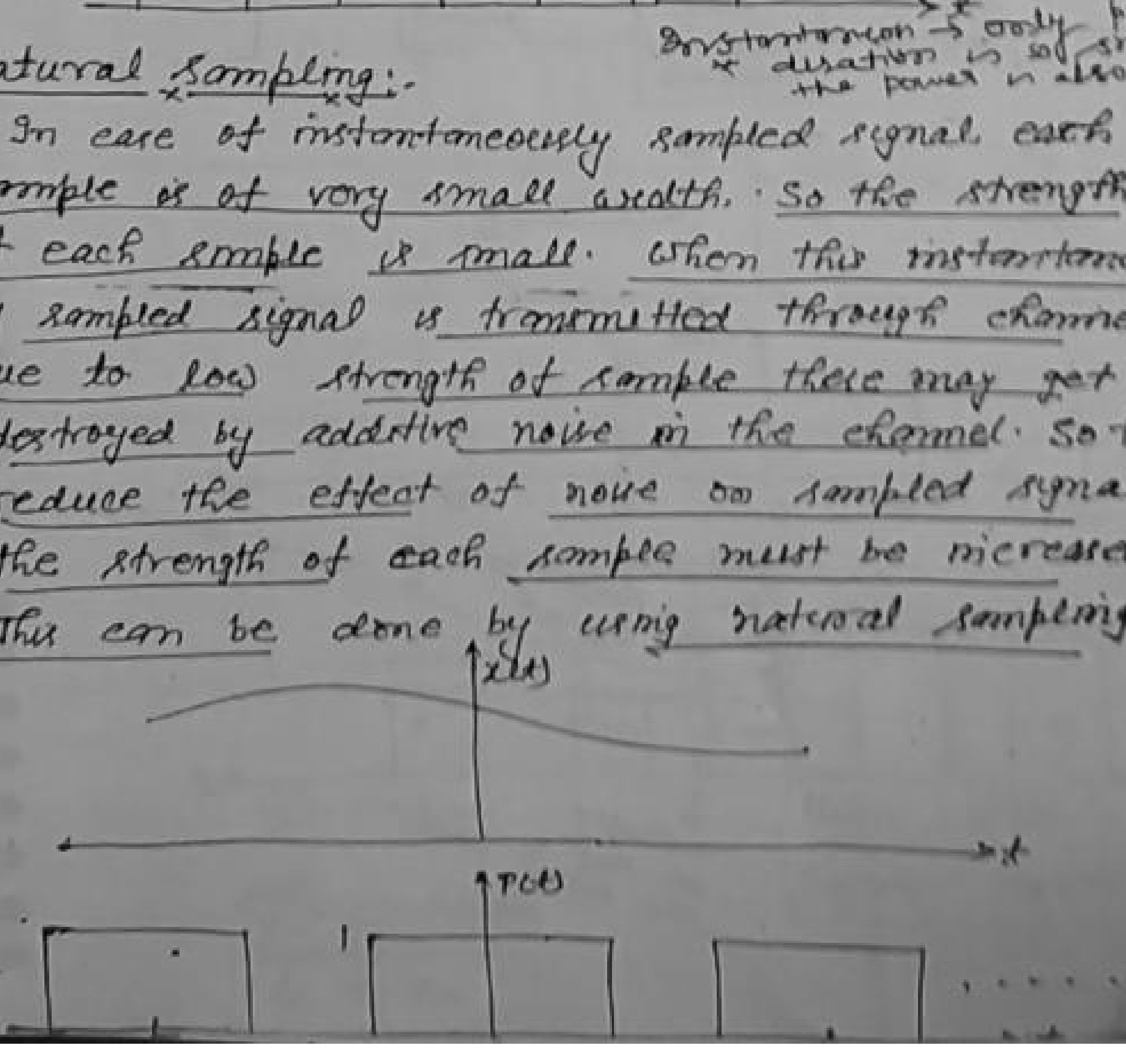
Base Band Pulse Modulation:

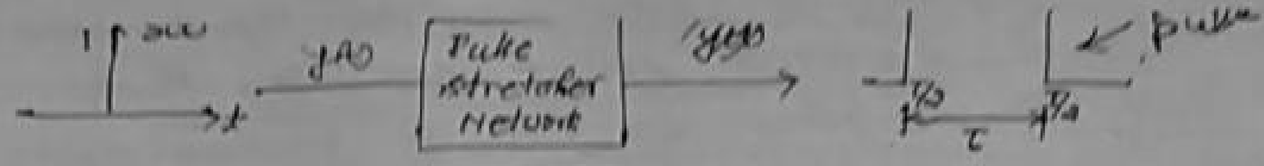


Instantaneous → only pulse duration is so small the power is also small

Natural Sampling:

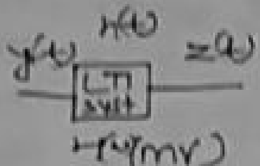
→ In case of instantaneously sampled signal each sample is of very small width. So the strength of each sample is small. When this instantaneously sampled signal is transmitted through channel due to low strength of sample these may get destroyed by additive noise in the channel. So to reduce the effect of noise on sampled signal the strength of each sample must be increased. This can be done by using natural sampling.





$y(t) = x(t) * h(t)$

$Y(\omega) = X(\omega) \cdot H(\omega)$



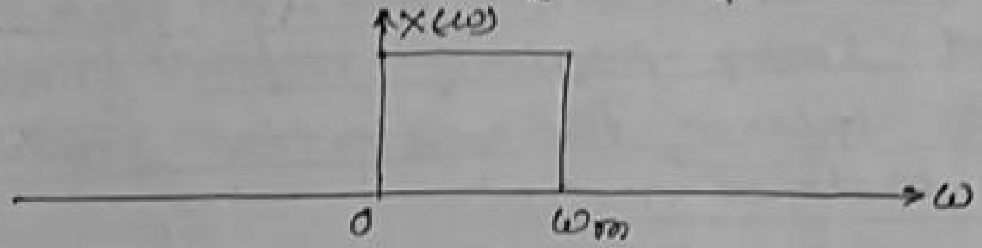
* whenever system is to be tested, always test with a impulse signal, since $\delta(t)$,

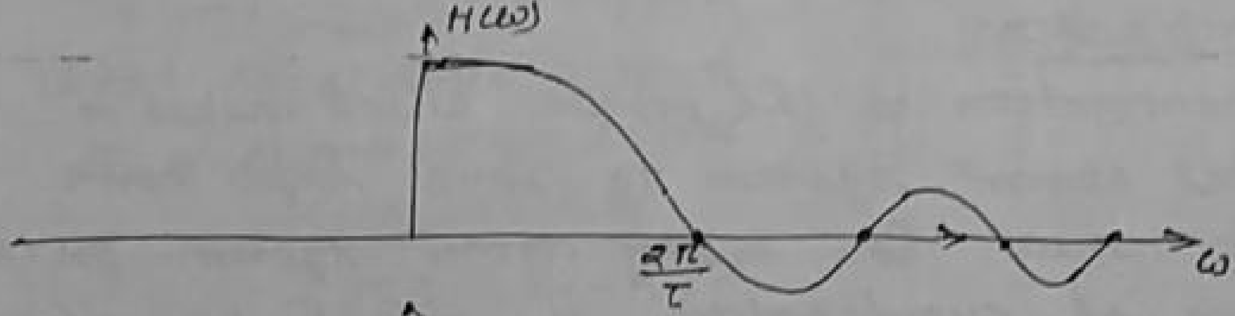
$$H(\omega) = Y(\omega) = \int_{-T/2}^{\tau/2} \cdot e^{-j\omega t} \, dt$$

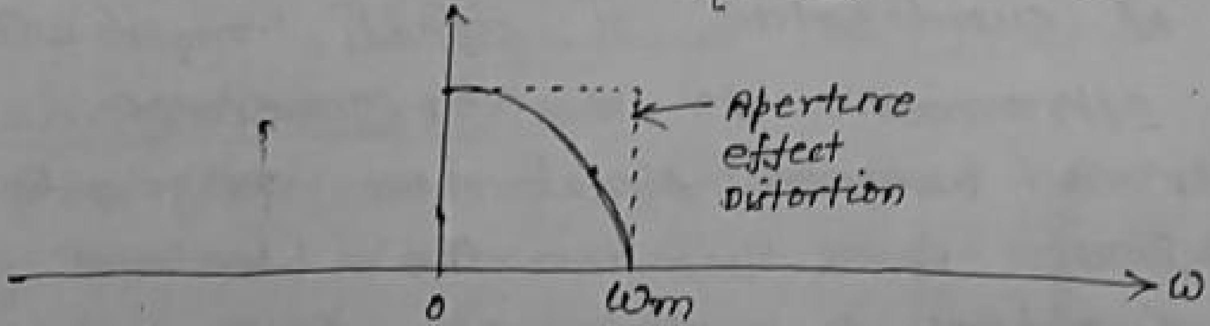
$$= \left| \frac{e^{-j\omega t}}{-j\omega} \right|_{-T/2}^{T/2}$$

$$= 2 \left| \frac{-e^{-j\omega\frac{\tau}{2}} + e^{j\omega\frac{\tau}{2}}}{2j\omega} \right|$$

$$= \frac{2}{\omega} \sin \frac{\omega\tau}{2}$$







$\frac{2\pi}{\tau} \gg \omega_m$

$\frac{2\pi}{\tau} \gg 2\pi f_m$

$\boxed{\frac{1}{\tau} \gg f_m}$

$H_1(\omega) = \frac{1}{H(\omega)}$

Equalizer circuit

→ Aperture effect distortion:-

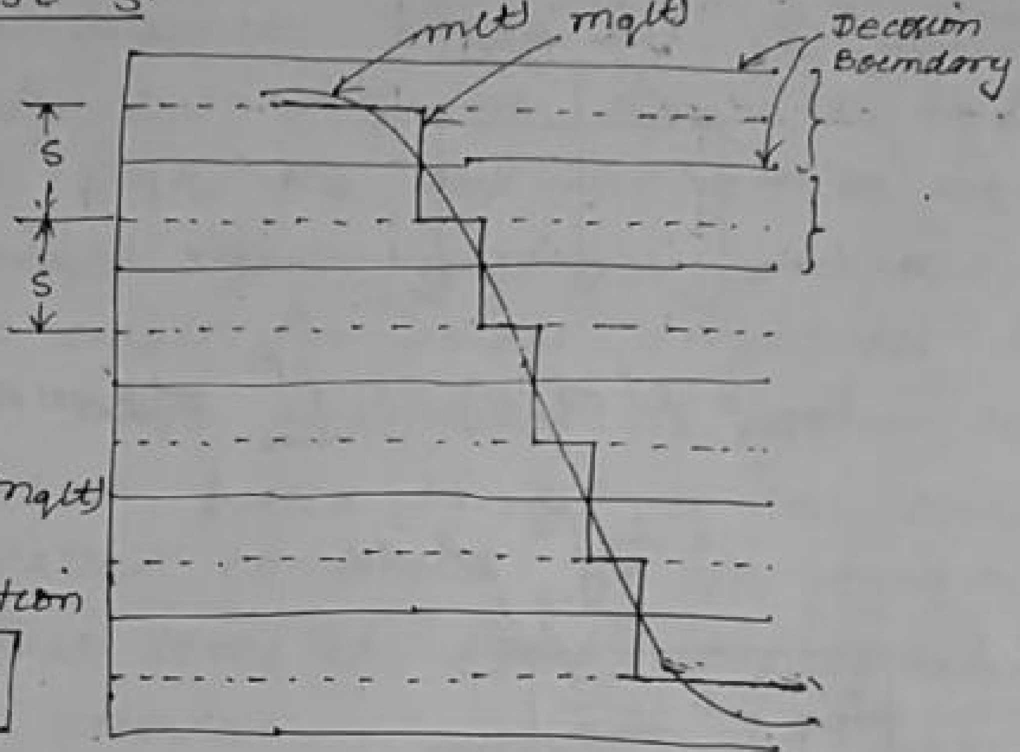
- Due to presence of sinc function, the sampled spectrum is distorted so we can say that in flat top sampling we can not recover our original signal back with a LPF due to aperture effect distortion. To minimize the distortion it should be adjusted such that the frequency corresponding to zero-cross value of magnitude of sinc function must be greater than the band limiting frequency of base band signal. So to minimize distortion $\boxed{\frac{1}{\tau} >> f_m}$. The aperture effect distortion becomes progressively smaller with decreasing τ and the above distortion is 0 for instantaneous sampling i.e $\tau \neq 0$.

At the receiving end, an equalizer ckt is used having frequency response $\boxed{H_r(\omega) = \frac{1}{H(\omega)}}$ to counter balance the distortion produced by pulse stretcher network.

Quantization:

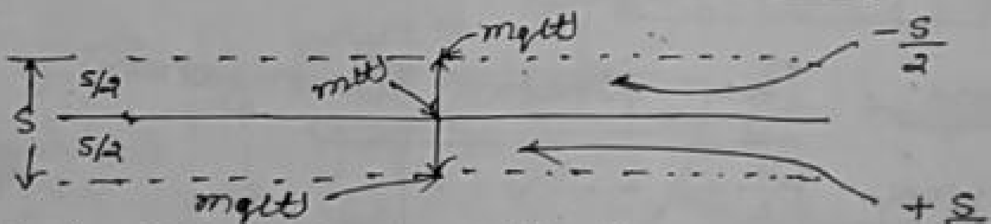
→ Quantization is the process which makes a digital commn system to have super noise performance than analog commn system. In the process of quantization, a signal $m_q(t)$ which is an approximation to $m(t)$ is generated. Consider a baseband signal $m(t)$ varying b/w lower limit V_L & V_H. Now the baseband signal is applied to a quantizer which gives a staircase waveform $m_q(t)$. The $m_q(t)$ is having an advantage that it is free from additive

Note: the continuous signal $m(t)$ is devided into M number of quantization levels and seperation between two successive levels is step size S.



$e_q(t) = m(t) - m_q(t)$

Max^m quantization error $= \boxed{\pm \frac{S}{2}}$



Total signal Range $= V_H - V_L$

No of quantization levels $= M$

Step size $= S$

$$\boxed{M \cdot S = V_H - V_L}$$

Quantization Error:-

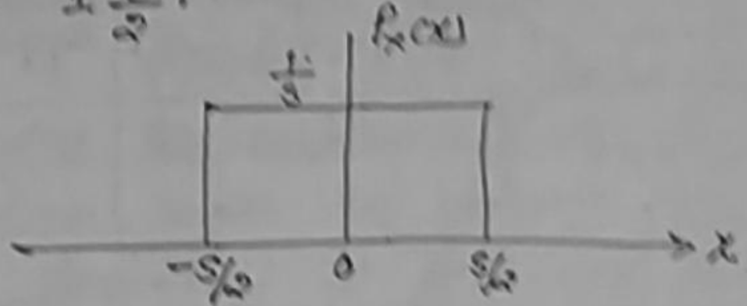
→ From the quantizer response it is clear that at any instant of time $m(t)$ and $m_q(t)$ are not exactly equal. The difference between $m(t)$ & $m_q(t)$ is called as quantization error.

$$e_q(t) = m(t) - m_q(t)$$

→ At the max the quantisation error can be $\pm \frac{S}{2}$.

Quantization Noise Power:-

Quantization noise is uniformaly distributed between $\pm \frac{S}{2}$.



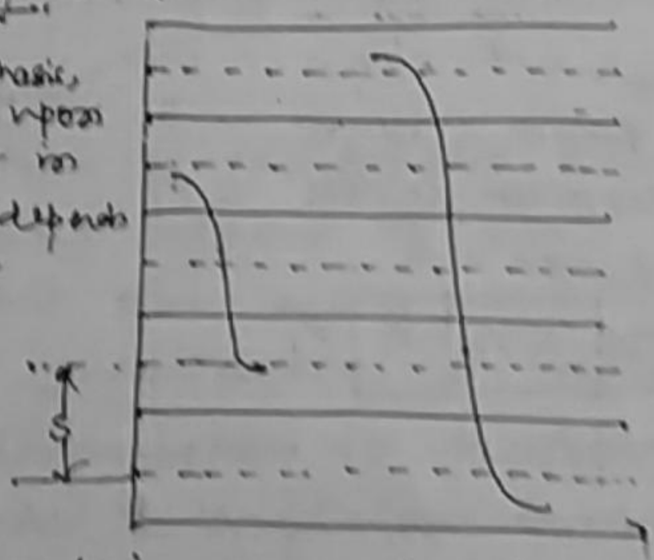
$$P_{nq} = E\left[e_q^2(t)\right] = \int_{-\infty}^{\infty} x^2 f_X(x)\,dx$$

$$= \frac{1}{S}\int_{-\frac{S}{2}}^{\frac{S}{2}} x^2\,dx$$

$$\boxed{P_{nq} = \frac{S^2}{12}}$$

Companding:- (Non uniform Quantization)

* In pre-emphasis, it is depended upon frequency, but in compression, it depends on amplitude content.

