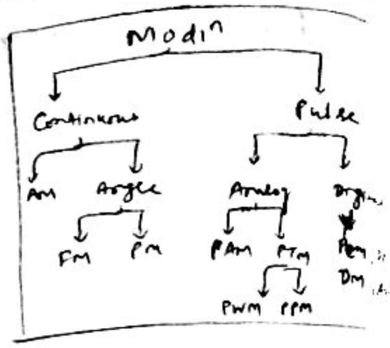
① (Singh, Sapre - Ch-7,8) ②

Module - III - : Pulse Modulation System :-

In pulse modulation systems the carriers are no longer continuous in nature but consist of several pulse form. So in pulse modulation the parameters of pulse are varied in accordance with instantaneous values of modulating signal



Pulse Mod^n are of 2 types.

- Pulse Amplitude modulation (PAM)
- Pulse Time Modulation (PTM)

PAM :- In PAM scheme the amplitude of the pulses of carrier signal is varied according to the modulating signal.

PTM :- In PTM scheme the timing of the pulses of the carrier signal are varied.

Again PTM scheme are of 2 types

(a) PWM (Pulse width modulation) / or Pulse duration mod^n (PDM) / or Pulse length modulation (PLM)

(b) PPM (Pulse pos^n mod^n)

PAM :-

The sampling the^m states that if a modulating signal is band limited to 'B' Hz then the sampling frequency & frequency of carrier signal are same at '2B' Hz.

So for PAM, the frequency of the carrier is decided by the sampling the^m.

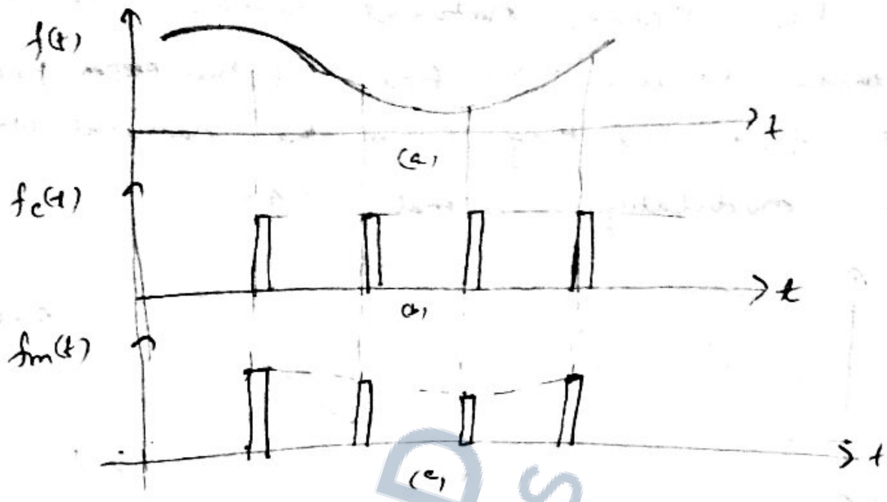


fig (a) shows baseband signal.

(b) " carrier pulse train.

(c) " pulse amplitude modulated signal.

→ Due to discrete on time axis and continuous in amplitude axis the PAM signal $f_m(t)$ is also called discrete time signal.

→ The baseband signal can have both positive as well as –ve polarity. As the transmission of such bipolar pulses is inconvenient, a clamping ckt is used so that the baseband signal with +ve polarity is ensured.

PAM can be obtained in 2 ways

(a) Natural sampling / Shaped-to sampling.

(b) Flat-top sampling.

(a) Natural sampling :-

Consider fig (2). Here we have assumed the amplitude of the carrier signal is 1, having duration τ and separated by T_s. By multiplying $f(t)$ & $f_c(t)$ of fig (2) we can get PAM signal $f(t).f_c(t)$.

(3) The name natural sampling is given to this method because the tops of the PAM signal are not flat but they follow the natural waveform of the modulating signal $f(t)$.

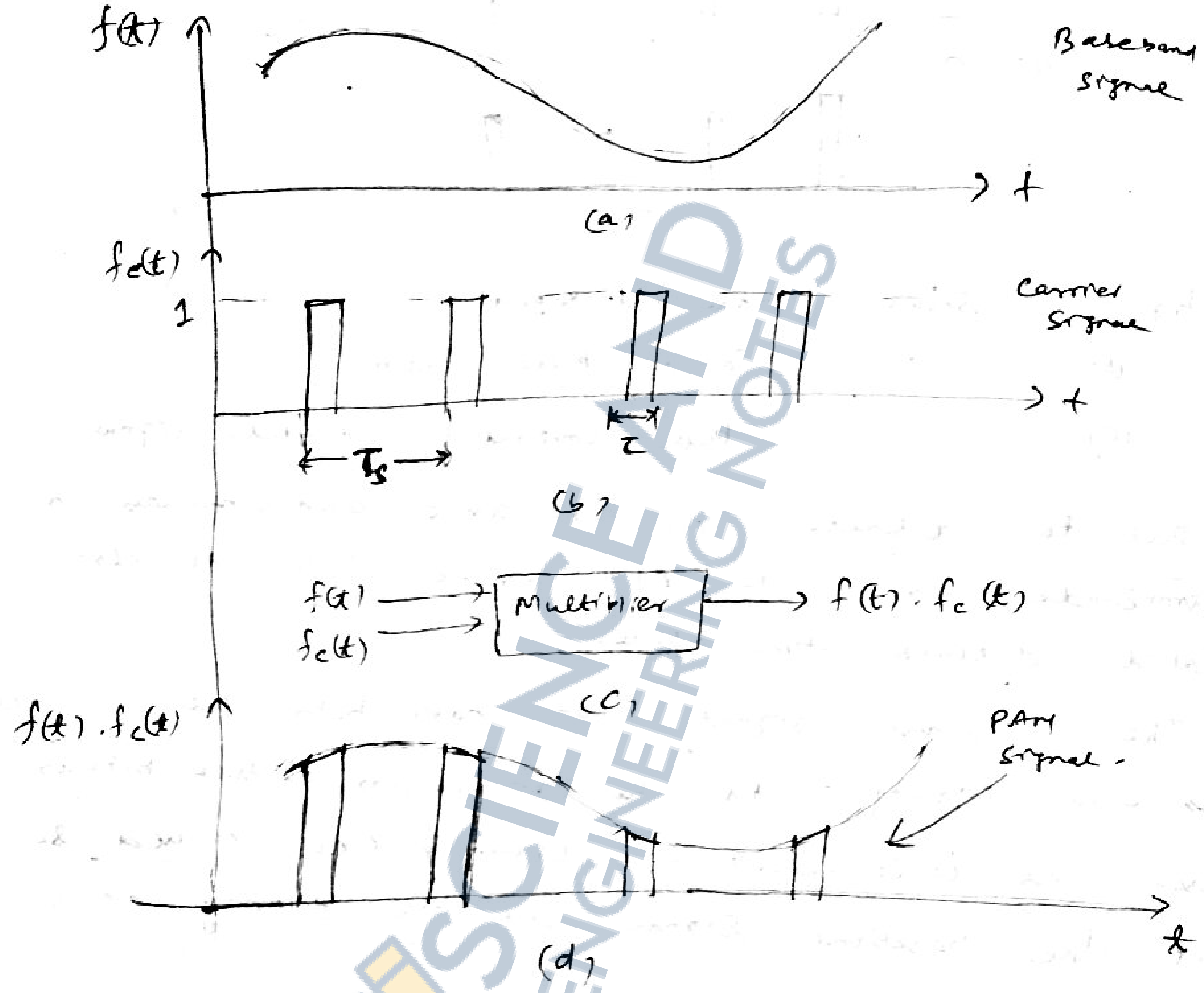


fig (2)

We know the Fourier series of a periodic pulse train is

$$V(t) = \frac{A\tau}{T_0} + \frac{2A\tau}{T_0} \sum_{n=0}^{\infty} C_n \cos \frac{2\pi n t}{T_0} \qquad (1)$$

where A = Amplitude of pulse

τ = duration of pulse

T_0 = period of pulse.

$$C_n = \frac{\sin(n\pi\tau/T_0)}{(n\pi\tau/T_0)}$$

For the carrier pulse train, we have

$v(t) = f_c(t)$

$A = 1,$

$T_0 = T_s$

$\therefore$ Eq^n (1) becomes,

$$f_c(t) = \frac{\tau}{T_s} + \frac{2\tau}{T_s} \sum_{n=0}^{\infty} C_n \cos\left(\frac{2\pi n t}{T_s}\right)$$

where $C_n = \dfrac{\sin(n\pi\tau/T_s)}{n\pi\tau/T_s}$

$$f_c(t) = \frac{\tau}{T_s} + \frac{2\tau}{T_s}\left[C_1 \cos\left(\frac{2\pi t}{T_s}\right) + C_2 \cos 2\times 2\pi \frac{t}{T_s} + \cdots\right]$$

The o/p of the multiplier is

$$f(t).f_c(t) = \frac{\tau}{T_s} f(t) + \frac{2\tau}{T_s}\left[f(t)\, C_1 \cos 2\pi \frac{t}{T_s} + f(t)\, C_2 \cos 2\times 2\pi\frac{t}{T_s} + \cdots\right] \quad \text{--- (2)}$$

By using sampling them^m, we have

$$T_s = \frac{1}{2f_m} \qquad \left[\begin{array}{l} f_s = 2f_m \\ \Rightarrow T_s = \frac{1}{2f_m} \end{array}\right]$$

f_m = max^m frequency component in $f(t)$.

Substituting $T_s = \frac{1}{2f_m}$ in eq^n (2), we have.

$$f(t).f_c(t) = \underbrace{\tau \times 2f_m . f(t)}_{\text{first term}} + \underbrace{2\tau \times 2f_m \left[f(t)\, C_1 \cos(2\pi t . 2f_m) + f(t)\, C_2 \cos(2\times 2\pi t \times 2f_m) + \cdots \right]}_{\text{2nd term}} \quad \text{--- (3)}$$

→ Neglecting the multiplication factor, then the first term in eq^n (3) is the base band $f(t)$ itself.

→ The second term is product of $f(t)$ and a sinusoidal frequency component $2f_m$

i.e $f(t) . \cos(2\pi.(2f_m)\,t)$

(5)

If $f(t) = A_m \cos 2\pi f_m t$

then $f(t) \cdot \cos 2\pi (2f_m) t$ will give

$$A_m \cos 2\pi (f_m) t \cdot \cos 2\pi (2f_m) t$$

$$= \frac{A_m}{2} \cdot 2 \underbrace{\cos 2\pi (f_m) t}_{\beta} \cdot \underbrace{\cos 2\pi (2f_m) t}_{\alpha}$$

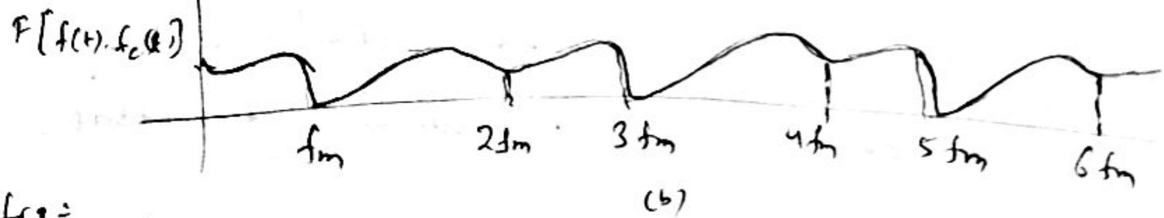
$$= \frac{A_m}{2} \left[\cos \left[2\pi (2f_m + f_m) t \right] + \cos \left[(2f_m - f_m) \right] t \right]$$

$$= \frac{A_m}{2} \left[\cos (2\pi (2f_m + f_m) t) + \cos (2\pi (2f_m - f_m)) t \right]$$

→ Hence the multiplication in second term will yield the frequency spectrum given by the sum $\underline{2f_m + f_m}$ and difference $\underline{2f_m - f_m}$. Thus the spectrum of second term is from $\underline{f_m \text{ to } 3f_m}$.