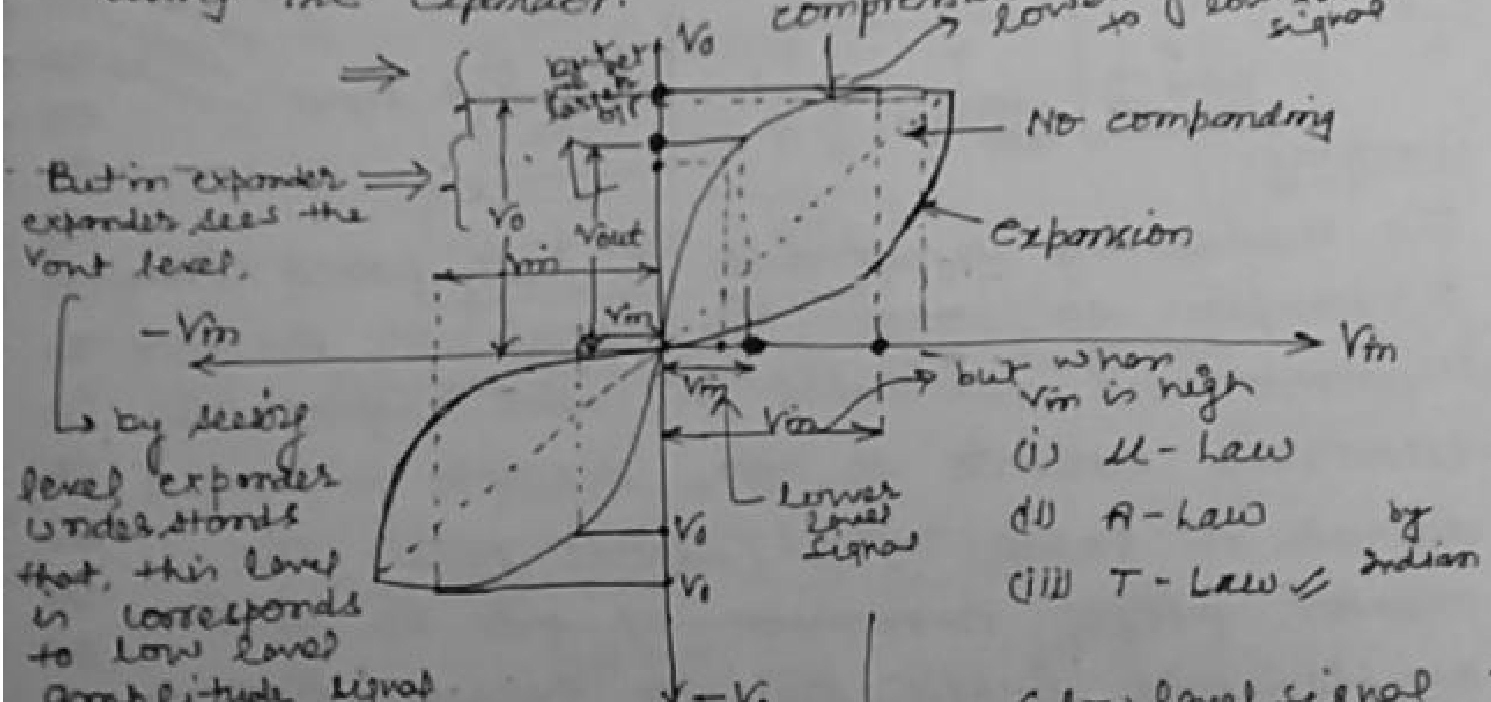


$$SNqR = \frac{\text{signal amplitude}}{\text{Frequency}}$$

depend only upon

→ The quantisation error is dependent on step size. Considering signal to quantization noise ratio (SNqR) when uniform quantiser is applied. lower amplitude signals will have a poorer SNR

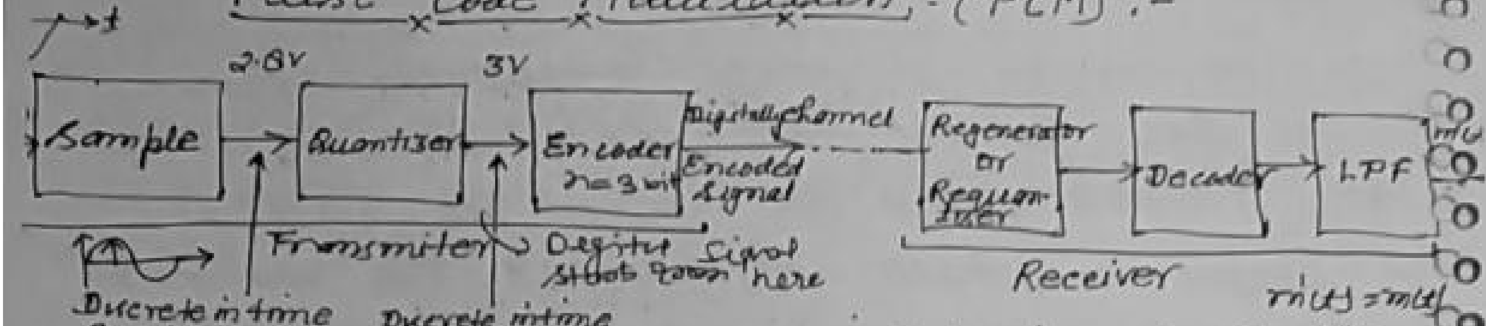
... compression is large amplitude signals. This is due to the fact that denominator of SNqR remains same for lower as well as for larger amplitude signals but the numerator is low for low amplitude signals. So to have a uniform SNqR, the step size is to be adjusted in accordance with signal i.e. step size should be small for small amplitudes and large for large amplitudes. To perform this the signal is passed through a non-linear network called compressor. At the receiving end original level of signals is maintained using the expander.



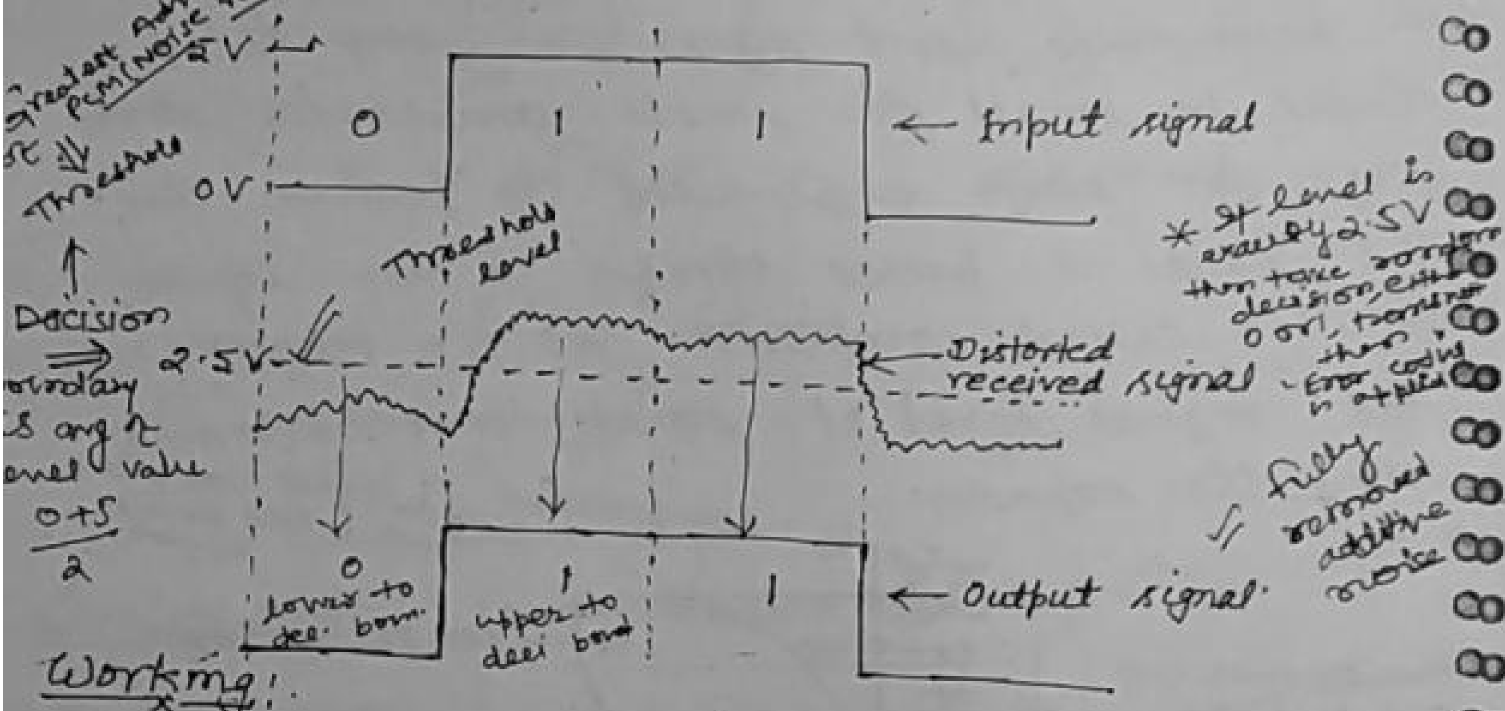
{ low level signal provide high gain
∴ V_{out} is larger
see in fig

* bcoz for any combination of input signal, compressor can not provide same o/p
So, there is no ambiguity for expander to recover the original signal.

Pulse Code Modulation :- (PCM) :-



* Noise drastically removed by FM, & PCM



Working :

- The modulating signal m(t) is first passed through a sampler to convert the signal in discrete in time and continuous in value signal. The quantizer converts it into discrete in time & discrete in value signal (digital signal). The encoder assigns n-number of bits to each sample and finally digitally encoded PCM signal is transmitted over channel.
- At the receiving end the signal is passed through the regenerator or requantizer which removes the additive noise present in the signal

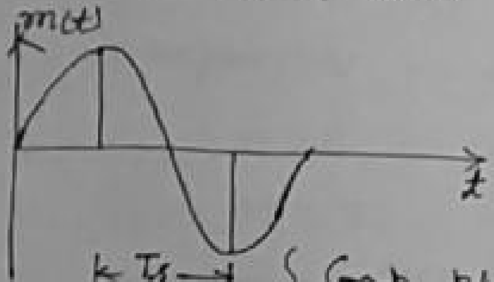
... generates given digital pulses. The decoder converts the digitally encoded signal into corresponding analog value and the low pass filter is used to reconstruct the original signal.

Date 1/01/2010

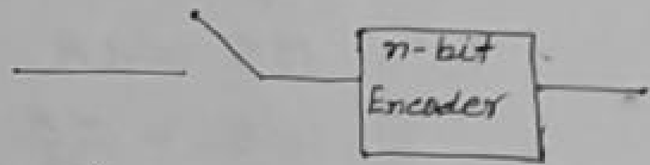
Bandwidth of PCM signal:

$$\boxed{2^n \geq M}$$

M → No of quantization level
n → No of bits



$f_s = \frac{1}{T_s}$



kT_s → {Gap b/w Two Samples}

T_s sec ⟶ n-bits

1 sec ⟶ $\frac{n}{T_s}$ bits/sec

$$\boxed{R_b = n f_s \text{ bits/sec}}$$

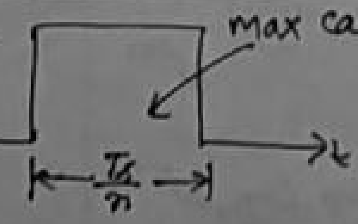
↑

Bit rate / Data Rate / Tx-rate / Pulse rate

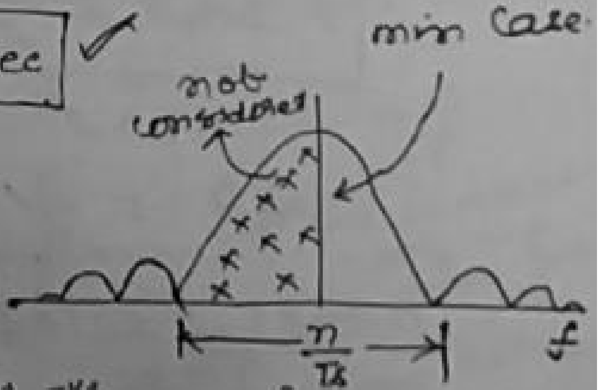
n-bits ⟶ T_s sec

$$\boxed{\text{Max Pulse width} = \frac{T_s}{n} \text{ sec}}$$ ✓

Expansion in time domain is comparison in Freq domain



PSD ⟶



$(BW)_{PCM} \geq \frac{n}{2T_s}$ ✓ {No -ve portion is used to calc BW}

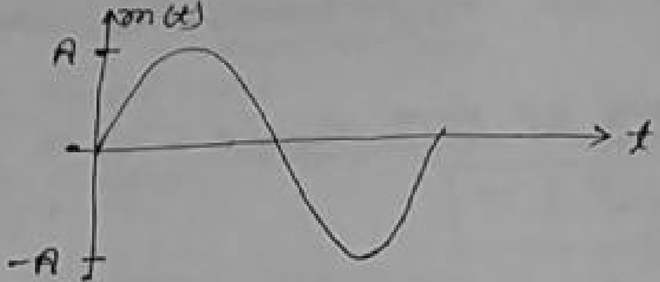
$$\boxed{(BW)_{pcm} \geq \frac{n \cdot f_s}{2}}$$

Min BW ⇒ $(BW)_{min} = \frac{n f_s}{2} = \frac{n \cdot (2 f_m)}{2} = \boxed{n f_m}$ ✓✓

SNqR in PCM :-

(Signal to quantization Noise Ratio)

→ Consider a sinusoidal modulating signal $m(t)$ varying between $\pm A$.



$$P_s = \frac{A^2}{2}$$

$$V_H - V_L = 2A$$

$$\downarrow$$

$$MS = 2A$$

$$\therefore S = \frac{2A}{M}$$

$$P_{nq} = \frac{S^2}{12} = \frac{\left(\frac{2A}{M}\right)^2}{12}$$

$$= \frac{4A^2}{12M^2}$$

$$SNqR = \frac{P_s}{P_{nq}} = \frac{\left(\frac{A^2}{2}\right)}{\left(\frac{4A^2}{12M^2}\right)}$$

$$= \frac{\cancel{A^2}}{2 \times \cancel{4} \cancel{A^2}} \times \cancel{12}^{3} M^2$$

$$\boxed{\therefore SNqR = \frac{3}{2} M^2}$$

$$M = 2^n$$

$$SNqR = \frac{3}{2}(2^n)^2 = \frac{3}{2} \cdot 2^{2n}$$

$$(SNqR)_{dB} = 10 \log_{10} \frac{3}{2} + 10 \log_{10} 2^{2n}$$

$$\boxed{(SNqR)_{dB} = 1.761 + 6.02n \text{ dB}} = \text{Dynamic Range}$$

Dynamic Range

$\boxed{SNqR \approx 6n\,dB}$ ← when n is Large.

(i) $n = 6$: $SNqR = 37.8\,dB$

(ii) $n = 7$: $SNqR = 43.8\,dB$

(iii) $n = 8$: $SNqR = 49.8\,dB$

6dB

6dB

→ In PCM system for each bit increase, the SNqR increases by 6 dB.

→ comparators → $2^n - 1$ Disadvantage

$n = 6$; 63

$n = 7$; 127

$n = 8$; 255

Disadvantage of PCM System:

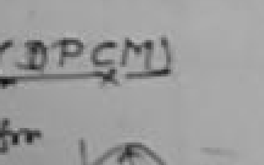
→ As the number of bits increases the system performance improves but at the same time the system complexity also increases.

→ Since bandwidth is also directly proportional to no of bits so it also increases proportionaly.

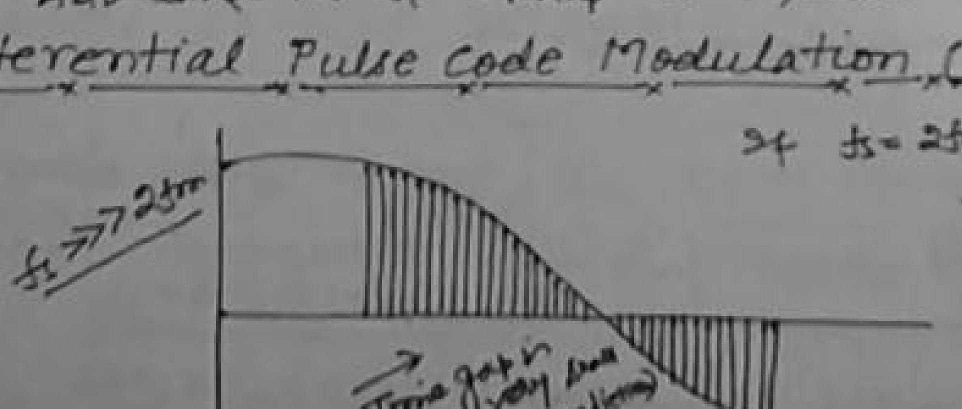
A2D conv ⇐ No of OpAmp $= 2^n - 1$, when n is increased

Differential Pulse Code Modulation (DPCM)

If $f_s = 2f_m$



No prediction due to huge time gap

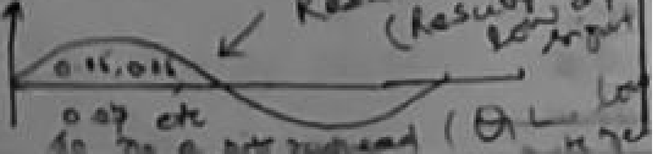


Previous → 2V

Predicted → 1.85 V

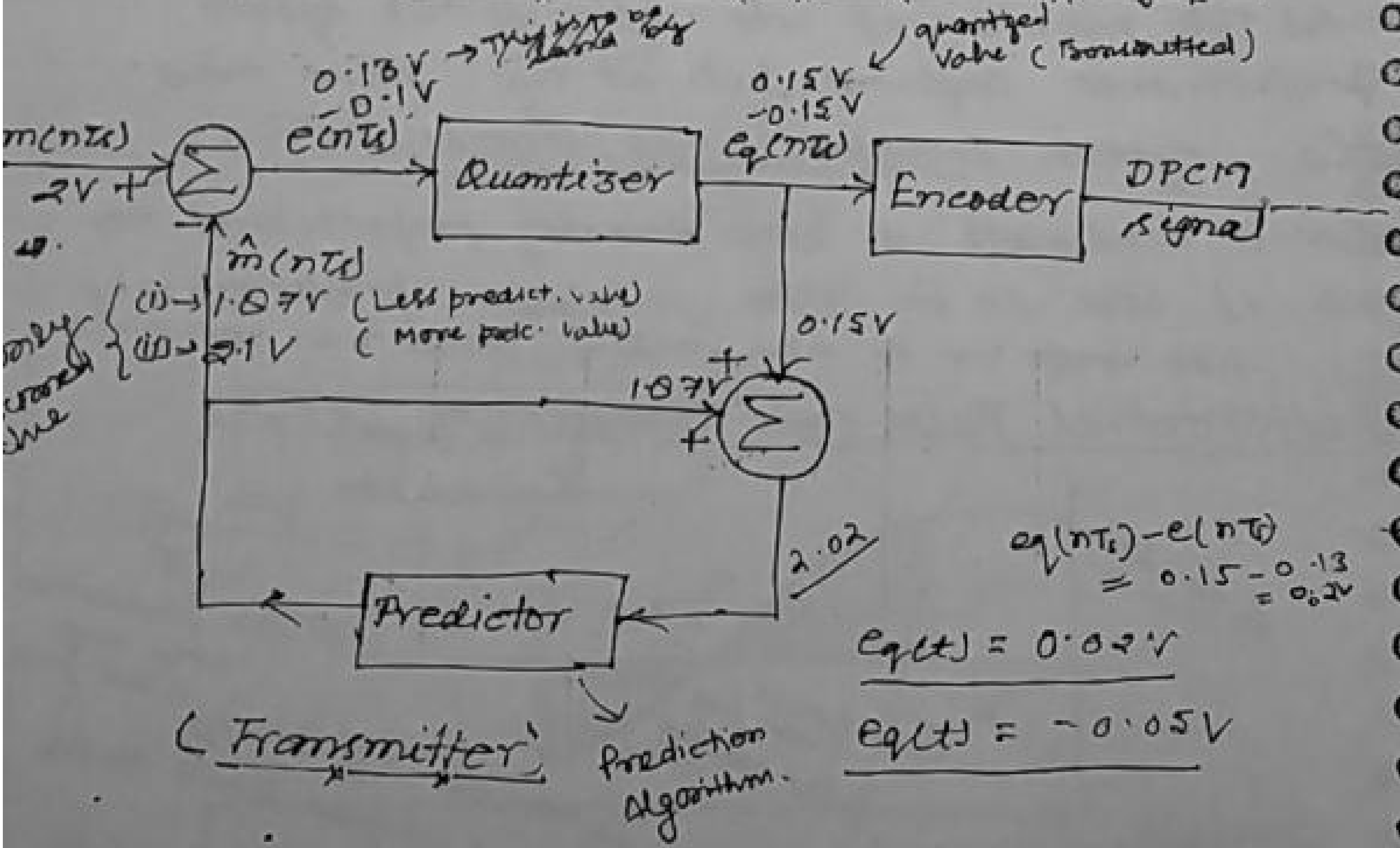
→ 0.05 V ✓ → very low ampl signal Encoded & then transmitted

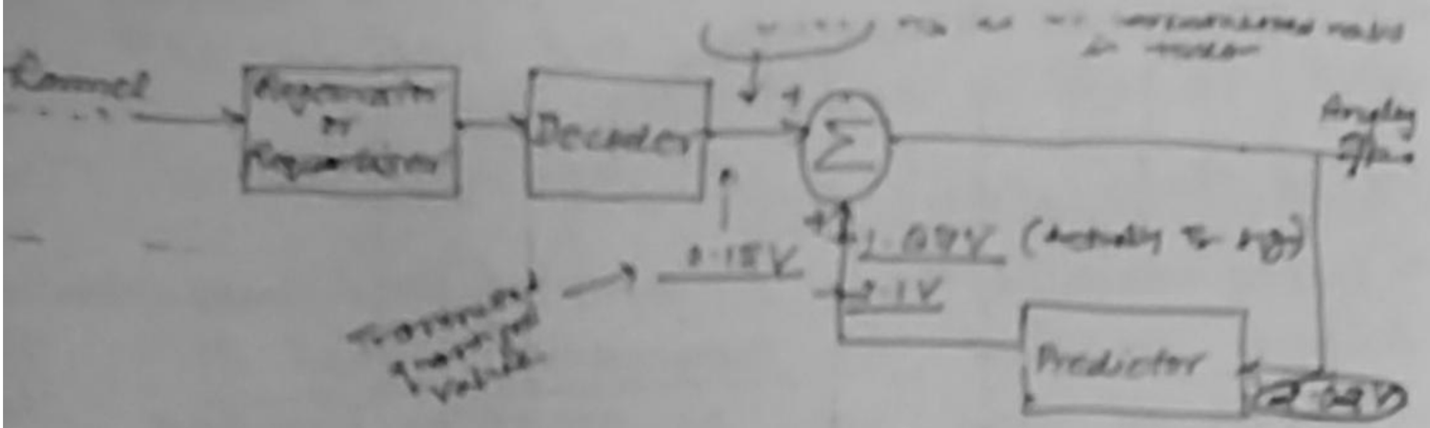
Resultant Signal (Resultant low ampl signal)



• When a signal is sampled at a rate higher than the nyquist rate, the resulting signal is found to exhibit a high correlation between adjacent samples i.e. the signal doesnot change rapidly from one sample to next. So instead of transmitting samples at each sampling instant difference between sample and its predicted value is transmitted. The difference can be added at the receiver to get the original signal mess. Such differential scheme has the advantage when being transmitted by using PCM. that the signal formed by all these above differences will be confined to smaller voltage range than the original signal. Hence no of quantization levels required will be less. and fewer bits will be needed to encode the message.

Generation and Detection of DPCM signal:-





(i) → Analog output $= 1.67 + 0.15V$

$= 2.04V$

(ii) → Analog output $= 2.1 + (-0.15V)$

$= 1.95V$ ← small quantization error

Working:

~~Delta Transmission~~

→ The sampled input and its predicted value is fed to an adder which generates the difference b/w the actual sample value and its predicted value

→ The difference value is quantized, encoded and transmitted as DPCM signal.

→ The quantized value is also fed to an adder to which another input is predicted value. The adder generates the sum of quantized value and predicted value and is given as input to predictor so that it can predict next sample value

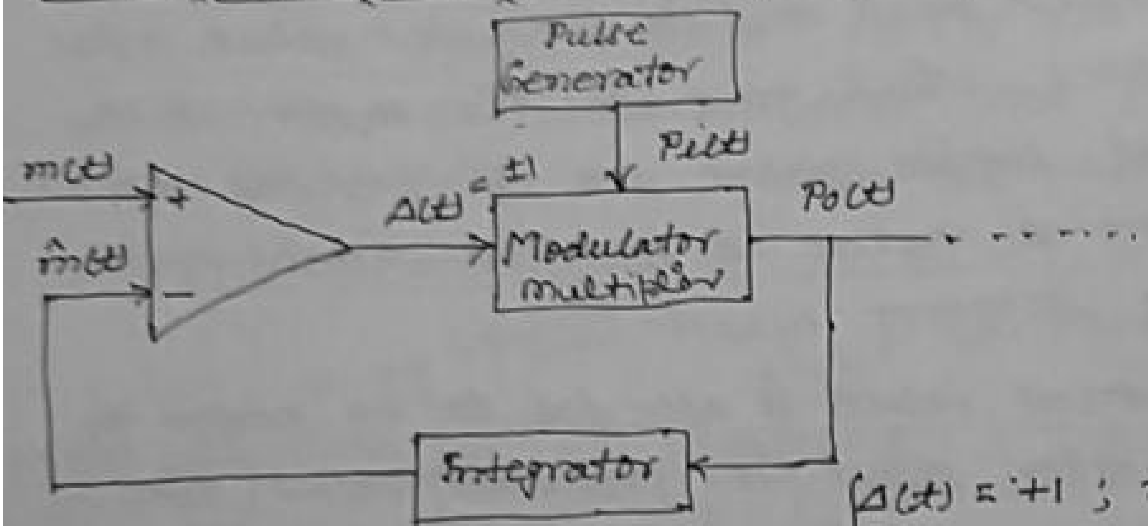
→ At the receiving end the DPCM signal is passed through regenerator which removes the additive noise from the signal and generates fresh pulses. The decoder ckt converts digitally encoded signal into corresponding analog value

→ This analog value is added with the o/p of receiving end predictor to generate the analog output the receiving end predictor predicts exactly the same as transmitting end predictor, it input to both is same.

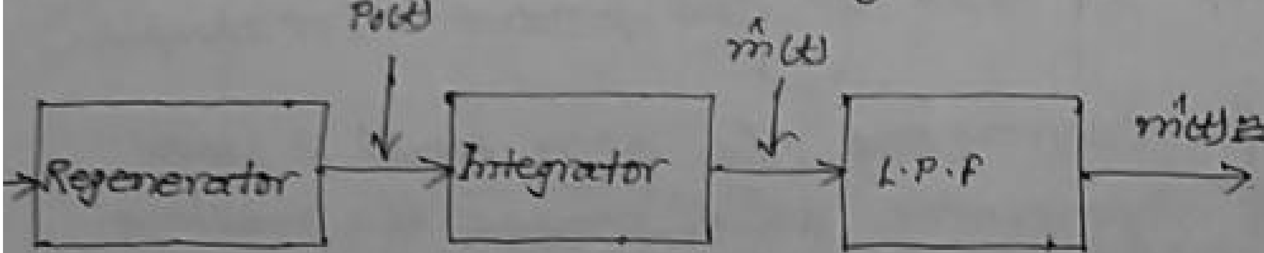
Delta Modulation :- (DM):

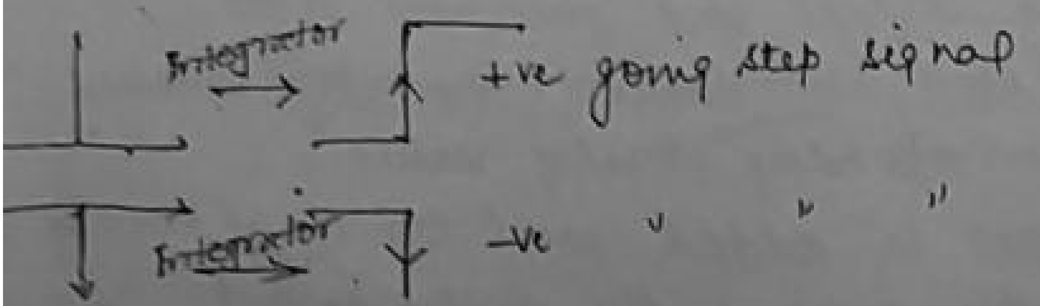
→ Delta modulation is a oversampled modulation scheme in which the transmitted signal is encoded in one bit. In DM the stair case approximation of original modulating signal m(t) is generated which is passed through a LPF to reconstruct the original signal back.

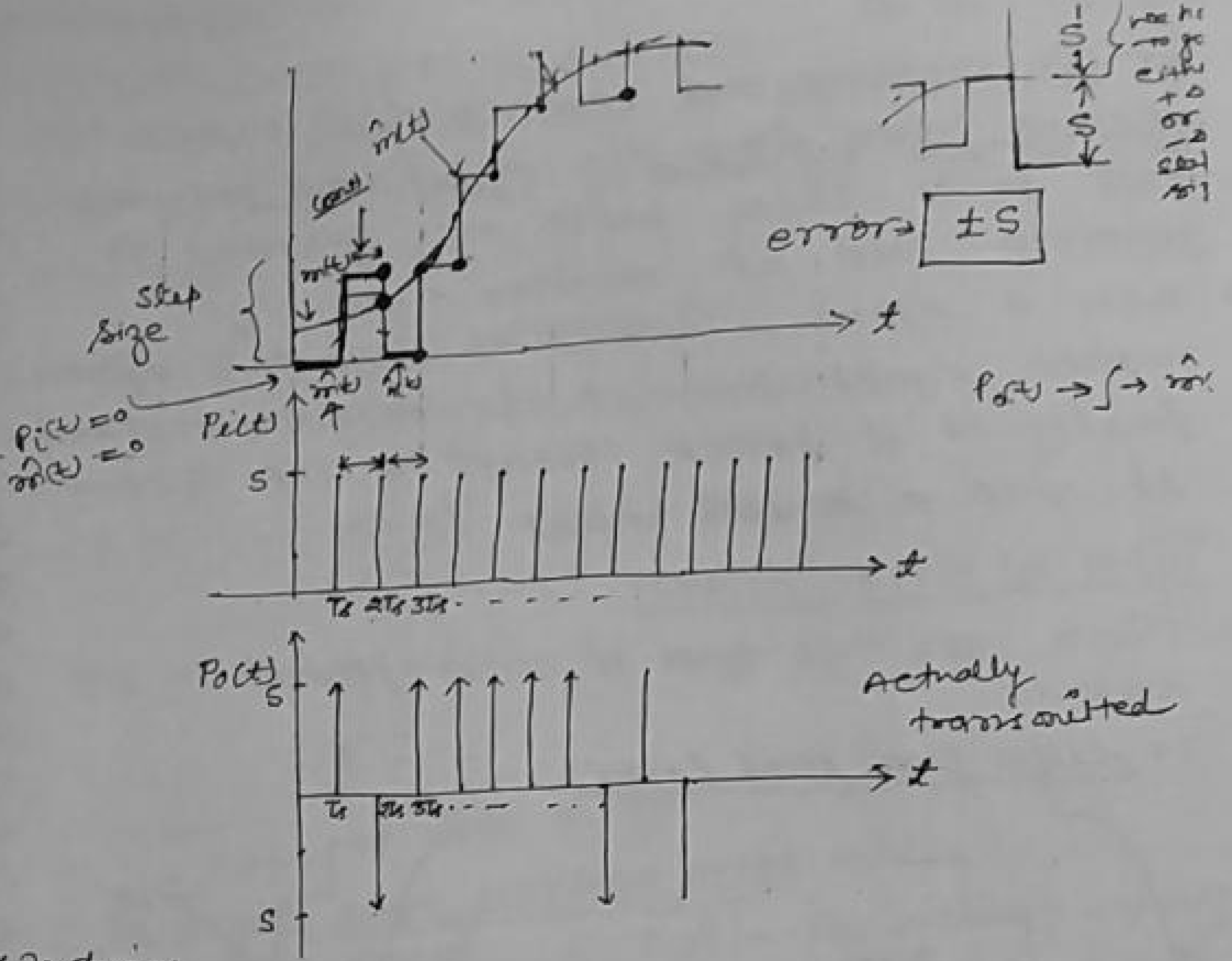
Generation and Detection of DM Signal:



$$\begin{cases} \Delta(t) = +1 \; ; & m(t) > \hat{m}(t) \\ \quad\;\; = -1 \; ; & m(t) < \hat{m}(t) \\ \quad\;\; = \text{Random} & m(t) = \hat{m}(t) \end{cases}$$







Working:

→ The modulating signal $m(t)$ and its approximation value $\hat{m}(t)$ are compared in an open comperator which generates.

$$\Delta(t) = +1 : m(t) > \hat{m}(t)$$
$$= -1 : m(t) < \hat{m}(t)$$
$$= \text{Random} : m(t) = \hat{m}(t)$$

{Never become zero, three level signal but $n=1$, $2^1 = 2$ levels

→ The modulator ckt multiplies the output of comperator and with output of pulse generator $P_i(t)$ to generate output impulses $P_o(t)$. The polarity of $P_o(t)$ is dependent on polarity of $\Delta(t)$.