→ Similarly the freq. spectrum of third term is from $4f_m - f_m = \underline{3f_m}$ to $4f_m + f_m = \underline{5f_m}$ and so on.

(a)



(b)

fig:

(a) Magnitude plot of spectral density of f(t)

(b) Magnitude plot of spectral density of $\underline{f(t) \,\&\, f_c(t)}$.

(6)

Flat top sampling :-

The electronic circuitry needed to perform natural sampling is somewhat complicated because the pulse-top shape is to be maintained. These complications are reduced by flat-top sampling. In this, in the modulated signal the pulses are flat. Thus the tops of the pulse have a const. amplitude within pulse interval.

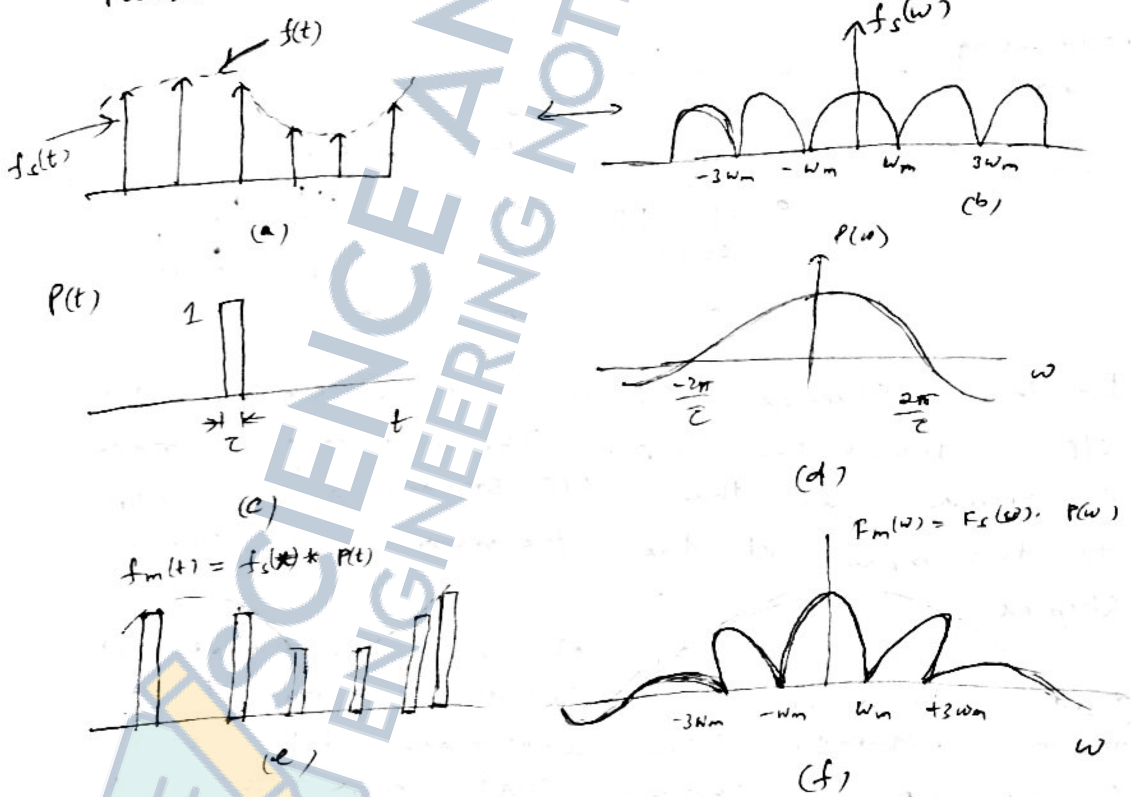


fig 1(a) Impulse sampled signal $f_s(t)$.

(b) Spectrum of $f_s(t)$

(c) Non periodic pulse $P_c(t)$ of width τ & height 1.

(d) Spectrum of P(t)

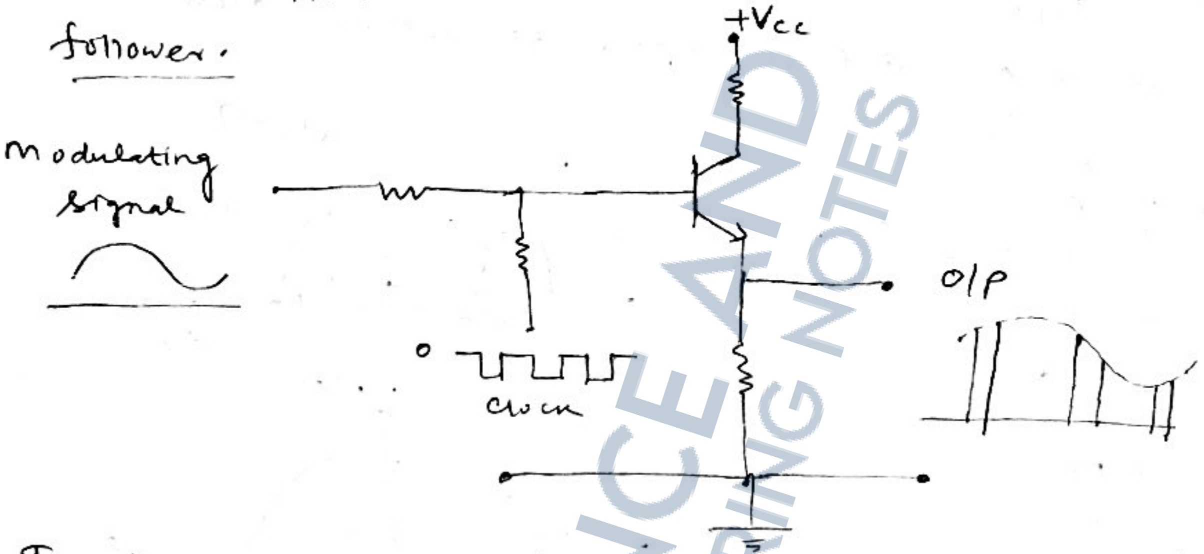
(e) Flat-top sampled PAM signal $f_m(t)$

(f) Spectrum of $f_m(t)$

The flat-top sampled signal fm(t), fig(e), may be considered as a convolution of the impulse sampled signal fs(t) [fig(a)] and non-periodic pulse p(t) of width τ & height 1 (fig(c))

A PAM Modulator Circuit :-

The PAM modulator is a simple emitter follower.



In the absence of the clock signal, the o/p follows the i/p. The modulating signal is applied as the i/p signal. Another i/p to the base of the transistor is the clock signal.

The frequency of the clock signal is made equal to the desired carrier pulse train frequency. The amplitude of the clock signal is so chosen that high level is at ground (0V) & low level at some -ve voltage which is sufficient to bring the transistor in the cut-off region.

(Thus, when the clock signal is high, the ckt behaves as an emitter follower and o/p follows the i/p modulating signal. When clock signal is low, the transistor is cut-off & o/p is zero. Thus the desired o/p (PAM) waveform is obtained at the o/p of the transistor shown in fig.)

(8) Demodulation of PAM Signal:-

Demodulation of natural sampled signal can be done with the help of an ideal LPF with cut off freq ω_m. But for this, the pulse-top shape is to be maintained after transmission. This is very difficult due to transmitter and receiver noise. Therefore, normally, flat-top sampling is preferred over natural sampling.

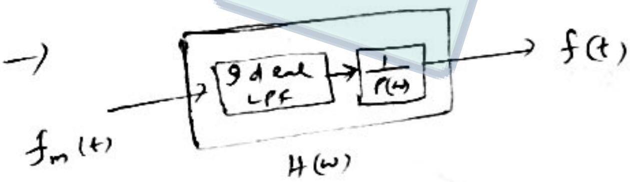
Methods for flat-top sampled signal demodulation.

(i) using an equalizer:-

→ If any flat top sampled signal is passed through an ideal LPF then the spectrum of the O/P will be $F(\omega) \cdot P(\omega)$.

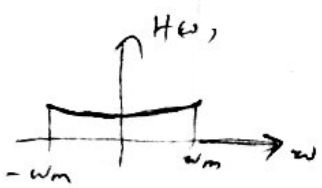
→ The time function of O/P will be distorted due to the multiplying factor $P(\omega)$.

→ If the LPF O/P is passed through a filter having transfer function $\frac{1}{P(\omega)}$ in the range $0 - \omega_m$ then the spectrum at the O/P of this filter will be $F(\omega) \cdot \cancel{P(\omega)} \cdot \frac{1}{\cancel{P(\omega)}} = F(\omega)$. The filter having transfer function $\frac{1}{P(\omega)}$ is called an equalizer.

→ 

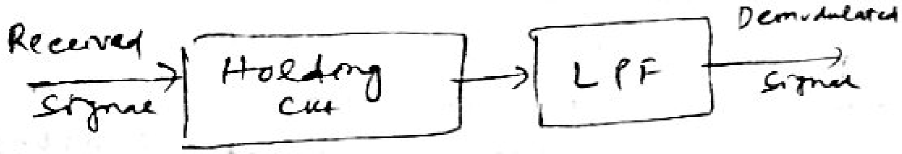
The combination of an ideal LPF and equalizer is known as composite filter.

$$H(\omega) = \begin{cases} \frac{1}{P(\omega)}, & |\omega| < \omega_m \\ 0, & \text{otherwise} \end{cases}$$

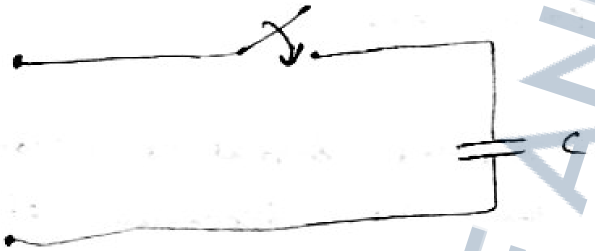


(iii) Using Holding circuit:-

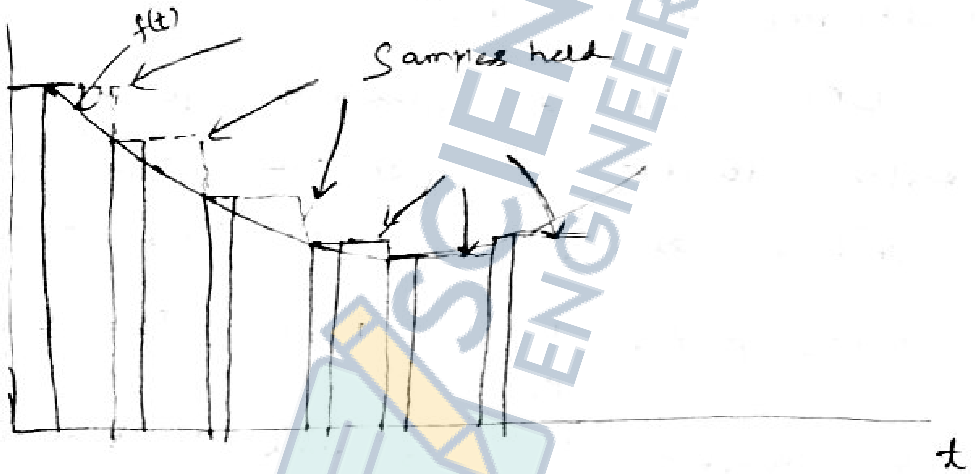
In this case the received signal is passed through a holding ckt and a LPF as shown in fig.