→ The $P_o(t)$ impulses are encoded and transmitted as DM signal. Simultaneously these impulses

are integrated to generate the approxim...
signal $\hat{m}(t)$

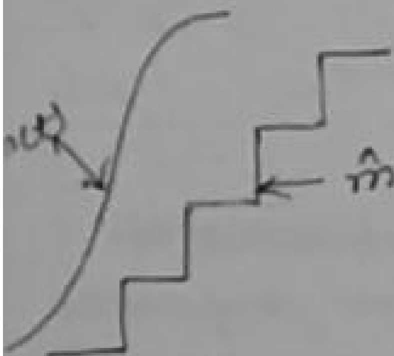
- At the receiving end regenerator removes the additive noise from the signal and generates fresh pulses. These pulses are decoded to generate $P_o(t)$ at receiving end.

→ $P_o(t)$ is again integrated to regenerate approximation signal $\hat{m}(t)$ at the receiving end and finally it is passed through a LPF to convert it into a smooth signal.

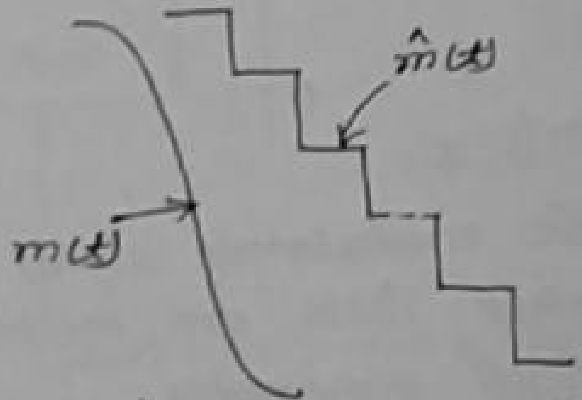
<u>Types of Error in DM:</u>

→ There are two types of errors present in DM system.

1→ <u>Slope Over load Error:</u>



(Positive slope overload Error)



(Negative slope overload Error)

→ "The error produced due to steep slope of input signal as compaired to slope of approximation signal is known as slope-overload error. There are two types of slope overload error."

(i) <u>Positive Slope-Over load Error:-</u>

The error produced due to positive steep slope of input signal is known as positive slope over-load error.

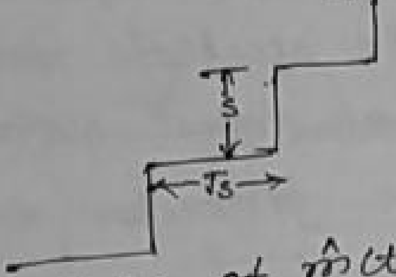
(ii) Negative slope overload error:-

The error produced due to negative steep slope of input signal is known as negative slope over load error.

$$* \quad m(t) = A \sin\omega_m t$$

$$\frac{dm(t)}{dt} = A \cdot \omega_m \cos\omega_m t$$

$$\left.\frac{dm(t)}{dt}\right|_{max} = A \cdot \omega_m$$



$$\text{Slope of } \hat{m}(t) = \frac{S}{T_s} = S \cdot f_s$$

To avoid slope overload error:-

Imp $$\boxed{\left.\frac{dm(t)}{dt}\right|_{max} \leq S \cdot f_s}$$ ← For arbitary signal

$$A \cdot \omega_m \leq S \cdot f_s$$

$$\boxed{A \leq \frac{S \cdot f_s}{2\pi f_m}} \quad \& \quad \boxed{S \geq \frac{A \cdot \omega_m}{f_s}}$$

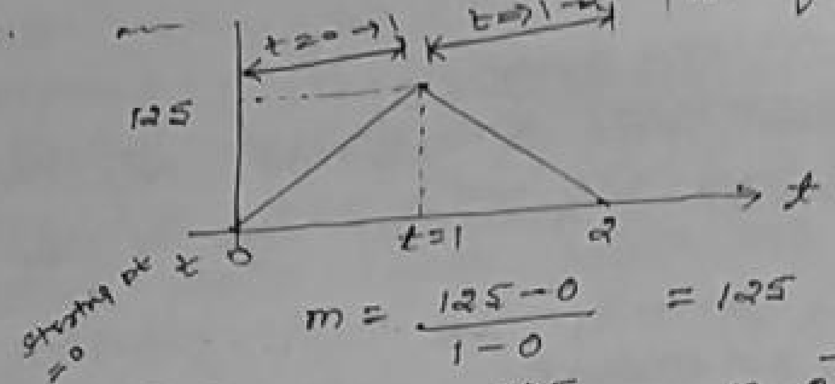
For sinusoidal signal

Q→ Consider a delta modulator with step size S and sampling frequency f_s = 32 K samples/sec. The input to the delta modulator is

$$x(t) = 125t\left[\underbrace{u(t) - u(t-1)}_{0 \longrightarrow 1}\right] + (250 - 125t)\left[\underbrace{u(t-1) - u(t-2)}_{1 \longrightarrow 2}\right]$$

calculate minimum step size to avoid slope over load error.

(in eqn) put



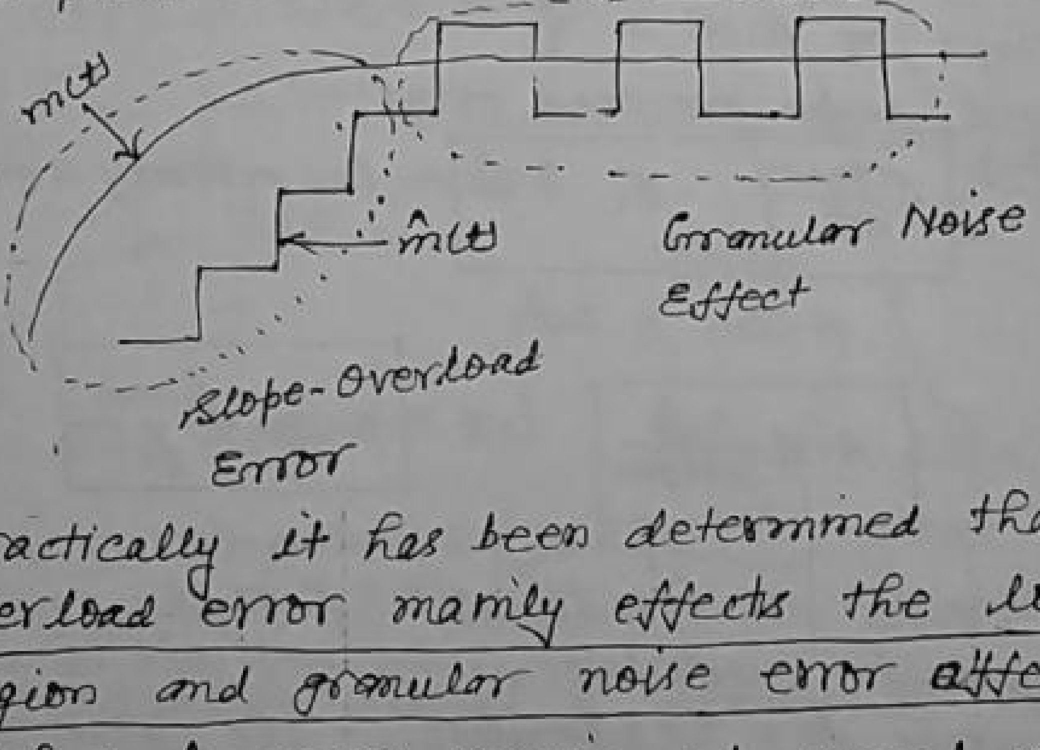
If slope is 125 & 250 then u have 2 calculate for max slope i.e 250

$$m = \frac{125-0}{1-0} = 125$$

$$S_{min} = \frac{125}{32 \times 10^3} = 2^{-8}$$

Granular Noise Effect :-

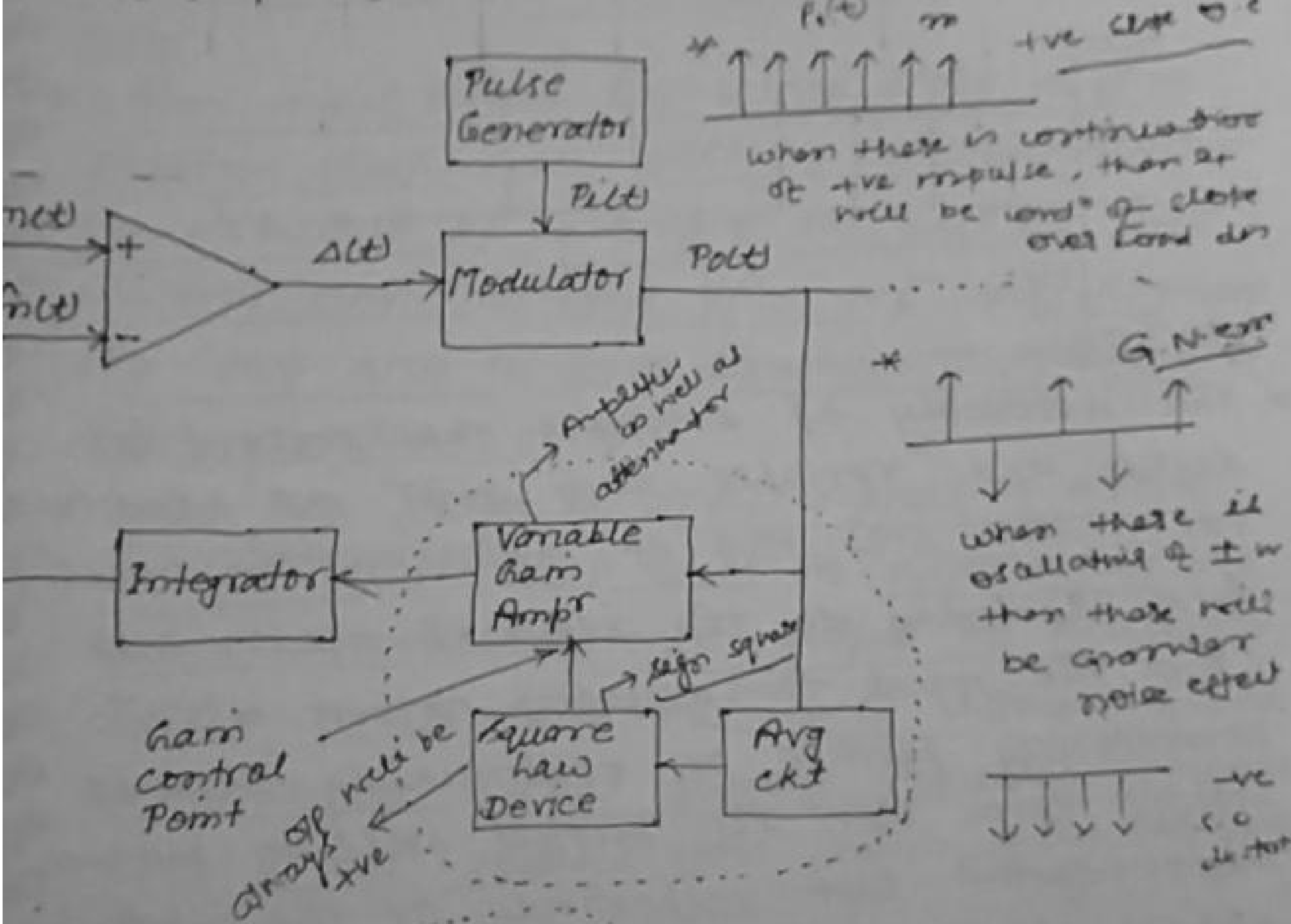
→ When step size S is too large relative to the local slope-characteristics of input waveform $m(t)$ thereby causing $\hat{m}(t)$ to oscillate around a relatively flat segment of input waveform. This phenomenon is known as granular noise

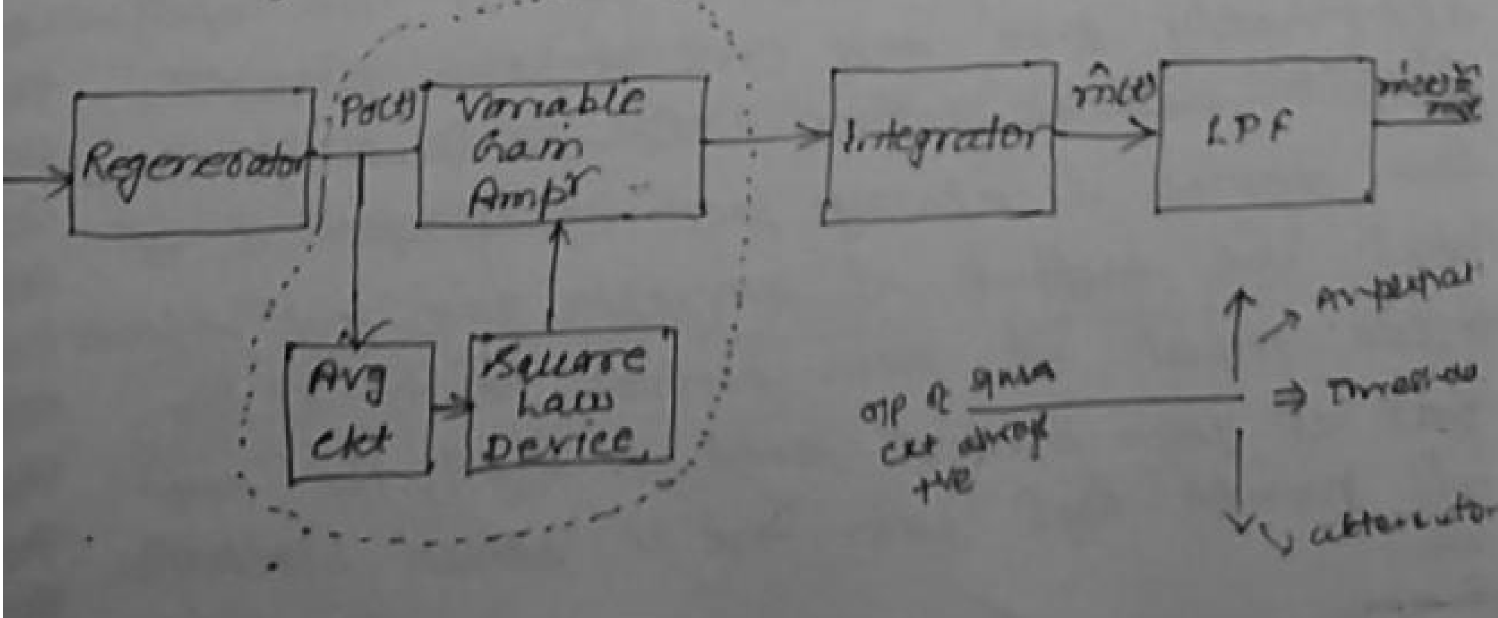


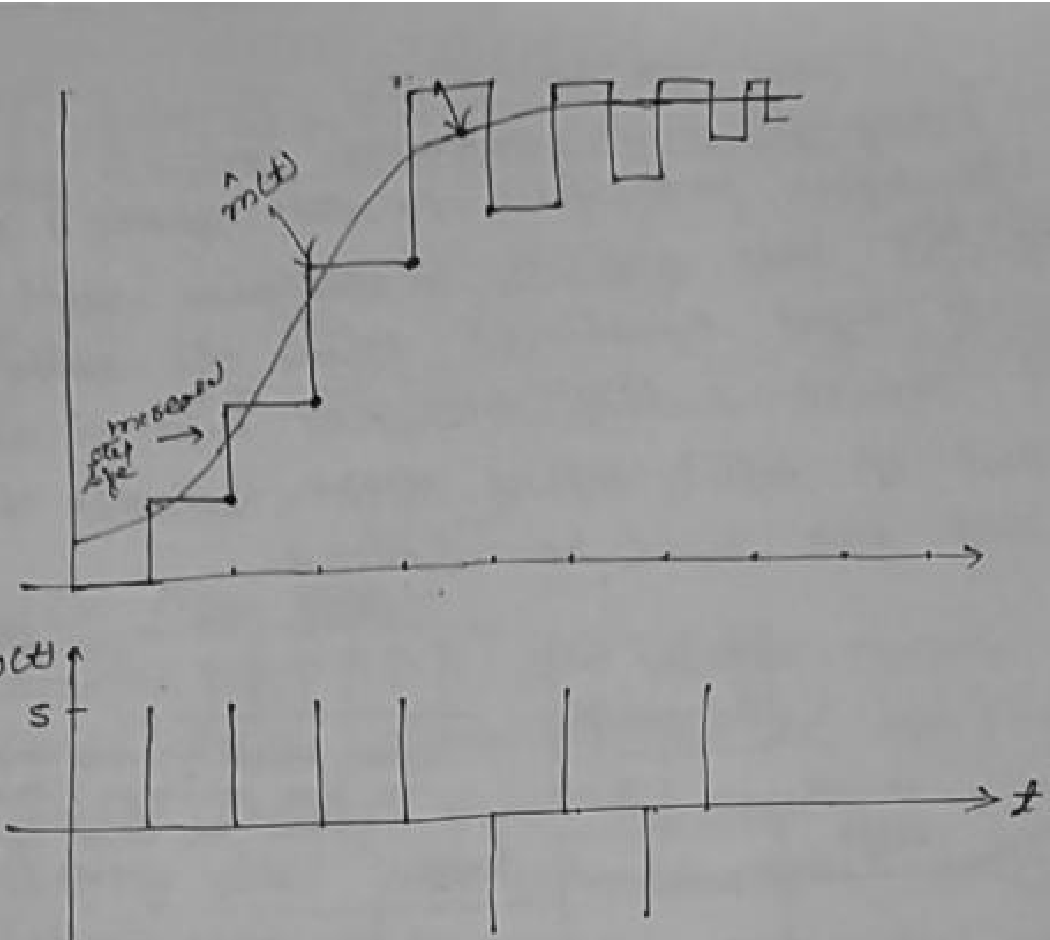
→ Practically it has been determined that slope overload error mainly effects the lower freq- region and granular noise error affects the higher frequency region of signal waveform. Since most of the information lies only in lower frequency region so slope-over load error is more objectionable.

02/01/2014 [illegible]

→ The basic principle of ADM system is to varry the step size in accordance with slope of input signal. i.e when the slope of input signal is steep, step size must be increase and if input signal slope is small then the step size must be reduced.







Working:

→ The assembly of averaging ckt, square law device and variable gain ampr are used to varry the step-size of approximation signal $\hat{m}(t)$ according to the slope of input signal. When the input to averaging ckt is either continuously positive or continuously negative impulse train then its output will be positive and or negative high respectively. If the input is alternatly ⊕ve and ⊖ve impulse then its output will be zero(0). The output of square law device is always positive irrespective of input. The output of variable gain ampr depends on the input at gain control point. If the input at gain control point is high, the ampr will provide high gain and it almost block the

... gain control point is
zero(s).

2nd ⇒ If the input signal has steep slope in +ve direction then Po(t) will be continuously +ve hence the output of averaging ckt will be +ve high and ampr will provide high gain so as the to minimize +ve slope overload error.

→ When input signal has steep slope in -ve direction then Po(t) will be continuously -ve
- hence output of averaging ckt will be high in -ve direction and the ampr will increase the step size in -ve direction to avoid -ve slope overload error

→ When the slope of input signal is almost constant then Po(t) will be alternately +ve and -ve impulses. Hence output of averaging ckt is zero(s) and ampr will almost block the incoming signal thereby reducing stepsiz

→ ADM is not suitable for continuously varying signal.

Disadvantage of ADM:-

→ Adaptive delta modulation is not suitable for periodic signals.

Comparision between PCM, DM, ADM and DPCM:-

Comparison between PCM, DM, ADM & DPCM:

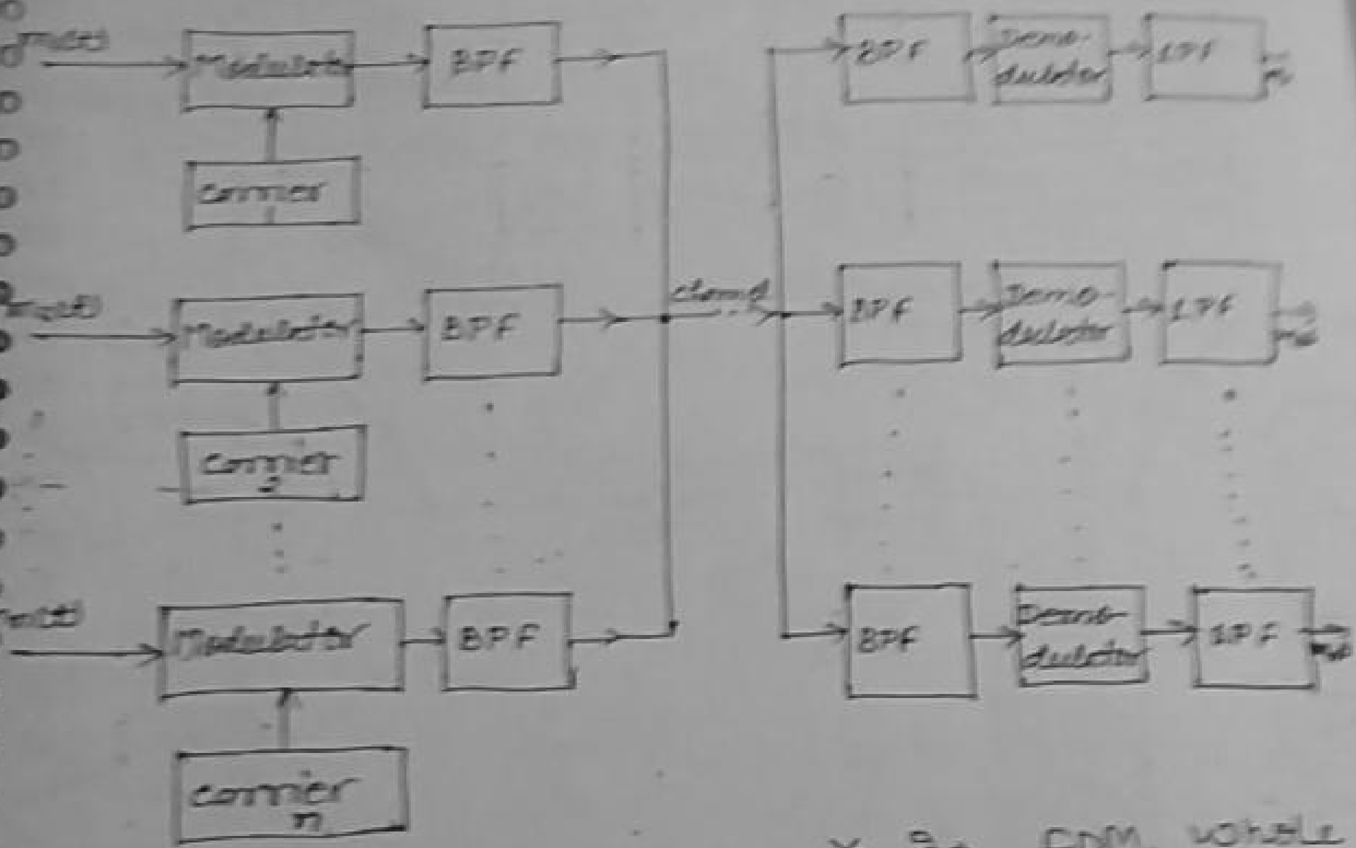
Sl	Parameters	PCM	DM	ADM	DPCM
•	No of bits per sample	8-bit, 12-bit & 16 bits/sample	1-bit	1-bit	greater than 1 bit less than that require in PCM
•	Step-size	fix	fix	Variable	fix
•	Sampling-Rate	slightly greater than adm	Oversampled	Oversampled	Oversampled
•	Complexity	Highly complex	Simplest	Simplest	More complex than DM & ADM but less complex than PCM
•	Feedback	No feedback	Yes	Yes	Yes
	Types of error	mainly Quantization error	Fair by slope overload error & granular noise effect	Both slope overload error as well as granular noise but very less than DM	mainly Quantization error.

Multiplexing of signals:

→ Transmission of more than one signal simultaneously over same channel is known as multiplexing. Multiplexing requires that signals must be kept apart so that they donot interfere each other and hence can be seperated at the receiving end. This is accomplished by seperating the signals either in frequency or in time. There are two types of multiplexing schemes.

Frequency division multiplexing (FDM)

→ The technique of separating signals in frequency is known as FDM



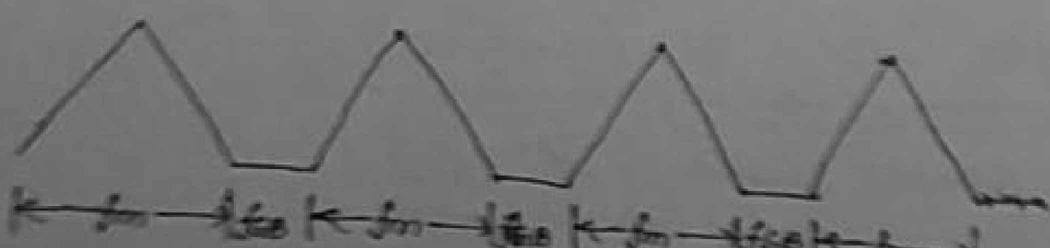
* In FDM, whole signal is transmitted but in TDM only samples are transmitted

BW of FDM System:-

①→

$$\boxed{BW = n f_m}$$

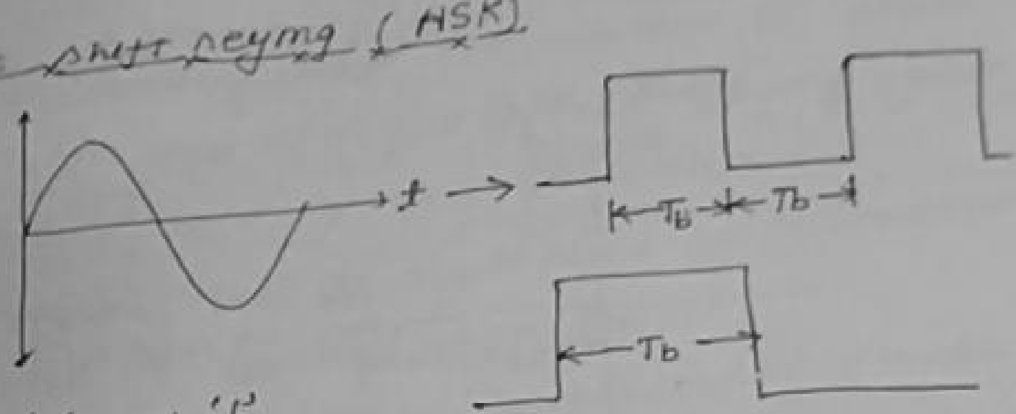
②→



$$\boxed{BW = \underline{n f_m} + \underline{(n-1) f_{GB}}}$$

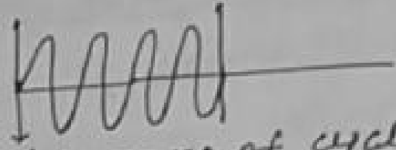
[illegible] for different transmitters

Amplitude shift keying (ASK)



$S_1(t) = A\cos 2\pi f_c t \longrightarrow$ '1'

$S_2(t) = 0 \longrightarrow$ '0'



integer no of cycles

$\underbrace{\frac{A^2}{2}}_{\text{Power}} \cdot \underset{\text{Time}}{T_b} = E_b \rightarrow$ Energy

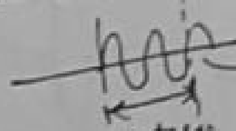
$A = \sqrt{\frac{2E_b}{T_b}}$

$\boxed{f_c = n f_b}$ $\quad n \rightarrow$ Any integer no. of bit freq f_b (must b)

$S_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_c t \longrightarrow$ '1'

$S_2(t) = 0 \longrightarrow$ '0'

$= \int_{T_b} \cos n\omega t = 0$ (must be)



but is not int so it will not be 0

$$S_{11} = \left[\int_0^{T_b} S_1^2(t)\,dt\right]^{\frac{1}{2}}$$

$$S_{11} = \left[\frac{2E_b}{T_b}\int_0^{T_b}\left(\frac{1+\cancel{\cos 4\pi f_c t}^{\,0}}{2}\right)dt\right]^{\frac{1}{2}}$$

$$S_{11} = \left[\frac{2E_b}{T_b}\int_0^{T_b}\frac{1}{2}\,dt\right]^{\frac{1}{2}}$$

$$S_{11} = \sqrt{E_b}$$

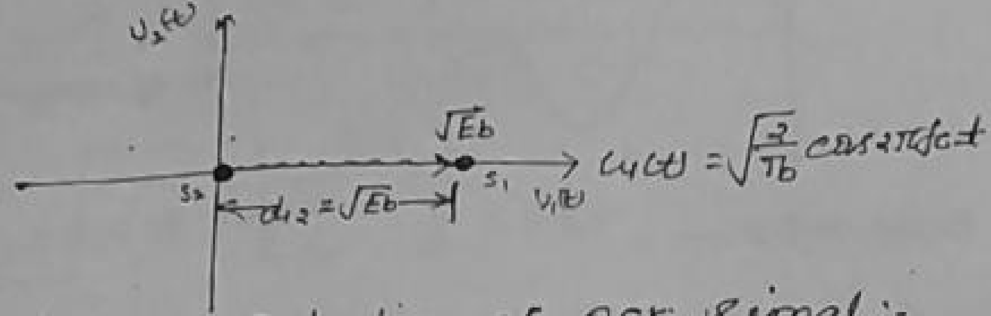
$$\phi_1(t) = \frac{S_1(t)}{S_{11}} = \frac{\sqrt{\frac{2E_b}{T_b}}}{\sqrt{E_b}}\cos 2\pi f_c t$$

$$\phi_1(t) = \sqrt{\frac{2}{T_b}}\cos 2\pi f_c t \Rightarrow \text{orthonormal function}$$

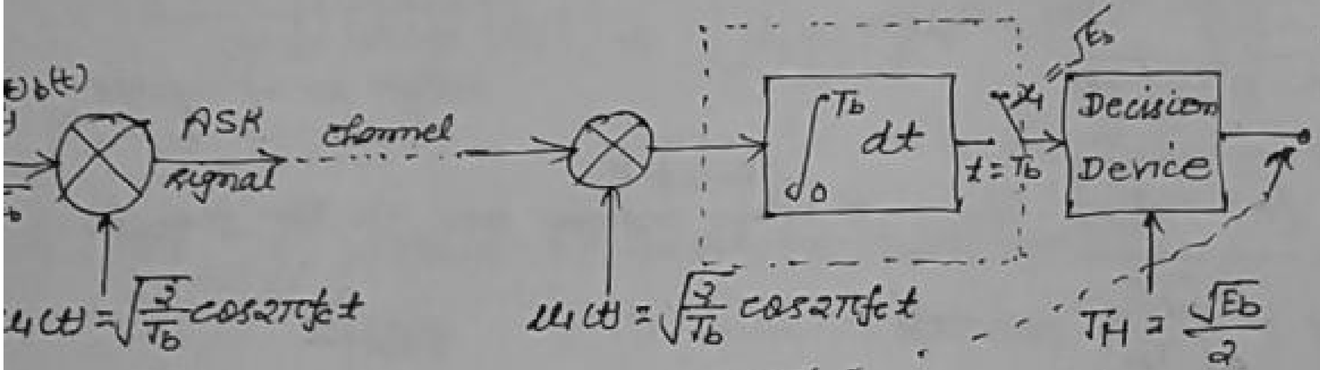
Constellation diagram:

$$s_1(t) = \underbrace{\sqrt{E_b}}_{s_{11}} \times \underbrace{\sqrt{\frac{2}{T_b}} \cos 2\pi f_c t}_{u_1(t)}$$

$$s_2(t) = 0$$



Generation and Detection of ASK Signal :-



$$\hookrightarrow \frac{0 + \sqrt{E_b}}{2} = \frac{\sqrt{E_b}}{2}$$

$$\begin{cases} \text{choose } 1, & x_1 > \frac{\sqrt{E_b}}{2} \\ 0; & x_1 < \frac{\sqrt{E_b}}{2} \\ \text{Random;} & x_1 = \frac{\sqrt{E_b}}{2} \end{cases}$$

For '1' :-

$$\rightarrow \sqrt{E_b} \times \sqrt{\frac{2}{T_b}} \cos 2\pi f_c t \times \sqrt{\frac{2}{T_b}} \cos 2\pi f_c t$$

$$= \int_0^{T_b} \sqrt{E_b} \times \frac{2}{T_b} \cos^2 2\pi f_c t \, dt$$

$$= \sqrt{E_b} \times \frac{2}{T_b} \times \int_0^{T_b} \left(\frac{1 + \cancel{\cos 4\pi f_c t}^{\,0}}{2} \right) dt$$

$\boxed{x_1 = \sqrt{E_b}} \longrightarrow$ Length of the signal

→ At the transmitting end bit stream $b(t)$ which is a unipolar NRZ waveform having value $\sqrt{E_b}$ for bit '1' and '0' for bit zero(0).

→ The bit stream $b(t)$ is multiplied by orthonormal function $u(t)$ to generate ASK signal.

→ The received ASK signal is multiplied by locally generated orthonormal function to facilitate coherent detection, the product is passed through integrator and dumped switch receiver which maximizes SNR. At $t = T_b$ the sample of signal + noise is taken and is fed to the decision device.

→ The decision device decides b/w '0 & 1' according to its threshold value.