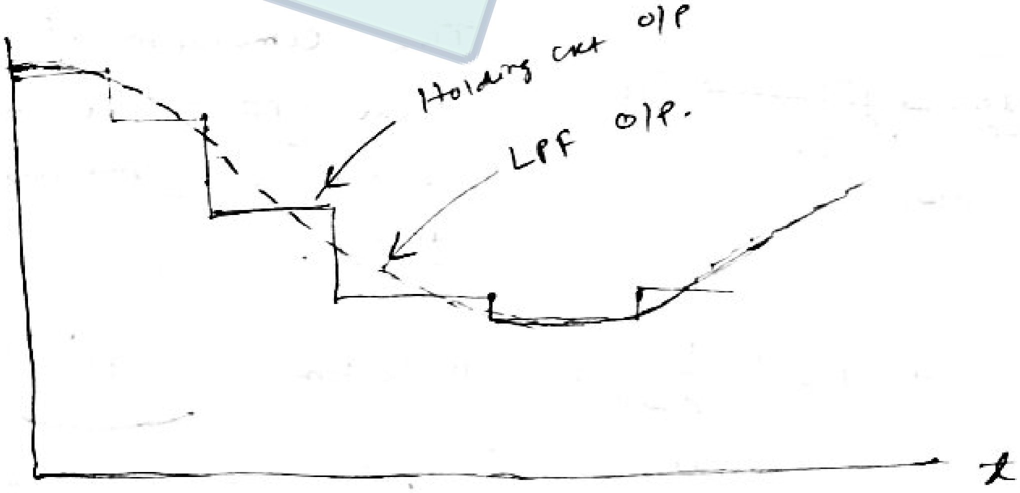
(a) Building blocks of demodulator.



(b) Zero order Holding ckt.



(c) O/P of Holding ckt



(d) O/P of LPF

Holding circuit:-

received signal is passed ... and a LPF as shown in fig.

LPF → Demodulated signal

of demodulator.

= C

ckt.

t

o/p

i/p.

t

10

→ The sample & hold ckt is shown in fig (b). After arrival of pulse the switch S closes and opens at the end of pulse.

→ The capacitor 'C' is charged to the pulse amplitude value & holds this value during the interval between 2 pulses.

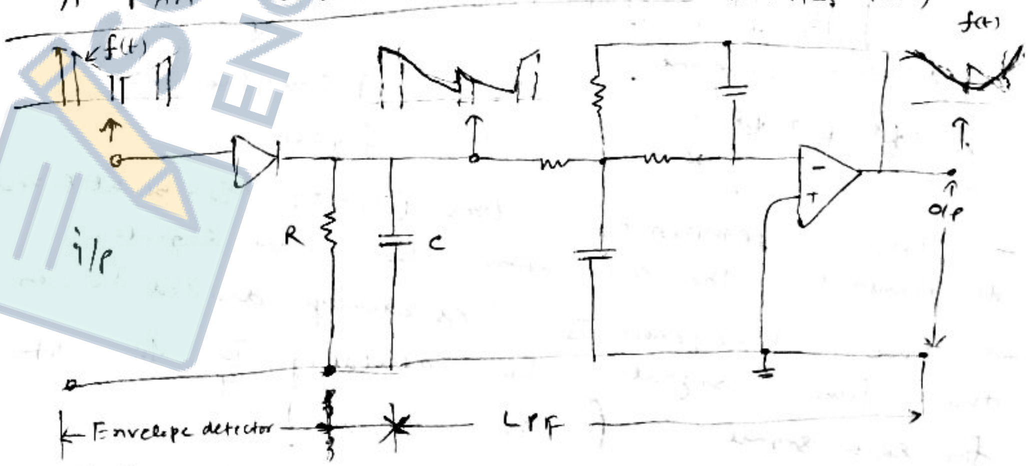
→ So sampled values are held as shown in fig (c).

→ The o/p of holding ckt is smoothened by LPF as shown in fig (d). Also some distortion is introduced due to the holding ckt.

→ Fig 4.7(b) is also called zero order holding ckt which considers only the previous sample to decide the value betn 2 pulses.

→ The first order holding ckt considers the previous 2 samples, the 2nd order holding ckt considers previous 3 samples & so on. As the order of the holding ckt increases the distortion decreases at the cost of the circuit complexity.

A PAM Demodulator ckt:- (Demodulation of Natural sampled PAM)

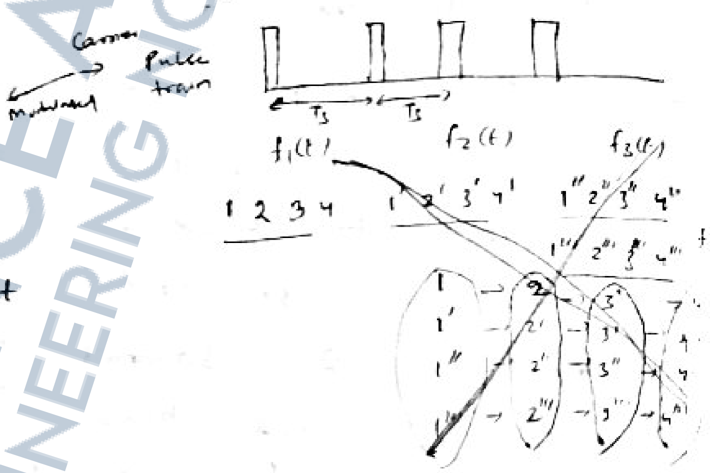


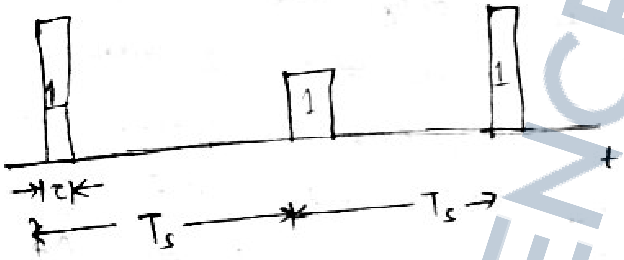
It is just an envelope detector followed by a low pass filter. The diode and R-C combination works as the envelope detector.

This is followed by a second order OP-AMP LPF to have a good filtering characteristics. Thus for the received pulse amplitude modulated signal as the i/p signal, the desired demodulated signal [i.e baseband f(t)] is the o/p.

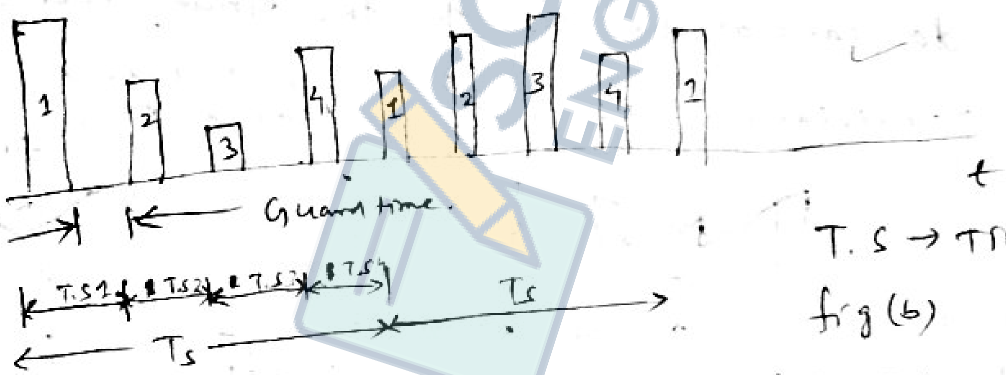
TDM (Time division multiplexed) PAM system

→ (In a PAM system the pulse duration (τ) is less than time period of pulse (T_s) i.e $\tau << T_s$. [shown in fig(a)]. Due to this no the system for information is transmitted through most of time.)





(a)



T.S → Time slot

fig (b)

→ (The remaining time ($T_s - \tau$) can be utilized to transmit the information from other signals.)

→ The time period T_s is equally divided between the four signals. So allocating $\frac{T_s}{4}$ time slot for each signal. (for fig (b))

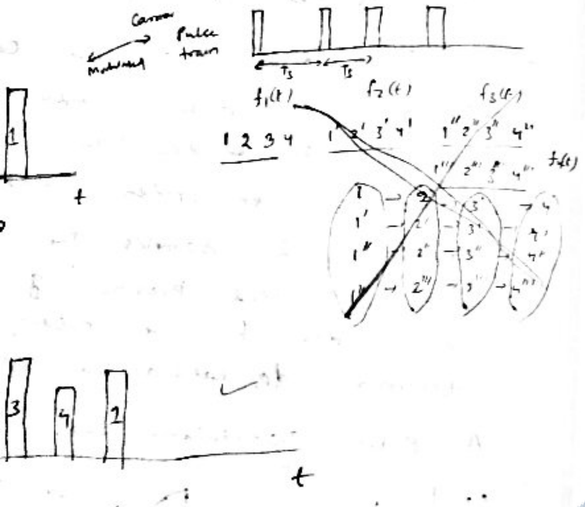
→ The duration of each slot is such that $\frac{T_s}{4} > \tau$.

→ The duration $\frac{T_s}{4} - \tau$ is called guard time

...llowed by a second order ... to have a good ...racteristics. Thus for the ... amplitude modulated signal ... signal, the desired demodulated ...band f(t)] is the O/P.

...ion multiplexed) PAM System

... system the pulse duration (τ) ...me period of pulse (T_s) i.e ... in fig(a)]. Due to this no ... ansmitted through the system for



T.S → Time slot

fig (b)

... time ($T_s - \tau$) can be utilized ...rmation from other signals.)

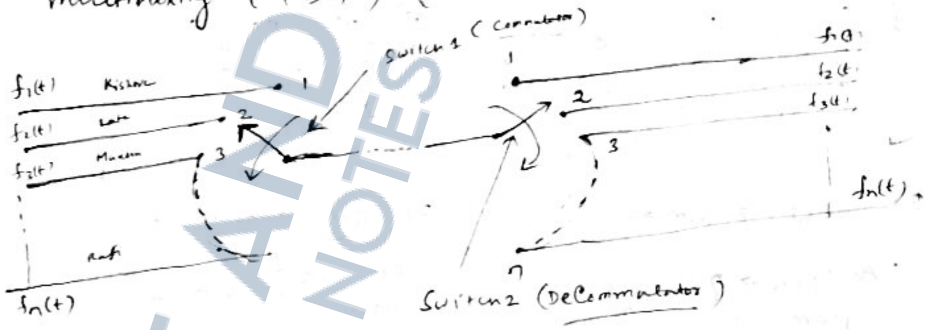
...s is equally divided between ... So allocating $\frac{T_s}{4}$ time slot ... for fig (b))

... of each slot is such that

...$s - \tau$ is called guard time

(12) between all successive sampling pulses.

→ The arrangement by which the information from more than one signal is transmitted in this manner is called Time division multiplexing (TDM) (shown in the fig. below)



(A TDM PAM System)

→ The ckt is used to transmit information from n signals. The switch I & switch II are known as commutators & decommutators. These 2 switches rotates at the same speed $2f_m$ rotations per second.

→ The commutator samples and combines the samples, while decommutator separates the samples belonging to individual signals.

→ To provide synchronization a synchronizing pulse is transmitted in every frame (time interval between 2 successive samples of the same signal i.e T_s.)

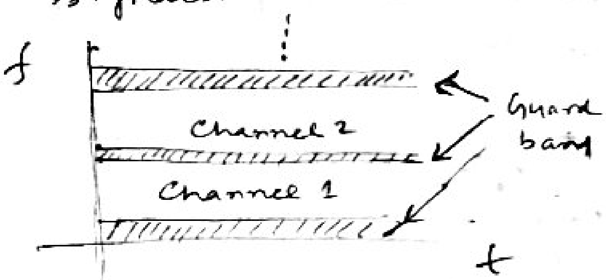
→ Thus to multiplex n channels, (n+1) time slots are provided in a frame i.e n for channels and 1 for synchronizing pulse.