Probability of Error in ASK:

$$\boxed{P_e = \frac{1}{2} \operatorname{erfc}\left[\sqrt{\frac{d_{12}^2}{4\eta}}\right]}$$

put $d_{12} = \sqrt{E_b}$

$$\boxed{P_e = \frac{1}{2} \operatorname{erfc}\left[\sqrt{\frac{E_b}{4\eta}}\right]}$$

Binary Phase Shift Keying:- (BPSK):

$$S_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_c t \longrightarrow \text{'1'}$$

$$S_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_c t + \pi) \longrightarrow \text{'0'}$$

$$S_2(t) = -\sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_c t \longrightarrow \text{'0'}$$

$$\boxed{S_2(t) = -S_1(t)}$$

Linear dependent signal

$$s_{11} = \left[\int_0^{T_b} s_1^2(t)\,dt\right]^{\frac{1}{2}}$$

$$= \left[\frac{2E_b}{T_b}\int_0^{T_b}\left(\frac{1+\cos 4\pi f_c t}{2}\right)dt\right]^{\frac{1}{2}}$$

$$s_{11} = \sqrt{E_b} \leftarrow$$

$$\phi_1(t) = \frac{s_1(t)}{s_{11}} = \sqrt{\frac{2}{T_b}}\cos 2\pi f_c t \leftarrow$$

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

$$= \int_0^{T_b} -\sqrt{\frac{2E_b}{T_b}} \times \sqrt{\frac{2}{T_b}}\,dt$$

$$= -\sqrt{E_b} \leftarrow$$

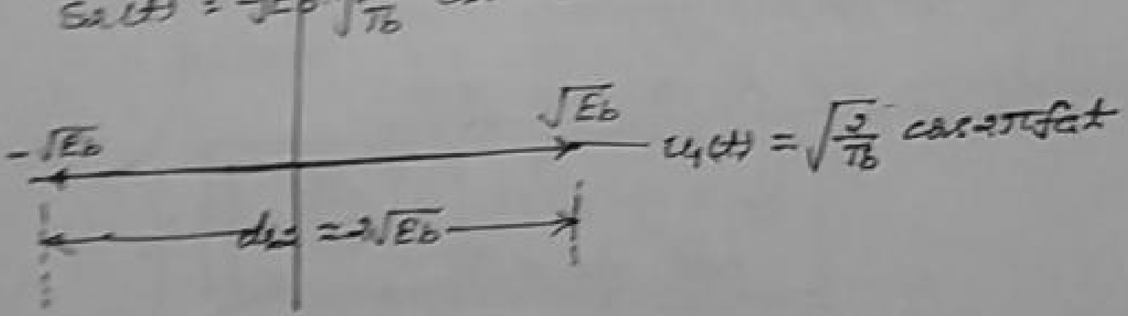
$$\boxed{s_{22} = 0 \leftarrow}$$

{linearly dependent no need of $\phi_2(t)$}

Constellation diagram:

$$s_1(t) = \sqrt{E_b}\cdot\sqrt{\frac{2}{T_b}}\cos 2\pi f_c t$$

$$s_2(t) = -\sqrt{E_b}\cdot\sqrt{\frac{2}{T_b}}\cos 2\pi f_c t$$

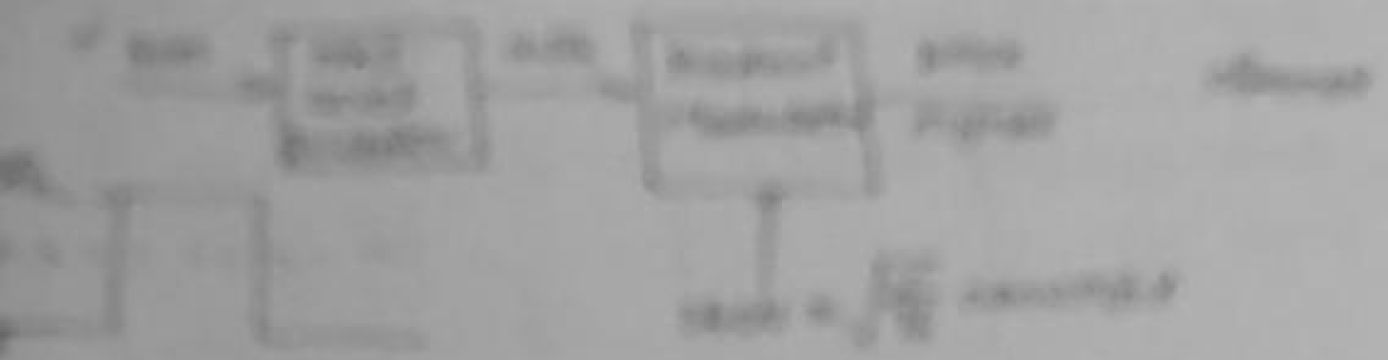


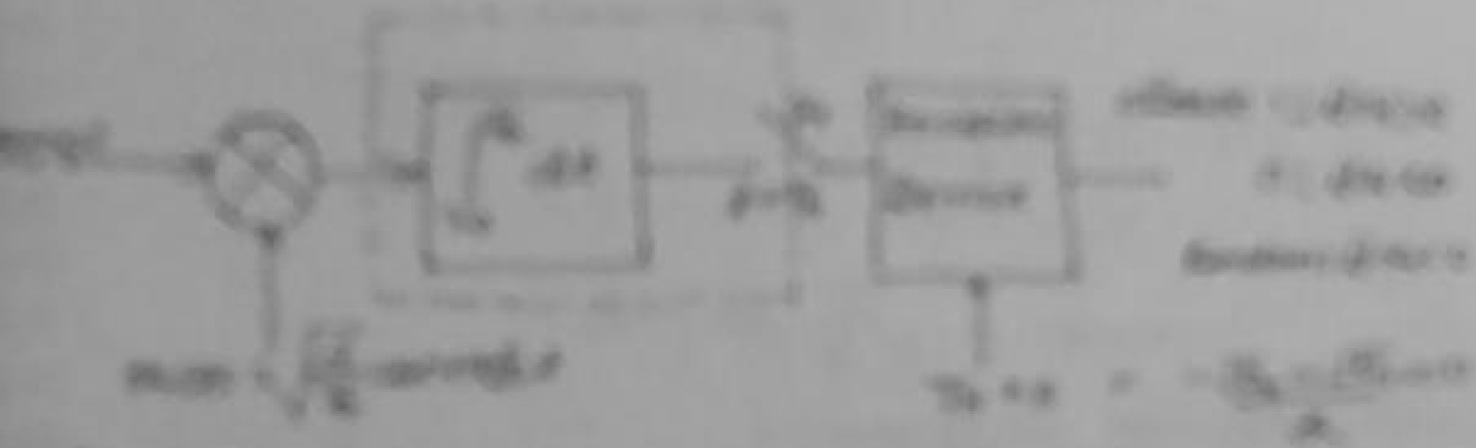
Probability of error:--

$$P_e = \frac{1}{2}\,\mathrm{erfc}\sqrt{\frac{d_{12}^2}{4\eta}}$$

$$d_{12} = 2\sqrt{E_b}$$

$$\boxed{\therefore P_e = \frac{1}{2}\,\mathrm{erfc}\left[\sqrt{\frac{E_b}{\eta}}\right]}$$





$$x_1 = \int_0^{T_b} -\sqrt{E_b}\,\sqrt{\frac{2}{T_b}}\cos 2\pi f_c t \cdot \sqrt{\frac{2}{T_b}}\cos 2\pi f_c t \; dt$$

$$\boxed{x_1 = -\sqrt{E_b}}$$

Binary Frequency shift keying (BFSK):-

→ This form of digital modulation is an example of continuous phase FSK. In BFSK system 1's & 0's are distinguished from each other by transmitting one of the two sinusoidal wave that differ in frequency by a fixed amount

In BFSK the transmitted signal is given by-

$$s_i(t) = \sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_i t \;; \rightarrow i = 1, 2$$

Must

$$s_1(t) = \sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_1 t \longrightarrow \text{'0'} \qquad \boxed{f_1 = n f_b}$$

$$s_2(t) = \sqrt{\frac{2E_b}{T_b}} \cos 2\pi f_2 t \longrightarrow \text{'1'} \qquad \boxed{f_2 = n \cdot f_b}$$

$$S_{11} = \sqrt{E_b}$$

$$u_1(t) = \sqrt{\frac{2}{T_b}} \cos 2\pi f_1 t$$

$$S_{21} = 0$$

{ Different signals }

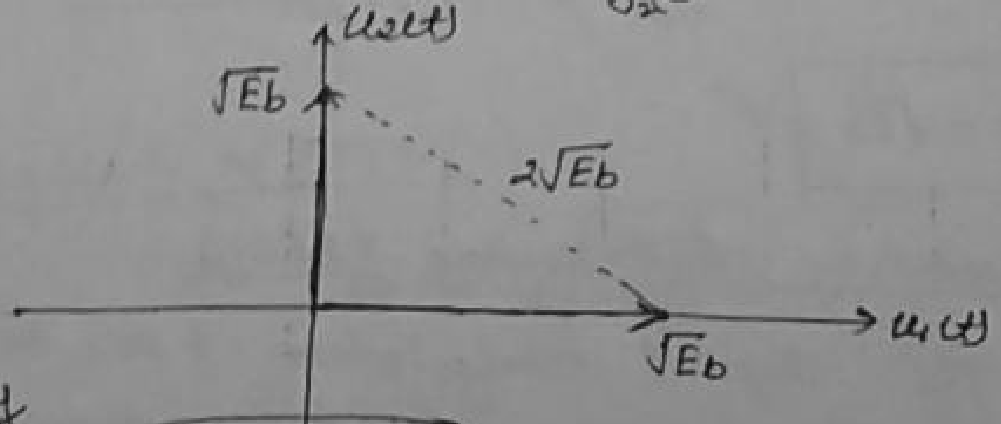
$$S_{22} = \sqrt{E_b}$$

$$u_2(t) = \sqrt{\frac{2}{T_b}} \cos 2\pi f_2 t$$

Constellation Diagram:-

$$S_1(t) = \underbrace{\sqrt{E_b}} \times \underbrace{\sqrt{\frac{2}{T_b}} \cos 2\pi f_1 t}_{u_1(t)}$$

$$S_2(t) = \underbrace{0 + \sqrt{E_b}} \times \underbrace{\sqrt{\frac{2}{T_b}} \cos 2\pi f_2 t}_{u_2(t)}$$

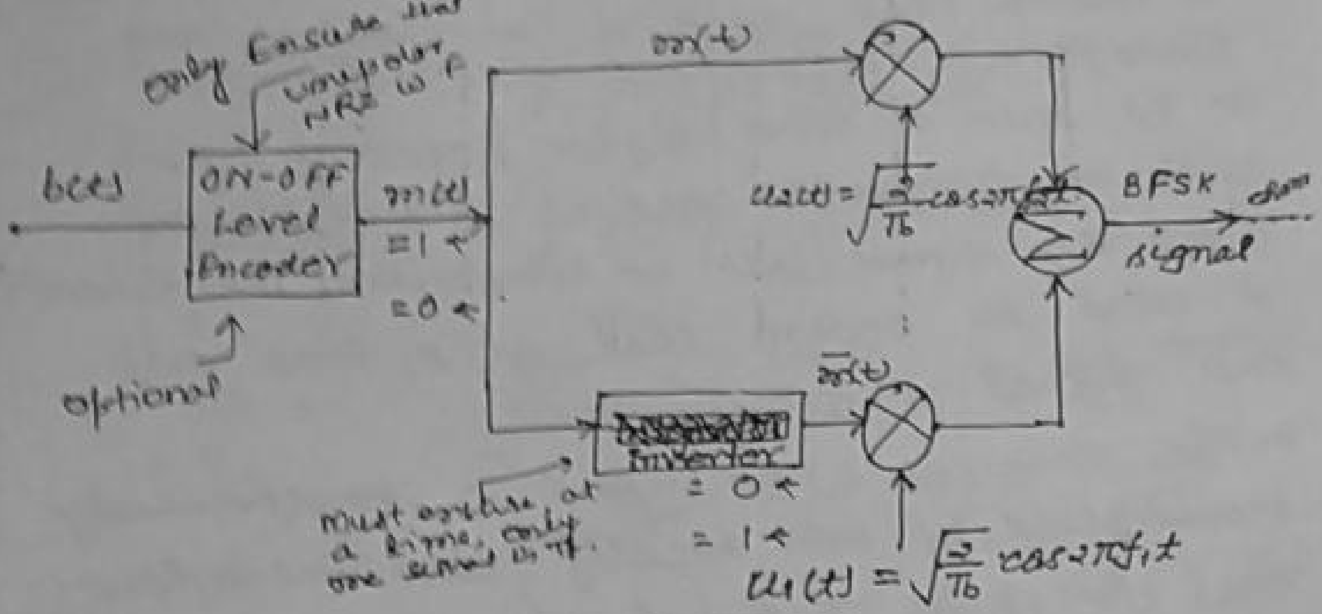


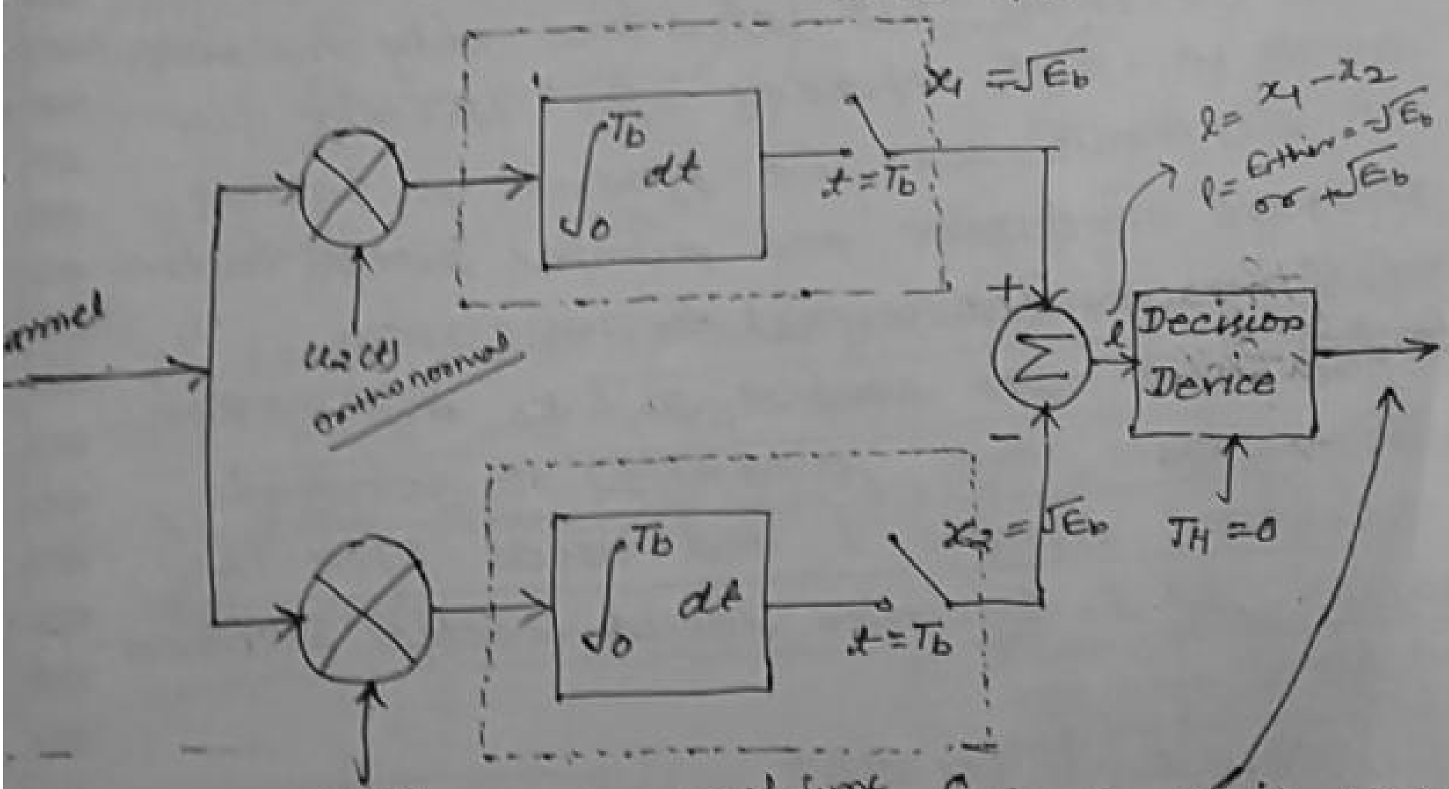
$$d_{12} = 2\sqrt{E_b}$$

Probability of error:

$$\boxed{P_e = \frac{1}{2} \operatorname{erfc}\left(\sqrt{\frac{E_b}{2\eta}}\right)}$$

Generation and Detection of BFSK signal:-





Two orthonormal func product = 0

For '1':

$\rightarrow \sqrt{Eb} \times \sqrt{\frac{2}{Tb}} \underbrace{\cos 2\pi f_2 t \cdot \cos 2\pi f_1 t}_{0} \times \sqrt{\frac{2}{b}}$

Choose 1 ; if $l > 0$

0 ; if $l < 0$

Random ; if $l = 0$

$\boxed{x_2 = 0}$

$\rightarrow \sqrt{Eb} \times \sqrt{\frac{2}{Tb}} \underbrace{\cos 2\pi f_2 t \cdot \cos 2\pi f_2 t}_{\cos^2 2\pi f_2 t}$

$\boxed{x_1 = \sqrt{Eb}}$

31/01/2010 → The binary data sequence b(t) is first applied to an optional on-off level encoder to ensure that m(t) is an unipolar NRZ waveform.

→ If m(t) = 1 then higher frequency signal will be transmitted and if m(t) = 0, lower frequency signal will be transmitted. The inverter is used to ensure that at a time only one signal is transmitted.

→ The received BFSK signal is simultaneously multiplied by both locally generated orthonormal functions but at a time only one output will be high depending upon whether "1" or "0" was transmitted. The product is passed through integrator and dumped switch receiver which maximized signal to noise ratio.

→ At $t = T_b$ two samples x_1 & x_2 are taken. The samples are subtracted to generate resultant sample l and based upon the value of l, Decision device decides between '0' & '1'.

$$u_1(t) = \sqrt{\frac{2}{T_b}} \cos 2\pi f_1 t$$

$$u_2(t) = \sqrt{\frac{2}{T_b}} \cos 2\pi f_2 t$$

$$\int_0^{T_b} \cos \omega_2 t \cdot \cos \omega_1 t \, dt = 0$$

imp for calculating numerical problems

Gate

$$= \int_0^{T_b} \cos(\omega_2 + \omega_1) t \, dt + \int_0^{T_b} \cos(\omega_2 - \omega_1) t \, dt = 0$$

$$\therefore \boxed{\begin{aligned} \omega_2 + \omega_1 &= n\omega_b \\ \omega_2 - \omega_1 &= m\omega_b \end{aligned}} \qquad \boxed{\begin{aligned} f_2 + f_1 &= nf_b \\ f_2 - f_1 &= mf_b \end{aligned}}$$

Comparison
$$\begin{cases} P_e(\text{ASK}) = \frac{1}{2}\,\mathrm{erfc}\left[\sqrt{\frac{E_b}{4\eta}}\right] \\ P_e(\text{BFSK}) = \frac{1}{2}\,\mathrm{erfc}\left[\sqrt{\frac{E_b}{2\eta}}\right] \\ P_e(\text{BPSK}) = \frac{1}{2}\,\mathrm{erfc}\left[\sqrt{\frac{E_b}{\eta}}\right] \rightarrow \text{Best} \end{cases}$$

$\frac{E_b}{\eta} = \frac{\text{watt/sec}}{\text{watt/Hz}} = \frac{\text{watt}}{\text{watt}}$ → $\frac{\text{Signal Power}}{\text{Noise Power}}$ → $\frac{S}{N}$ ratio

PSD $= \frac{\eta}{2}$

Power half
$$\begin{cases} \frac{E_b}{\eta} = 0\ dB \text{ only for BPSK} \\ \frac{E_b}{2\eta} \longrightarrow -3\ dB \end{cases}$$

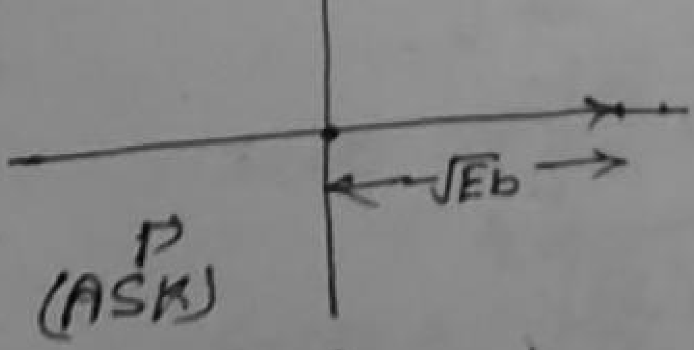
$\frac{S}{N}$ min $\searrow \frac{E_b}{4\eta} \longrightarrow -6\ dB$

* more $\frac{S}{N}$ ratio ... less will be P_e

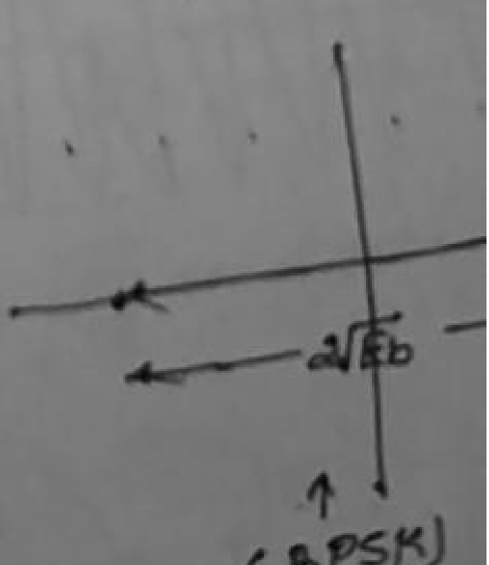
→ In BPSK signal to noise ratio is 3 dB better than BFSK and 6 dB better than ASK.

→ In BFSK signal to noise ratio is 3 dB better than ASK.

* Distance b/w two ... then Prob. ... will be ...



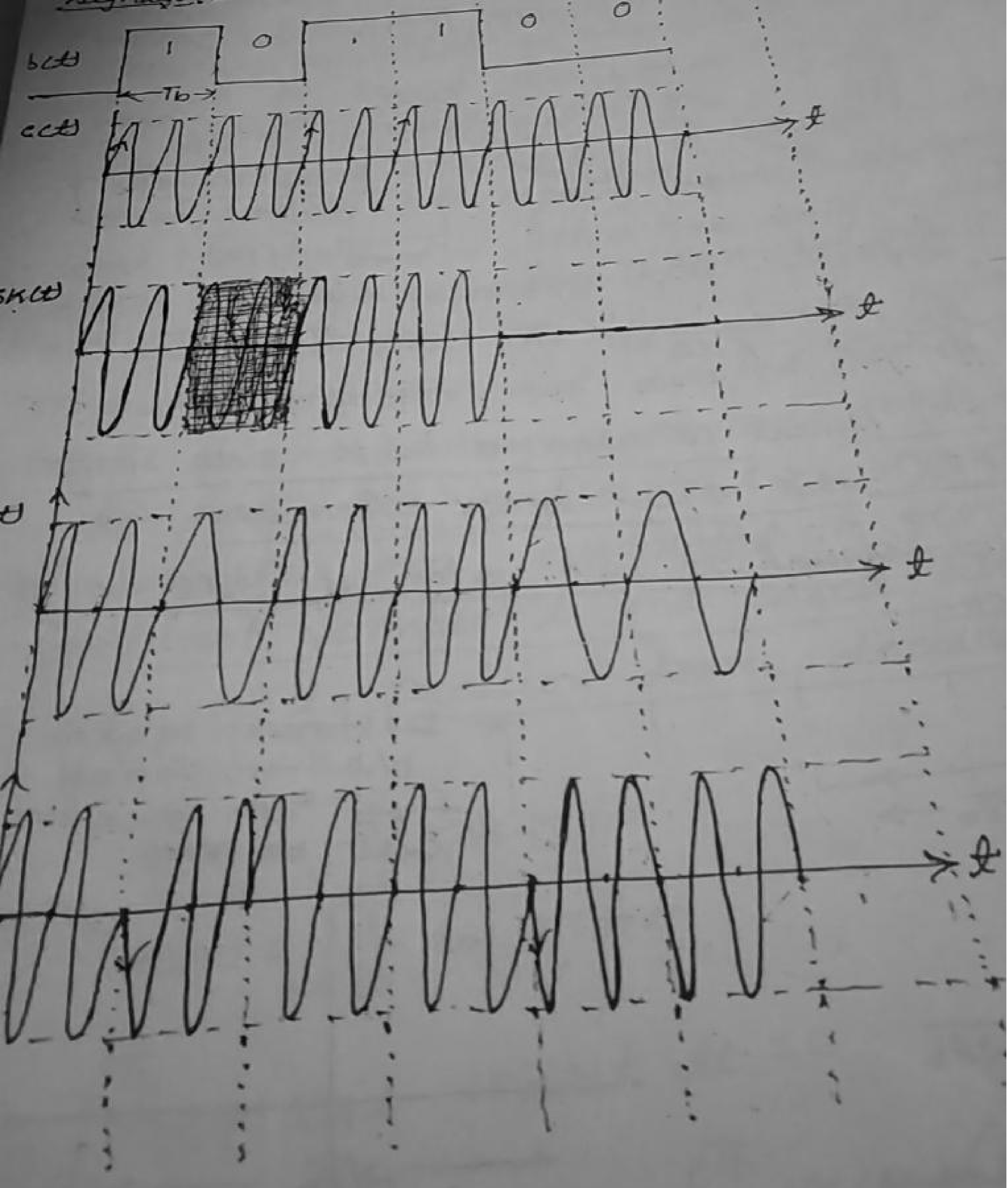
(ASK)



(BPSK)

to noise ratio increases ... error decreases."

Time Domain Representation of ASK, FSK, ... signals:-



Quadrature Phase Shift Keying (QPSK)

→ In QPSK as with binary PSK, information carried by transmitted signal is contained in the phase. In particular the phase of carrier takes on one of four equally spaced values such as $\frac{\pi}{4}, \frac{3\pi}{4}, \frac{5\pi}{4}$ & $\frac{7\pi}{4}$. and the transmitted signal is given by-

* Double data rate → no. of bits/sec

$$s_i(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + (2i-1)\frac{\pi}{4}\right] : i = 1,2,3,4$$

* Two BPSK signals in a single bit duration

T_b

Two sample or

more Tr. BW is required

$$\left.\begin{aligned} E_s &= 2E_b \\ T_s &= 2T_b \end{aligned}\right\} \text{symbol form}$$

$$s_1(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + \frac{\pi}{4}\right]$$

$$s_2(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + \frac{3\pi}{4}\right]$$

$$s_3(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + \frac{5\pi}{4}\right]$$

$$s_4(t) = \sqrt{\frac{2E_s}{T_s}} \cos\left[2\pi f_c t + \frac{7\pi}{4}\right]$$

$$\therefore s_1(t) = (+)\sqrt{\frac{E_s}{T_s}} \cos 2\pi f_c t \ (-) \sqrt{\frac{E_s}{T_s}} \sin 2\pi f_c t \rightarrow \text{'1 0'}$$

+ve, -ve

$$s_2(t) = (-)\sqrt{\frac{E_s}{T_s}} \cos 2\pi f_c t \ (-) \sqrt{\frac{E_s}{T_s}} \sin 2\pi f_c t \rightarrow \text{'0 0'}$$

-ve

$$s_3(t) = -\sqrt{\frac{E_s}{T_s}} \cos 2\pi f_c t + \sqrt{\frac{E_s}{T_s}} \sin 2\pi f_c t \rightarrow \text{'0 1'}$$

$$s_4(t) = \sqrt{\frac{E_s}{T_s}} \cos 2\pi f_c t + \sqrt{\frac{E_s}{T_s}} \sin 2\pi f_c t \rightarrow \text{'1 1'}$$

'1 0' $\longrightarrow \frac{\pi}{4}$

'0 0' $\longrightarrow \frac{3\pi}{4}$

'0 1' $\longrightarrow \frac{5\pi}{4}$

'1 1' $\longrightarrow \frac{7\pi}{4}$

$\pi/2$ $\pi/2$ $\pi/2$ $\pi/2$ π π

... when you ... there is a phase shift of 90° in the resultant signal and when both bits changes simultaneously then there is a corresponding phase shift of 180° in the resultant signal"

Constellation Diagram:

$\boxed{S_3(t) = -S_1(t)}$

$\boxed{S_4(t) = -S_2(t)}$

$u_1(t) = \sqrt{\frac{2}{T_s}} \cos 2\pi f_c t$

$u_2(t) = \sqrt{\frac{2}{T_s}} \sin 2\pi f_c t$

Two linearly independent & Two are linearly independent. So two orthonormal functions are required.

* compare the S_1, S_2, S_3, S_4 to the QPSK signals

$S_{11} = \sqrt{\frac{E_s}{2}}$

$S_{12} = -\sqrt{\frac{E_s}{2}}$

$S_{21} = -\sqrt{\frac{E_s}{2}}$

$S_{22} = -\sqrt{\frac{E_s}{2}}$

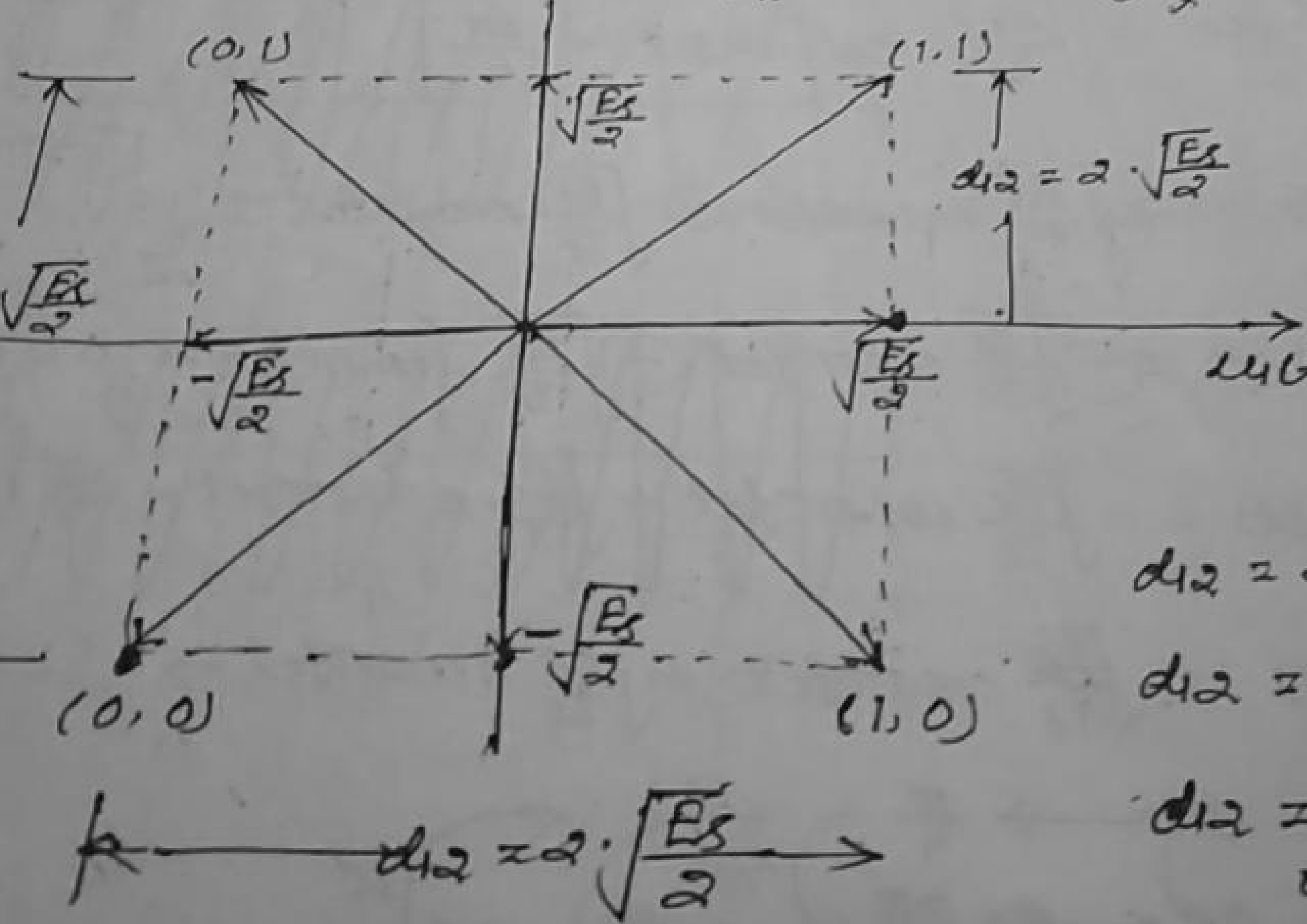
$S_{31} = -\sqrt{\frac{E_s}{2}}$

$S_{32} = \sqrt{\frac{E_s}{2}}$

$S_{41} = \sqrt{\frac{E_s}{2}}$

$S_{42} = \sqrt{\frac{E_s}{2}}$

$u_2(t) = \sqrt{\frac{2}{T_s}} \sin 2\pi f_c t$ * $P_e = \frac{1}{2} \operatorname{erfc}\left[\sqrt{\frac{E_b}{\eta}}\right]$