[1 2 3 4 5 | 1 2 3 4 5] $n + 1 = n+1$ ex: 1 1' 1'' 1''' s 2 2' 2'' 2''' s   frame $n + 1 = n+1$

(13)

FDM

1) Frequency scale is shared by different signals.



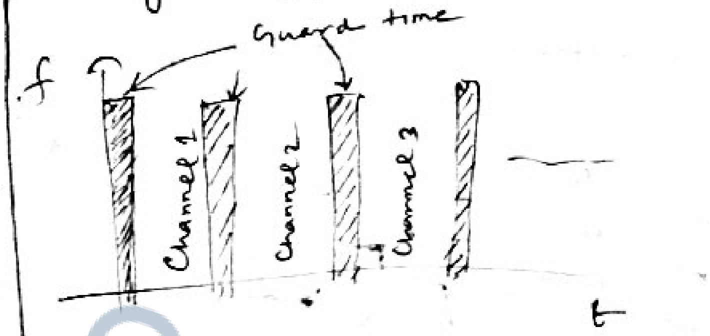
2) BW requirement:

AM/SSB → } nf_m
PAM →

AM/DSB } → $2nf_m$
AM

TDM

1) Time scale is shared by different signals.



2) BW requirement

AM/SSB } → nf_m.
PAM

AM/DSB } $2nf_m$
AM

So BW requirement in FDM, TDM are same.

But TDM is superior to FDM in following ways

(i) In FDM system, different carriers are to be generated for different channels. Also, as each channel occupies a different frequency band, different band pass filters are required.

On the other hand, in TDM system, all the channels require identical ckts, consisting of simple synchronous switches, gates & LPF. The circuitry needed in the TDM system is much simple than the one needed in the FDM system.

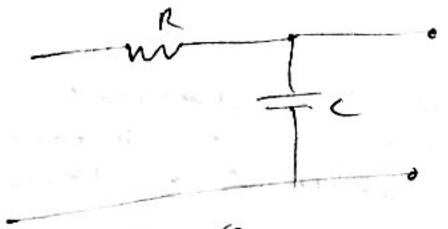
(ii) The non-linearities in the various amplifiers of an FDM system produce harmonic distortion and hence, they introduce interference within the channels.

But in TDM, the signal from different channels are allotted different time slots & they are not applied to the system simultaneously. Thus, the

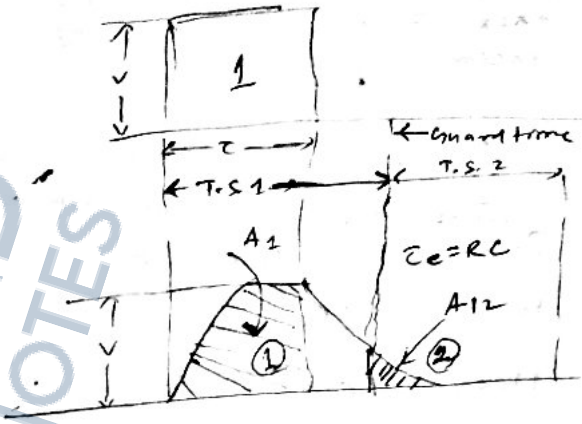
(14) TDM is relatively immune to interference within the channels as compared to FDM system.

Cross-Talk due to HF cut off channel :-

To band limit the communication channel, we represent the communication channel as RC low pass filter.



(a)



(b)

→ upper cut off freq $f_c = \frac{1}{2\pi RC}$.

→ After applying a pulse to the channel the pulse o/p ~~with~~ is shown in fig (b), It is due to HF limitation of the channel. The overlapping of time slot 1 into time slot 2. [cross talk is nothing but unwanted coupling of information from one channel to other.]

Cross-talk is due to

The cross talk factor 'K' is defined as ratio betn cross talk signal to the desired signal.

$$K = \frac{A_{12}}{A_2} = \frac{A_{12}}{A_1} \quad \text{—— (1)}$$

(∵ Avg area for A1 should be equal for A2)

In T.S. 1, the pulse is almost rectangular.

$$A_1 = \underline{V.\tau}.$$

$$A_{12} = V\tau_c e^{-\tau_g/\tau_c}(1 - e^{-\tau/\tau_c})$$

where
τ_g → guard time
τ_c → Time const.
τ → width of the pulse

to minimize cross talk,

$$\tau_c << \tau.$$

$$A_{12} \simeq V\tau_c e^{-\tau_g/\tau_c} \quad \text{—— (2)}$$

(15)

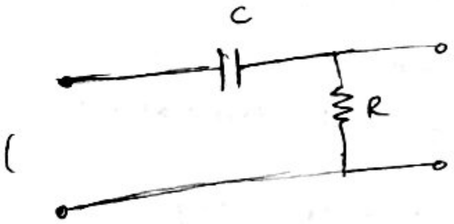
Putting eqn ② in eqn ①, we have

$$K = \frac{A_{12}}{A_1} = \frac{\cancel{V}\tau_c \, e^{-\tau_g/\tau_c}}{\cancel{V}\tau}$$

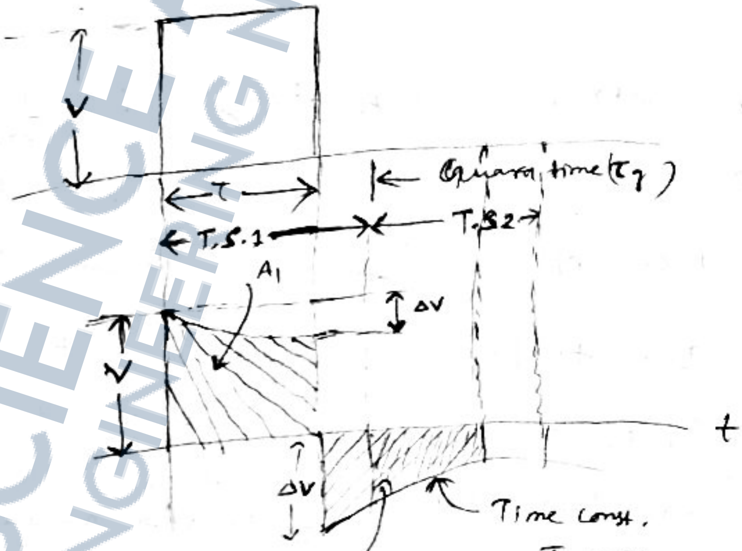
Cross talk factor = $\boxed{K = \frac{\tau_c}{\tau} e^{-\tau_g/\tau_c}}$

Crosstalk due to LF cutoff the Channel:-

Just the channel has upper cut off freq, it also has lower cutoff freq. Since we have Lower cut off freq, Hence the Channel can be represented in RC High pass filter.



(a)



(b)

High pass filter has lower cutoff freq,

$$f_1 = \frac{1}{2\pi RC}$$

Now, when a Pulse is applied to this Channel, the O/P of the Channel will be distorted due to LF limitation of the Channel. In this case, τ_c should be much greater than τ to reduce Cross-talk; ie $\underline{\tau_c >> \tau}$. (should discharge ~~slowly~~ slowly)

Now $\Delta V = V(1 - e^{-\tau/\tau_c}) \simeq V \cdot \frac{\tau}{\tau_c}$ (ΔV, derivative $e^{-\tau/\tau_c} \cdot \frac{-\tau}{\tau_c}$)

$(\because \tau_c >> \tau)$ $\left[\because e^x = 1 + x + \frac{x^2}{2!} + \frac{x^3}{3!} + \cdots \right]$

$$A_{12} \simeq \Delta v \cdot \tau = V \cdot \frac{\tau}{\tau_c} \cdot \tau = \frac{V\tau^2}{\tau_c}$$

But $A_2 \simeq A_1 \simeq V\tau$ [Rectangle V τ]

$\therefore$ Cross-talk factor,

$$K = \frac{A_{12}}{A_2} = \frac{A_{12}}{A_1} = \frac{V\tau^2}{\tau_c} \times \frac{1}{V\tau} = \frac{\tau}{\tau_c}$$

In this case, τ_c is very large, and hence the pulse may extends to many time slots. Hence this type of cross-talk extends to more than one channel.

Transmission of PAM :-

If PAM signal to be transmitted directly, say over a pair of wires, no further signal processing is necessary. If they are to be transmitted through the space using an antenna, they must be amplitude, frequency or phase modulated. The over-all system would be known as PAM-AM, PAM-FM, PAM-PM.

Bandwidth of PAM signals:-

BW requirement for transmitting 'n' signals, each band limited to 'f_m' Hz is $n f_m$ Hz.

S/N ratio of PAM system:-

The noise performance of PAM is identical to AM-SC signal.

The figure of merit $\gamma = \frac{S_o/N_o}{(S_i/N_i)} = \frac{o/p\ SNR}{i/p\ SNR}$

$$= \frac{S_o}{N_o} \times \frac{N_i}{S_i} = \left(\frac{S_o}{S_i}\right) \times \left(\frac{N_i}{N_o}\right)$$

For PAM signal, $S_o = S_i = \overline{f^2(t)}$, $\therefore \frac{S_o}{S_i} = 1$

17

For white noise,

$$N_o = N_i = n f_m \Rightarrow \frac{N_o}{N_i} = 1 .$$

$$\gamma = \left(\frac{S_o}{S_i}\right) \times \left(\frac{N_i}{N_o}\right) = 1 \times 1 = 1 .$$

$\therefore$ $\boxed{\text{figure of merit} = 1}$

Pulse Time Modulation :-

PTM systems are two types.

(i) PWM (Pulse width modulation)

(ii) PPM (Pulse position modulation)

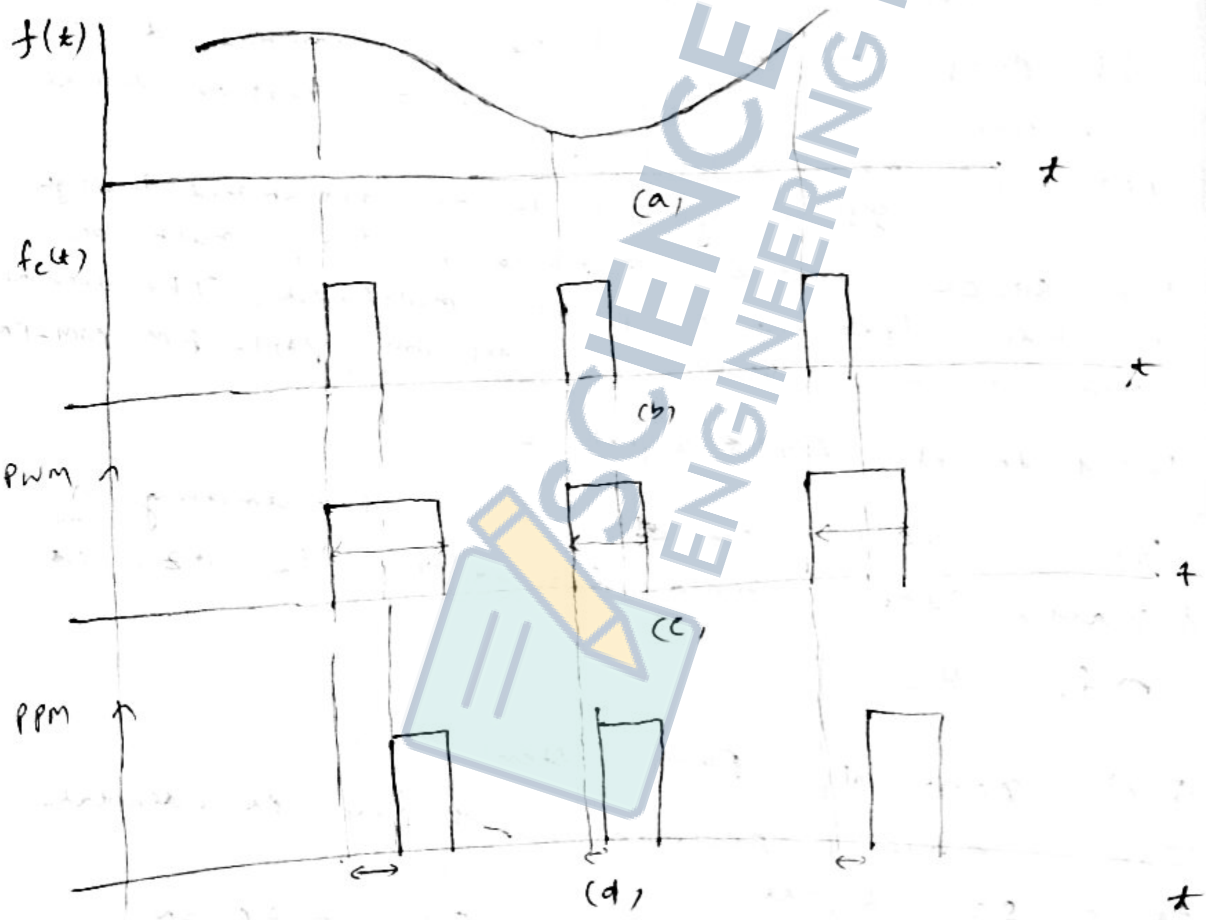


fig (a) Base band signal f(t)