$d_{12} = 2 \times \sqrt{\frac{E_s}{2}}$

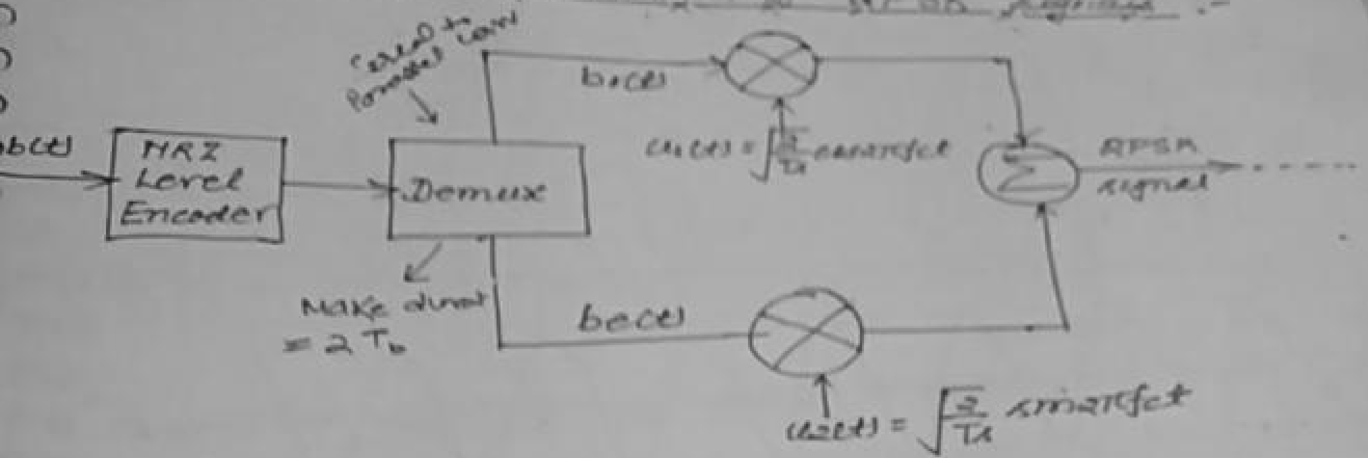
$d_{12} = \sqrt{2E_s}$

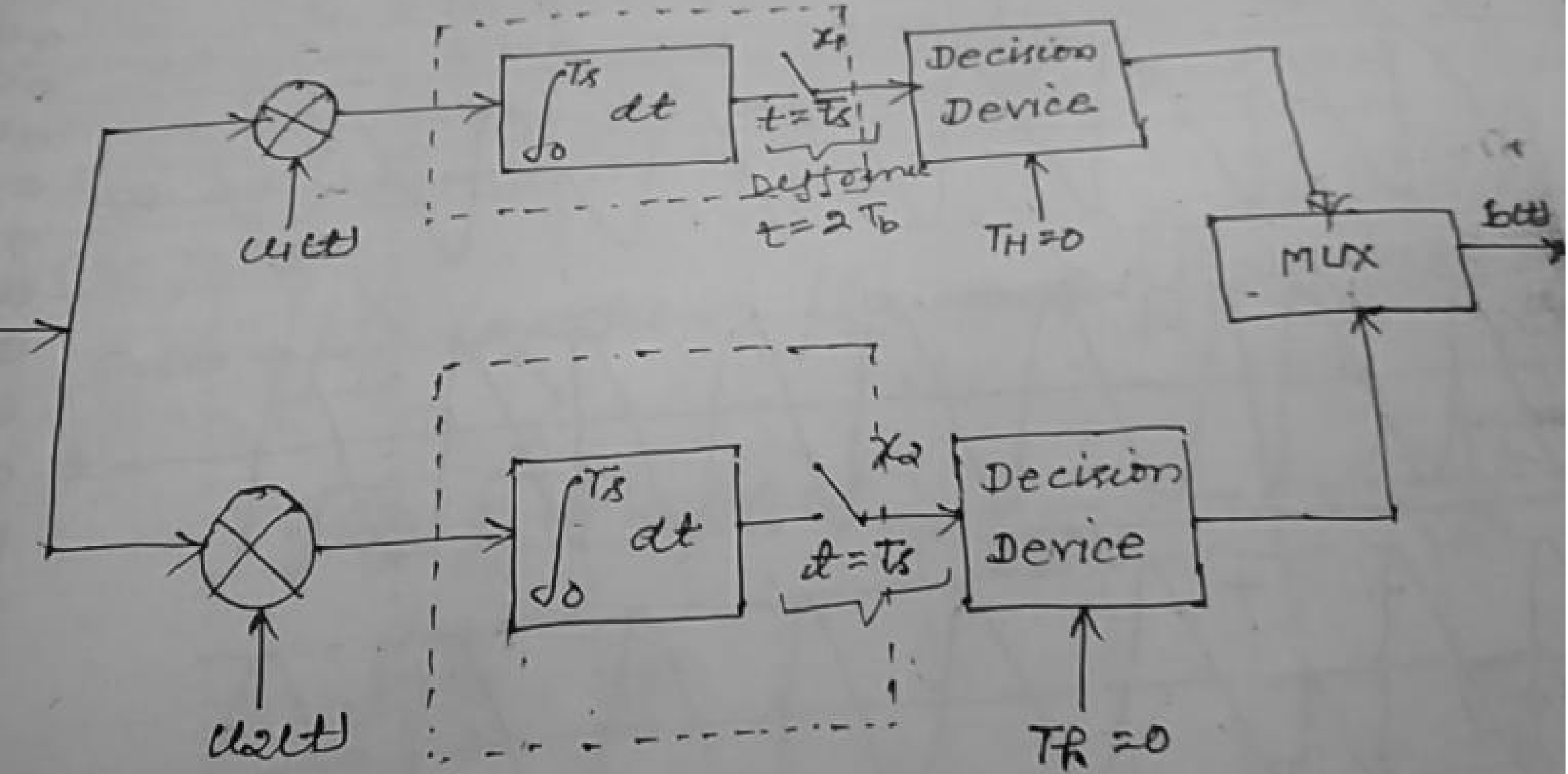
$d_{12} = 2\sqrt{E_b}$

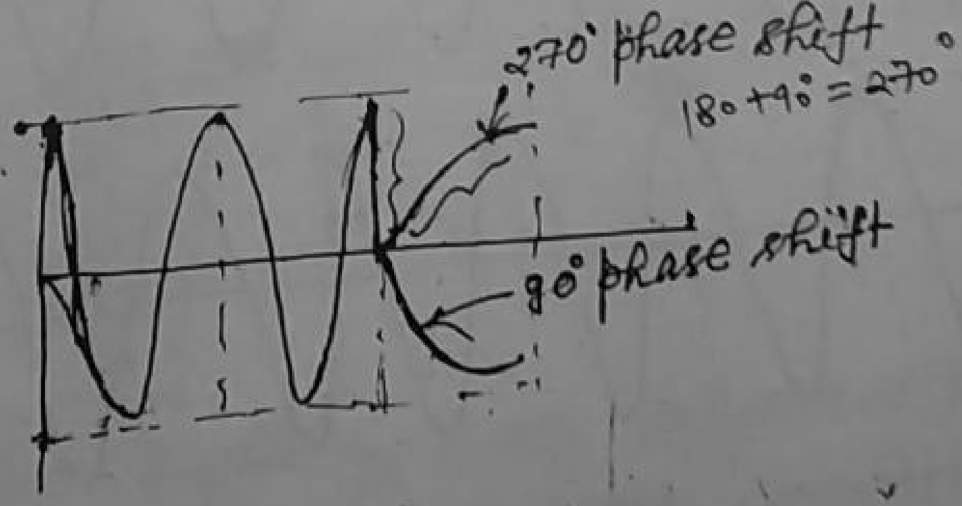
or

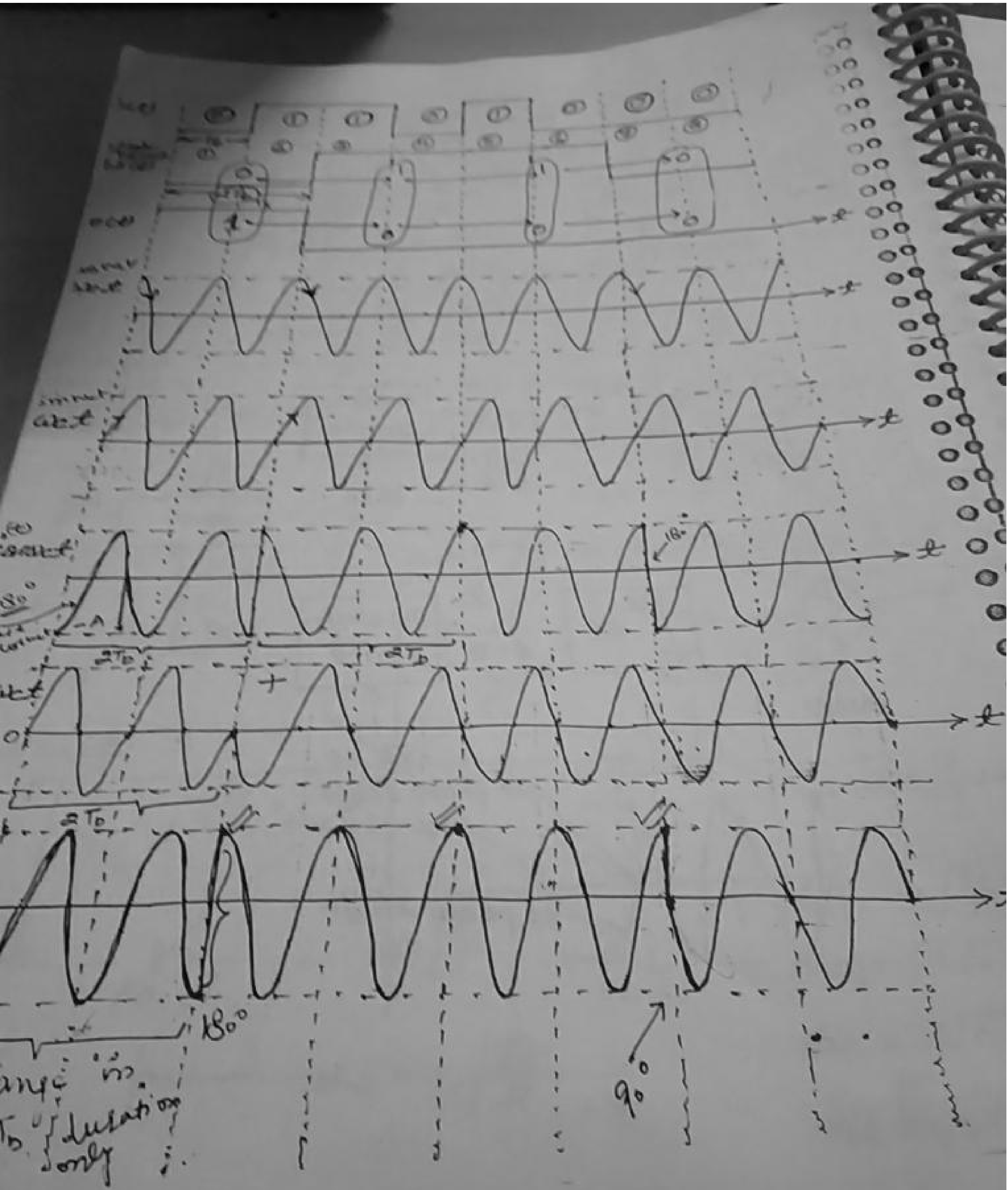
$d_{12} = \sqrt{4E_b}$

Generation and detection of QPSK signals :-

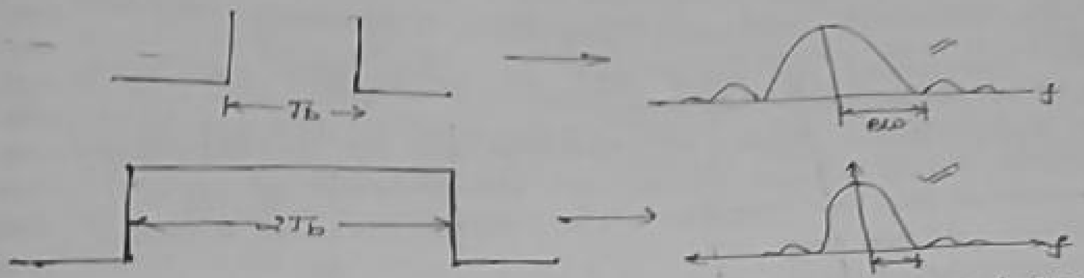








2Tb
2Tb
2Tb
180°
180°
90°
ange in
Tb duration
only



Hence BW = $\frac{1}{2}$ BW

→ "In QPSK for same data transmission rate only half the transmission bandwidth is required as compared to BPSK and alternatly for same transmission bandwidth double the ~~data~~ data rate can be achieved."

For 10:

$$S_1(t) = \left[\sqrt{\frac{E_s}{T_s}} \cos 2\pi f_c t - \sqrt{\frac{E_s}{T_s}} \sin 2\pi f_c t\right] \sqrt{\frac{2}{T_s}} \cos 2\pi f_c t$$

$$= \sqrt{\frac{2E_s}{T_s}} \int_0^{T_s} \cos^2 2\pi f_c t \, dt$$

$$= \sqrt{\frac{2E_s}{T_s}} \cdot \frac{T_s}{2}$$

$$= \sqrt{\frac{E_s}{2}}$$

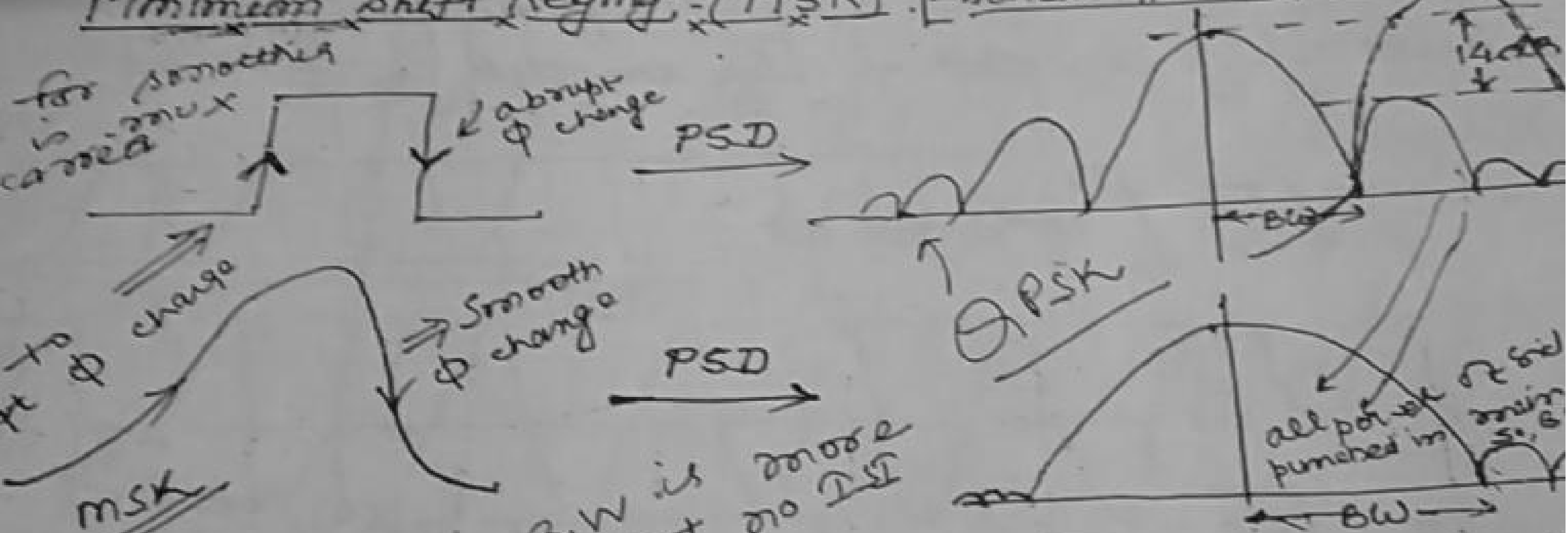
$$\boxed{x_1 = \sqrt{\frac{E_s}{2}}}$$

x_2 will also be $\sqrt{\frac{E_s}{2}}$

Working: → The binary data sequence b(t) which is unipolar NRZ waveform is first converted into bipolar NRZ waveform by passing it through a NRZ level encoder.

... demultiplexer program two ...
→ It converts bit stream bits into even and odd part bits.
→ The duration of each bit in even and odd part has a duration of $2T_b$.
The bit stream $b_e(t)$ is multiplied by orthonormal function $\phi_1(t)$ and $b_o(t)$ is multiplied by $\phi_2(t)$.
The addition of these two waveform is transmitted as QPSK signal.
→ The received QPSK signal is simultaneously multiplied with locally generated orthonormal function $\phi_1(t)$ & $\phi_2(t)$. The integrator and dumped switch receiver maximises SNR and the samples are taken at $t = T_b$
→ Samples x_1 & x_2 are applied to the decision device which decides accordingly b/w '0' & '1' and finally the odd & even bit stream are fed to a multiplexer which converts the parallel data into serial o/p binary sequence bits.

Minimum Shift Keying :- (MSK) :



$\boxed{(BW)_{MSK} = 1.5(B.W)}$

QPSK due to abrupt phase change as ...
...als the level of side lobe power incr...
...fically it has been determined that ...
...rence b/w highest point of main lobe ...
...lobe level is only 14 dB. So there ...