(b) Carrier Pulse train fc(t)

(c) Pulse width Modulated signal

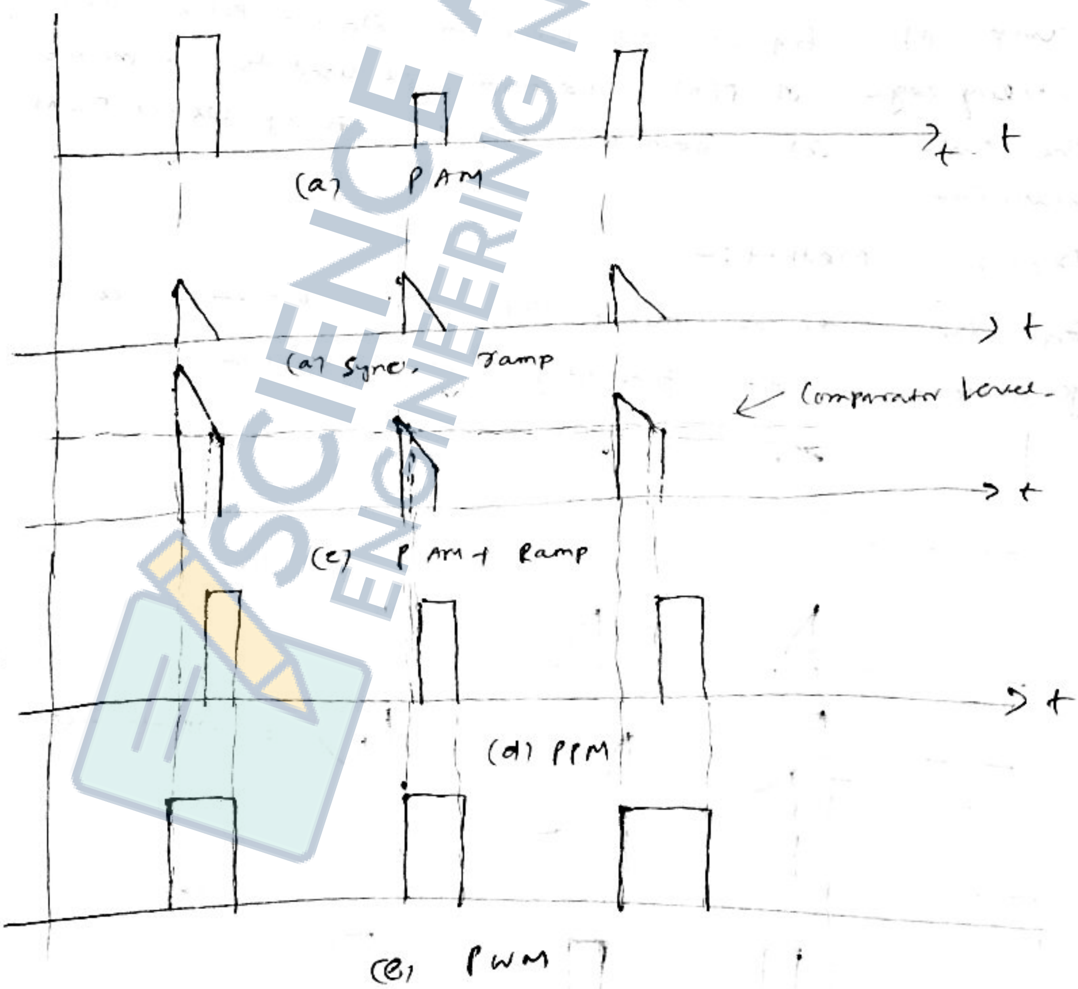
(d) Pulse Position modulated signal.

(18) Fig (c), is the PWM signal where the width of each pulse depends on the instantaneous value of the baseband signal at sampling instant.

Fig (d) is the PPM signal where the shift in the position of each pulse depends on the instantaneous value of the base band signal at sampling instant.

Generation of PTM signal :-

Indirect Method:-



Refer - fig:- 7.2.2. (Singh, Sapre) - 398 page

Steps:-

(I) First flat top PAM are generated shown in fig(a)

(II) The synchronized ramp waveform is generated in each pulse interval as shown in fig (b).

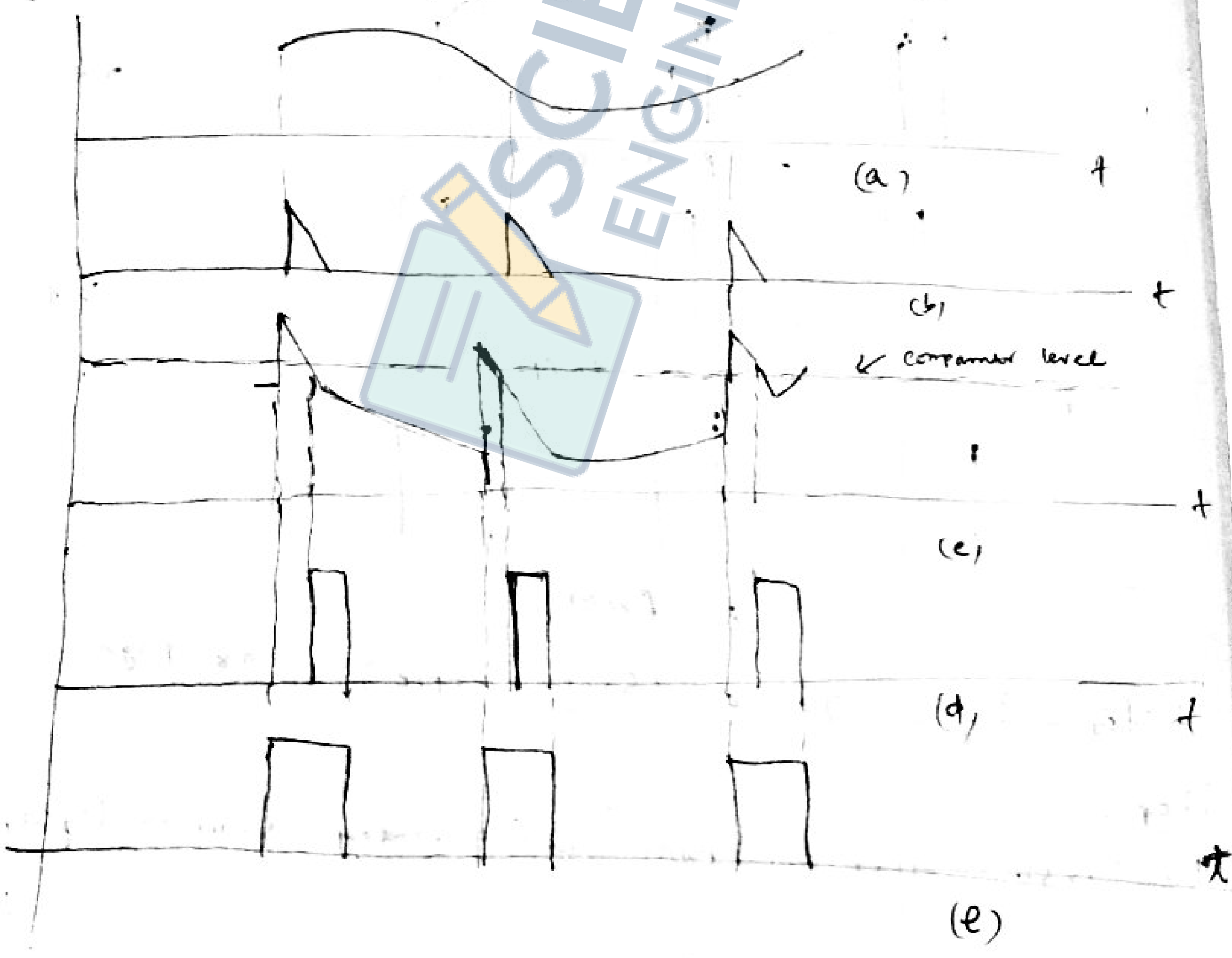
(III) These two signals are added as shown in fig (c). And the sum is applied to a comparator ckt whose reference level is shown by broken line as shown in fig (c).

(IV) The second crossing of the comparator reference level by the waveform of fig (c) is used to generate the pulse of const. amplitude and width as shown in fig (d).

The leading edge of the synch. ramp of fig (b) is used to start a pulse and trailing edge of PPM waveform is used to terminate the pulse as shown in fig (e) giving desired PWM waveform.

Direct Method :-

In the direct method the PTM waveform are generated without generating PAM waveform.



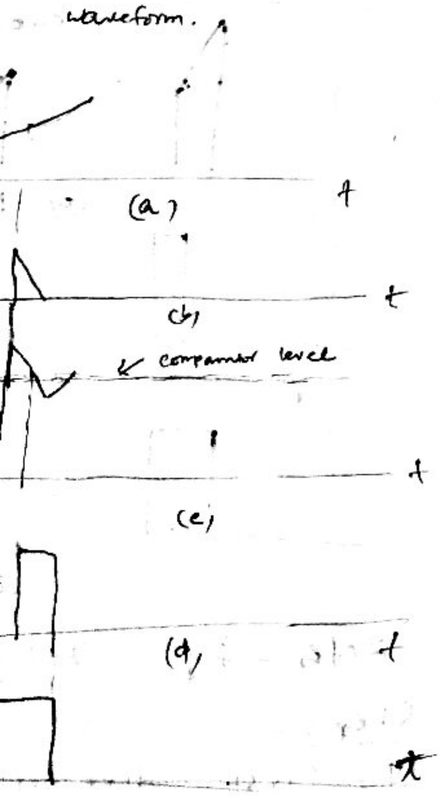
ramp waveform is
interval as shown

are added as shown
n is applied to
e reference level is
as shown in fig (c).

f the comparator
aveform of fig (c)
the pulse of const.
shown in fig (d).

edge of the synch.
start a pulse and
is used to terminate
giving desired PWM

TM waveform are
waveform.



(20) fig:-

(a) Base band signal f(t)

(b) Synchronized Ramp.

(c) f(t) + Synchronized Ramp.

(d) PPM signal

(e) PWM signal.

Step: (i) Here the baseband signal f(t) of fig(a) and ramp signal of fig(b) are added to give the waveform as shown in fig (c).

(ii) This is compared in a comparator whose reference level is shown in fig (c), horizontal & dotted line.

(iii) The PPM and PWM are obtained in the same way as explained in indirect method.

Note:- If we want to modulate the leading edge of pulses in the PWM waveform the ramp waveform shown in fig 2(a) is used.

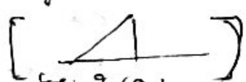


fig-2(a)

If we want to modulate both the edge of the pulse in PWM waveform, the ramp waveform shown in fig 2(b) is used

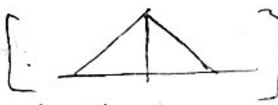
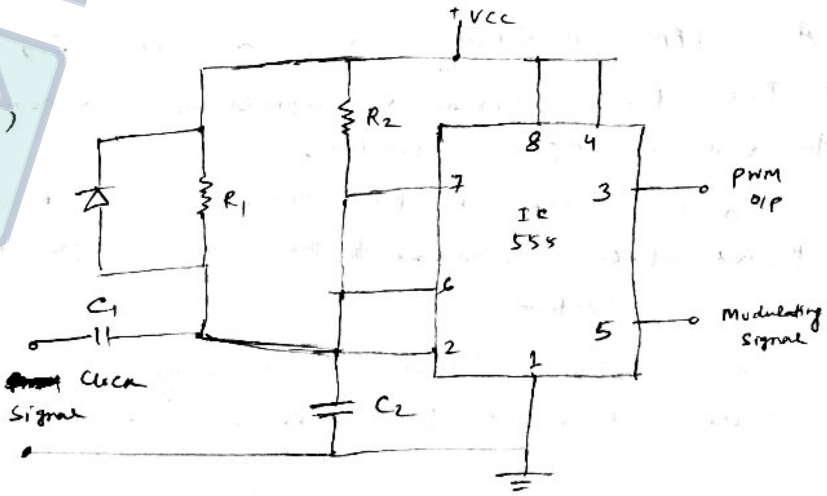


fig:- 2(b)

PWM Modulation ckt:-

fig:- 3(a)

PWM Modulator



→ For generation of PWM modulated signal we are using IC 555 as shown in fig 2(a).

→ The clock signal of desired frequency is applied as shown in fig. , from which the -ve trigger pulses are derived with the help of a diode and a R_1-C_1 combination. The R_1-C_1 combination is also called differentiator.

→ These -ve trigger pulses are applied to the pin number 2 of the 555 timer which is working in the monostable mode. They decide the starting time of the PWM pulses.

→ The end of the pulses depends on an R_2-C_2 combination and on the signal at pin number 5 to which the modulating signal is applied. (i/p pin)

→ Therefore, the width of the pulses depends on the value of the modulating signal.

→ The o/p at pin no. 3 is desired pulse width modulated signal.

PPM Modulator ckt:-

→ A PPM modulator is shown in fig 3(b).

→ The PWM signal is applied to pin number 2 through the diode and R_1-C_1 combination.

→ So the i/p to pin number 2 is the -ve trigger pulses which correspond to the trailing edge of the PWM waveform.

→ The 555 timer is working in a monostable mode & width of the pulse is constant.

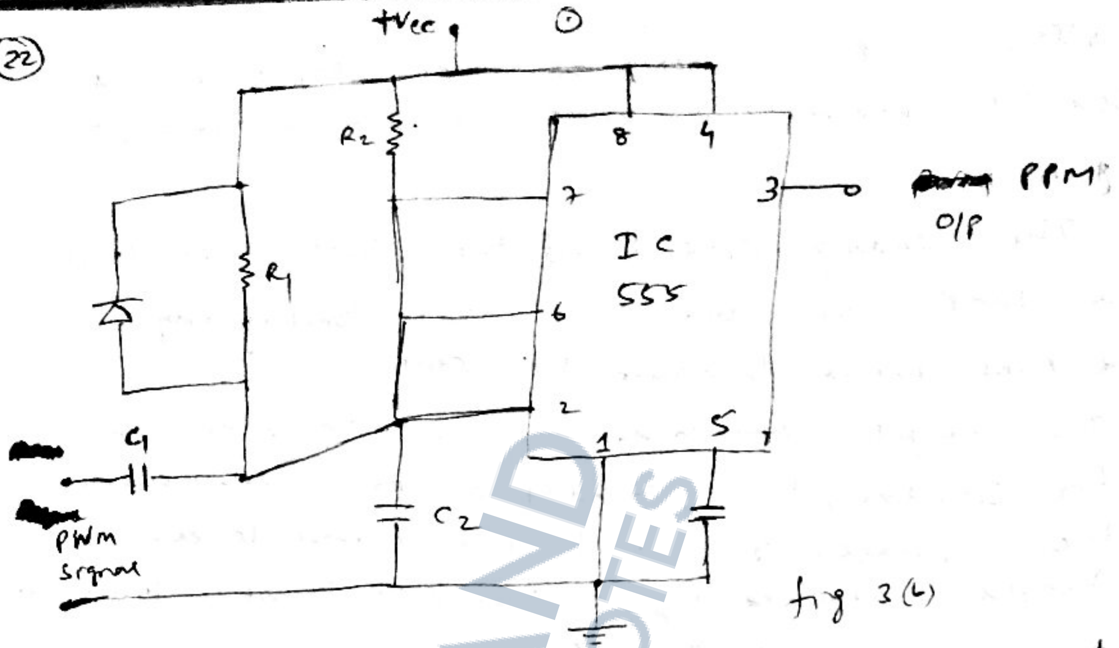


fig 3(b)

→ The −ve ~~pulses~~ trigger pulses decide the starting time of the O/P pulses and thus, the O/P at pin no. 3 is the desired PPM signal.

Demodulation of PTM signals :-

(a) Demodulation of PWM signal.

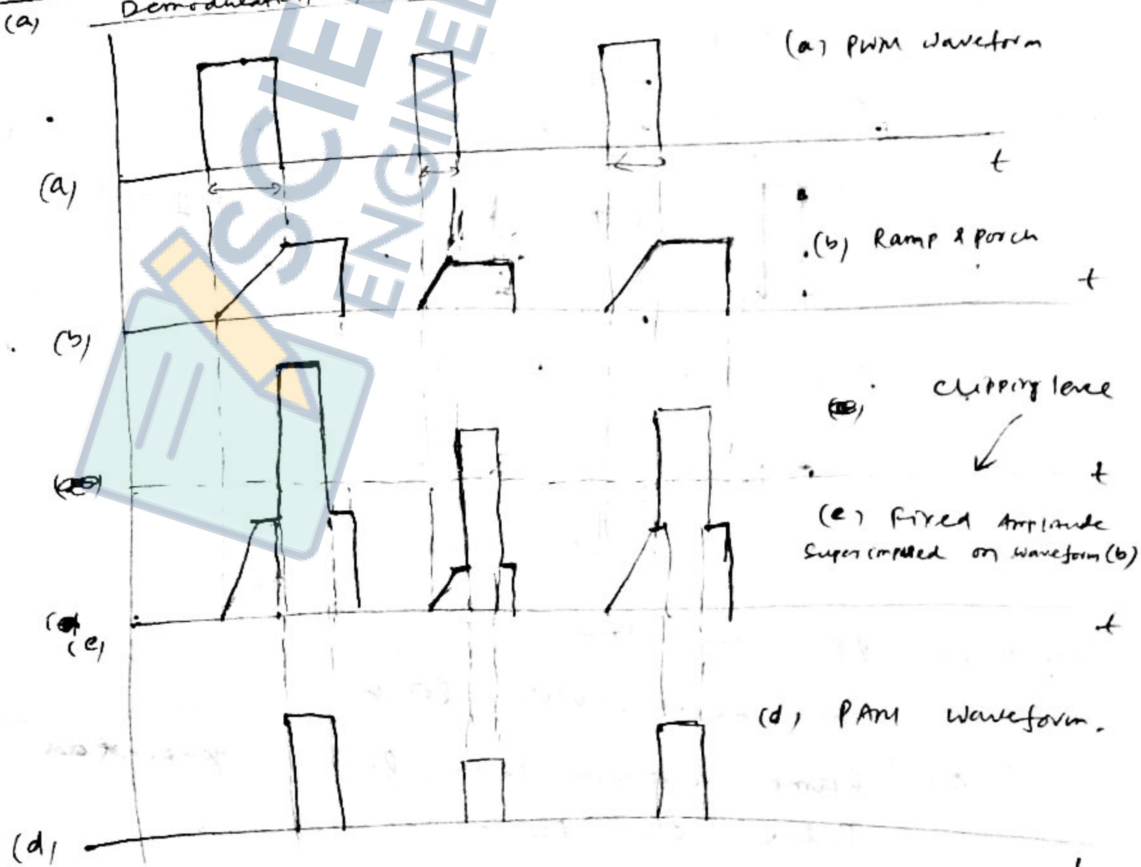


fig 4: a, b, c, d

(23) → The PWM waveform of fig 4-(a) is used to generate a ramp waveform as shown in fig 4-(b):

→ The leading edges of the PWM pulses start the ramp of same slope and trailing edges of the PWM pulses terminate the ramp.

→ The height attained by the ramp is sustained for some time, thus creating a porch, after which the voltage returns to its central level. Here height attained by the ramp is, therefore, proportional to the width of PWM pulses.

Demodulation of PPM Signal :-

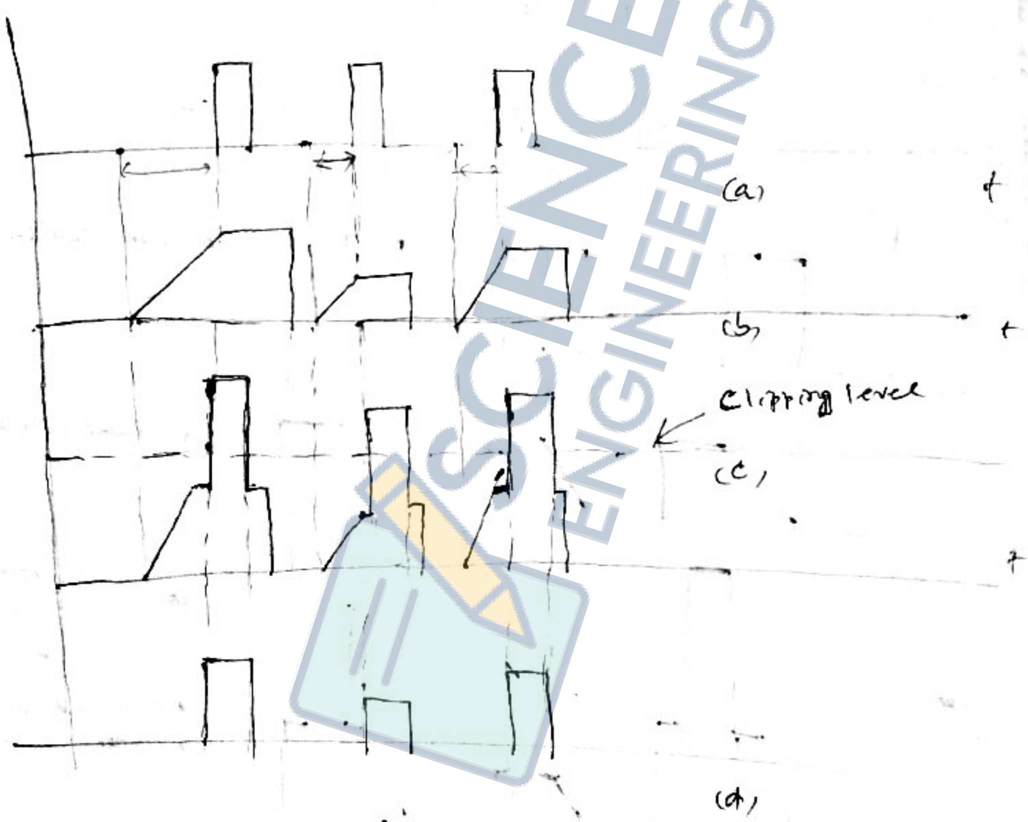


fig 5 (a) PPM waveform

(b) Ramp waveform with Porch

(c) Ramp waveform with locally generated Pulse on porch.

(d) PAM waveform.

(24) For P
fig 5(b) ramp is g
fig 5(a).
The begin
terminated b
pulse.
Thus
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for both PWM
generated pulses
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fig 4(d) &
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PWM dem

A PWM

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Capacitor

①

(24) For PPM demodulation, similar type of ramp is generated with PPM pulses shown in fig 5(b) fig 5(a). Here the ramp is initiated at the beginning of the time slot and it is terminated by the leading edge of the PPM pulse.

Thus the height attained by the ramp is proportional to the displacement of the ~~ed~~ leading edge of the PPM pulses from the beginning of the time slot. Here too, the height attained by the ramp is sustained for some time, thus creating a porch and then it is returned to the initial value.

The ~~remaing~~ remaining procedure for both PWM & PPM is same. A sequence of locally generated pulses of a fixed amplitude are added to the synchronized ramp on the porch as shown in 4(c) & 5(c). The lower portions in these waveforms are clipped by a clipping ckt, with the clipping level adjusted in such a way that it never crosses the ramp. The o/p of the clipper is a PAM waveform fig 4(d) & 5(d), from which the baseband signal can be recovered as discussed previously.

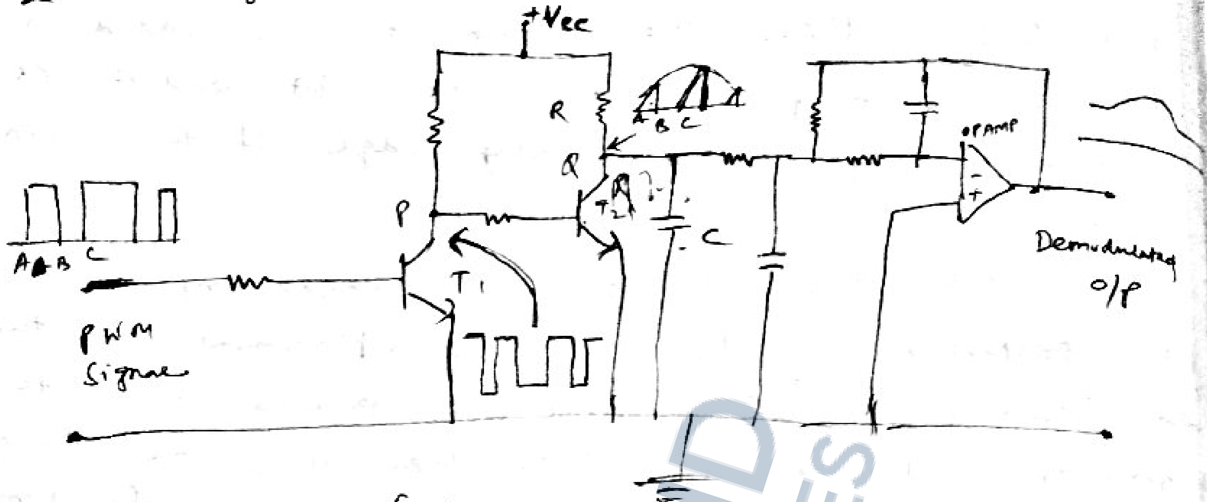
*

PWM demodulator ckt:-

A PWM demodulator ckt is shown in fig 7. The transistor T_1 works as an inverter. Hence, during the time interval A-B, when the PWM signal is high, the i/p to the transistor T_2 is low. Therefore, during this time interval, the transistor T_2 is cut off and the capacitor C gets charged through R-C combination.

During the time interval B-C when the

(25) PWM signal is low the i/p to the transistor T_2 is high, and it gets saturated.



fig(7)

→ The Capacitor C then discharges very rapidly through T_2. The Collector voltage of T_2 during the interval B–C is then low. Thus the waveform at the collector of T_2 is more or less a saw-tooth waveform whose envelope is the modulating signal.

→ When this is passed through a second order OP-AMP LPF, we get desired demodulated o/p.

A PPM demodulator ckt:-

→ A PPM demodulator ckt is shown in fig (8). This utilizes the fact that the gaps between the pulses of a PPM signal contain the information regarding the modulating signal. ~~During~~ ~~A-B~~

→ During the gap A-B between the pulses, the transistor is cut off, and the capacitor C gets charged through R–C combination.

→ During the pulse duration B–C, the capacitor discharges through the transistor, and the collector voltage becomes low.

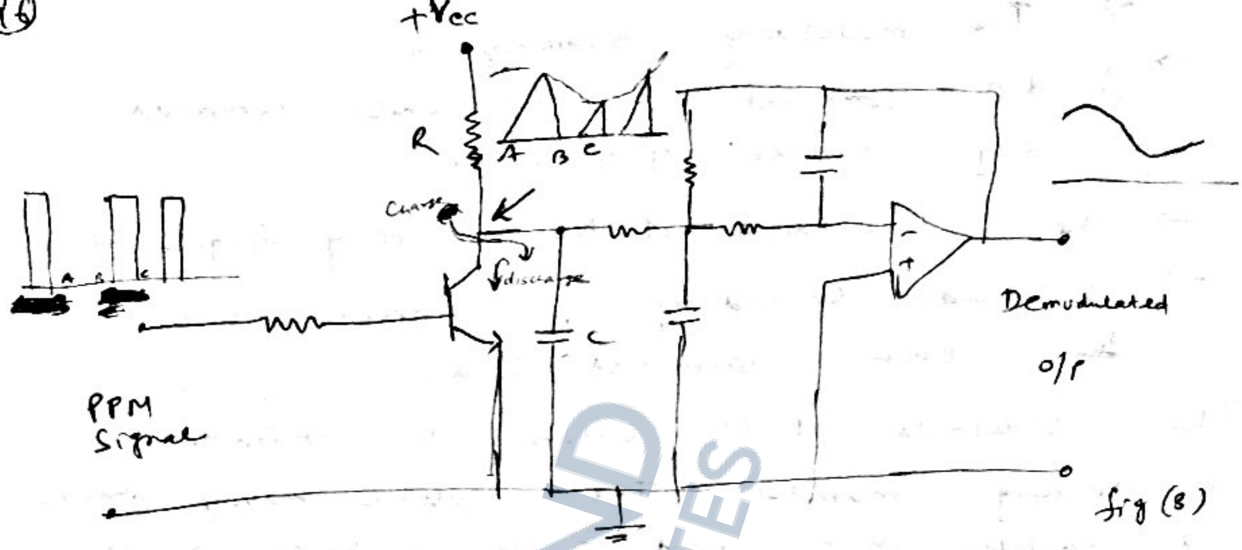


fig (8)

→ Thus, the waveform at the Collector is approximately a saw-tooth waveform whose envelope is the modulating signal. When this is passed through a second order OP-Amp LPF, we get the desired demodulated o/p.

Band width of PTM signal:-

From the spectrum of the PTM signals we can get the desired BW.

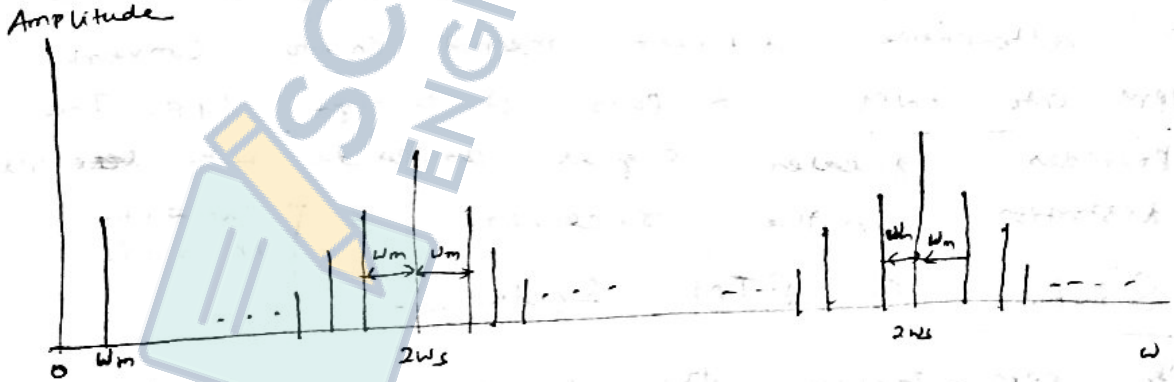


fig:- One sided spectrum of PWM signal.

Assume

ω_m = frequency of the modulating signal,

ω_s = sampling frequency.

Then the PWM spectrum has the following components.

P.T.O

→ The modulating frequency ω_m.

→ A d.c component at $\omega = 0$ which represents the avg. value of the pulses.

→ The harmonics of the sampling frequency ω_s.

→ Sidebands spaced by ω_m centered around ~~the~~ each harmonics of ω_s

The sidebands of ω_s extends to infinity with a decaying magnitude. The useful message band is available in a band $0 - \omega_m$ and hence a LPF is used to recover the message from PWM.

The spectrum of ~~a~~ naturally sampled PPM wave similar to PWM wave, only the difference that it contains a component proportional to the derivative of the modulating signal in place of modulating component itself.

Therefore the PPM detection can be achieved by an LPF followed by an integrator. An alternative detection method is to convert PPM into PWM & pass it through LPF. This provides greater signal amplitude with ~~less~~ less distortion in the receiver.

PWM = 5 f_m
PPM = 10 f_m

SNR of PTM Systems:-

→ In PPM signal, the sample values of the modulating signal $f(t)$ are transmitted in terms of the pulse position over a channel of bandwidth B HZ.

→ This finite bandwidth of the channel causes dispersion in the received PPM pulse. The resulting trapezoidal pulse is shown in fig 8 (a).

The rise time of the pulse is given as

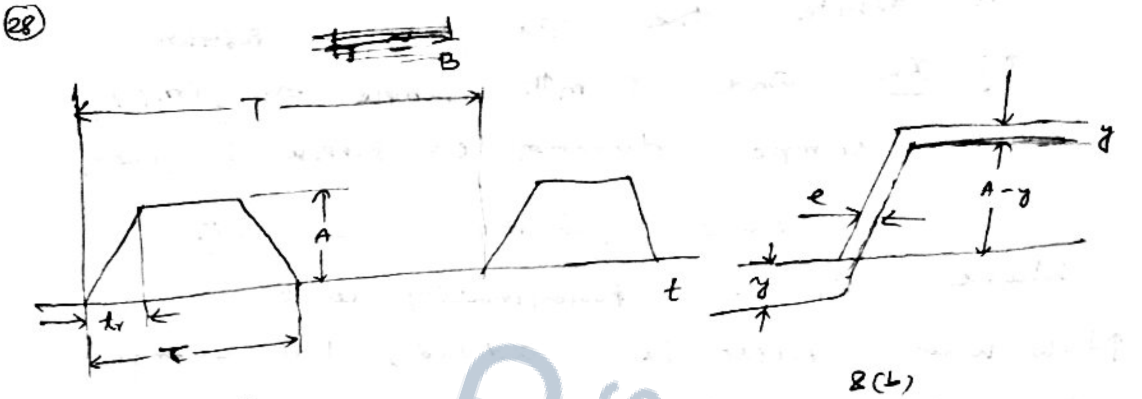


fig 8 (a)
(Trapezoidal pulse due to dispersion.)

8(b)
(Shift in the pulse of fig 8(a), due to noise.)

$$t_r = \frac{1}{B} \quad \text{——— (1)}$$

→ The position of the pulse is sensitive to any additive noise. If a noise signal 'y' is added at any instant, the position is shifted by 'e', shown in fig 8(b).

So we have

$$\frac{A}{y} = \frac{t_r}{e} \quad \text{——— (2)}$$

As y varies randomly, 'e' also varies randomly. The mean squared value of y and e are related as

$$\frac{\overline{e^2}}{\overline{y^2}} = \left(\frac{t_r}{A}\right)^2 \quad \text{——— (3)}$$

Therefore for an additive noise n(t) the mean squared value of e is given as

$$\overline{e^2} = \left(\frac{t_r}{A}\right)^2 \overline{n^2(t)} \quad \text{——— (4)}$$

Assume that change in position of the m^{th} pulse is proportional to the value of m^{th} sample of the modulating signal f(t). Suppose

29

We denote the the change in position by x_m and m^{th} sample by $f(mT)$. each sample duration is suppose T. Then,

$$x_m = K_1 f(mT) \quad \text{--- (5)}$$

Where K_1 is proportionality const.

Now when noise is added, the change in the position of the m^{th} pulse become $x_m + e_m$, where e_m is random and is defined by eqn (4). Lets denote this changed posn of the m^{th} pulse by x_m', then

$$x_m' = x_m + e_m = K_1 f(mT) + e_m \quad \text{--- (6)}$$

At the receiver, the pulse positions are converted back into samples, which are then passed through LPF for detection. It can be seen that o/p of LPF at the receiver is given by

$$x'(t) = K_1 f(t) + e(t) \quad \text{--- (7)}$$

Therefore, the useful message in this o/p is

$$S_o(t) = K_1 f(t) \quad \text{--- (8)}$$

and the noise signal in the o/p is

$$n_o(t) = e(t) \quad \text{--- (9)}$$

The mean square value provides the o/p signal & noise power as follows.

$$S_o = K_1^2 \overline{f^2(t)} \quad \text{and} \quad N_o = \overline{n_o^2(t)} = \overline{e^2(t)} \quad \text{--- (10)}$$

The mean square value e(t) is same as

on position
by $f(mT)$.
pose T. Then,
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———(9)
provides the
s follows.
$\overline{n_o^2(t)} = \overline{e^2(t)}$ ———(10)
(t) is same as

(20) mean square value of the samples.

$$\overline{e^2(t)} = \overline{e_m^2} = \left(\frac{t_r}{A}\right)^2 \cdot \overline{n^2(t)}$$

(from eqn 4)

and $N_o = \left(\frac{t_r}{A}\right)^2 \cdot \overline{n^2(t)}$ (from eqn 10) & eqn (4)

Therefore,

$$\frac{S_o}{N_o} = \frac{K_1^2\,\overline{f^2(t)}}{\left(\frac{t_r}{A}\right)^2 \cdot \overline{n^2(t)}}$$

$$= K_1^2 \cdot \left(\frac{1}{t_r}\right)^2 \cdot \overline{f^2(t)} \cdot \frac{A^2}{\overline{n^2(t)}}$$

$$\frac{S_o}{N_o} = K_1^2 \cdot B^2\,\overline{f^2(t)} \cdot \frac{A^2}{\overline{n^2(t)}}$$

(where $t_r = \frac{1}{B}$, from eqn 1)

Let the duty ratio 'd' of the pulse be given by

$$d = \frac{\tau}{T}$$

The power contained in the PPM wave of amplitude 'A' is given by

$$S_i = d A^2$$

The i/p noise power is given by,

$$N_i = \overline{n^2(t)}$$

Therefore, ~~noise~~ figure of merit,

$$\gamma = \frac{S_o/N_o}{S_i/N_i} = \frac{K_1^2 \cdot B^2\,\overline{f^2(t)} \cdot \cancel{\frac{A^2}{\overline{n^2(t)}}}}{\cancel{\frac{d A^2}{\overline{n^2(t)}}}}$$

$$\boxed{\gamma = \frac{K_1^2}{d}\,\overline{f^2(t)} \cdot B^2}$$

(31)

Thus, $\frac{S}{N}$ ratio improvement $\propto$ square of channel BW

$$\boxed{\gamma \propto B^2}$$

→ For PWM also $\boxed{\gamma \propto B^2}$

→ Noise performance of PWM is inferior to PPM.

→ PPM system preserves all signal information at the terminating instant of the pulses and yet it avoid a considerable loss of power which PWM system expands during pulse.

→ So PWM is less efficient than PPM, with regard to transmitter power utilization.

Ex:- 1) For a PAM transmission of voice signal having maxm freq. equal to $f_m = 3\,kHz$, Calculate Transmission BW. It is given that sampling freq $f_s = 8\,kHz$ and pulse duration $\tau = 0.1\,T_s$.

Ans:- $T_s = \frac{1}{f_s} = \frac{1}{8 \times 10^3} = 0.125\,ms.$

$\tau = 0.1 \times T_s = 0.1 \times .125 \times 10^{-3} = 12.5\,\mu s$

$BW \geq \frac{1}{2\tau}$ (For PAM transmission BW)

$BW \geq \frac{1}{2 \times 12.5 \times 10^{-6}}$

$BW \geq \frac{10^2 \times 10^4}{25}$

$$\boxed{BW \geq 40\,kHz}$$

2/ Compare PAM, PWM, PPM.

PAM	PWM	PPM
1) Amplitude of pulse is proportional to the amplitude of the modulating signal.	1) Width of the pulse is proportional to the amplitude of the modulating signal.	1) The relative position of the pulse is proportional to the amplitude of modulating signal
2/ BW of transmission channel depends on the width of the pulse.	2/ BW of transmission channel depends on rise time of the pulse.	2/ BW of transmission channel depends on rise time of pulse.
3) The instantaneous power of the transmitter varies.	3) Instantaneous power of the transmitter varies.	3) Instantaneous power of the transmitter remains const.
4) Noise interference high.	4) Noise interference is medium.	4) Noise interference is minimum.
5) Similar to AM	5) Similar to FM	5) Similar to PM.
6) SNR is low	6) SNR is medium.	6) SNR is high.

Pulse Code Modulation (PCM)

Introduction :- → Pulse modulation systems are not completely digital.

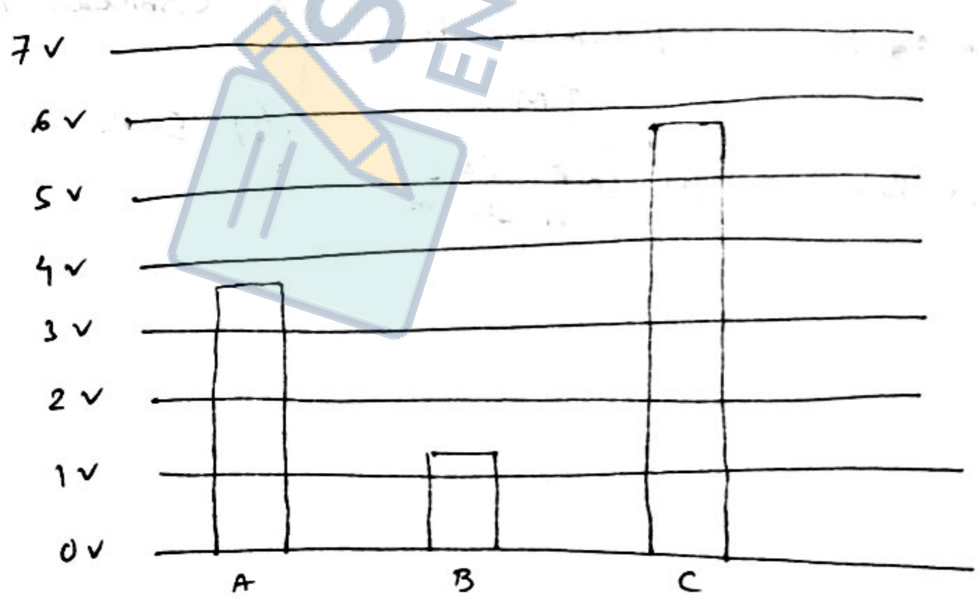
→ e.g PAM signal is a discrete time signal where signal on time axis is discrete and on amplitude axis it is continuous.

→ To ~~achieve~~ get a digital signal from this signal, the signal on amplitude axis should also be made discrete.

→ The process of converting a continuous amplitude & discrete time signal into discrete amplitude & discrete time signal is called quantization.

→ Example :- Lets consider a PAM signal whose amplitude varies from -0.5 V to 7.5 V.

→ This range is divided into 8 (2^3) levels known as quantization levels.



fig(1)

→ The pulses having values from -0.5 to 0.5 are approximated (quantized) to a value 0V. Then pulse having values from 0.5 V to 1.5 V are

~~app~~ approximated to a value 1V & so on.

→ It is assumed that, as per probability theory exact values as 0.5, 1.5V etc. will never occure.

→ Thus, any pulse can be approximated to one ~~case~~ of the values of quantization levels 0V, 1V etc.

Voltage Range (in volt)	Quantization Level	Binary code
−0.5 to 0.5	0	000
0.5 to 1.5	1	001
1.5 to 2.5	2	010
2.5 to 3.5	3	011
3.5 to 4.5	4	100
4.5 to 5.5	5	101
5.5 to 6.5	6	110
6.5 to 7.5	7	111

→ In fig (1), pulses A, B, C have amplitude 3.8V, 1.2V & 5.7V respectively, they will be approximated to 4V, 1V, 6V respectively.

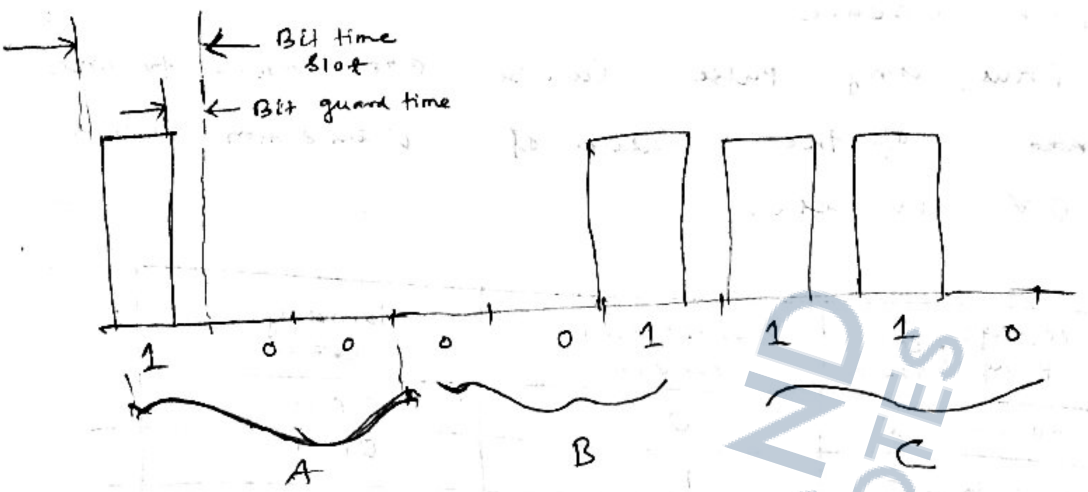
→ These are known as quantized pulses.

These quantised pulses are then coded. So three quantized pulses having values 4V, 1V, 6V will be encoded as 100, 001 & 110 respectively.

→ The presence of pulse may be represented by 1 and its absence by a 0. The transmitted signal for pulses A, B, C will be

(35)

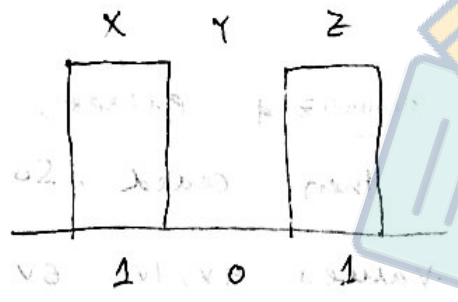
as shown in fig 2. Note that there is ~~guard~~ a time slot allotted to each bit, a portion of which is guard time.



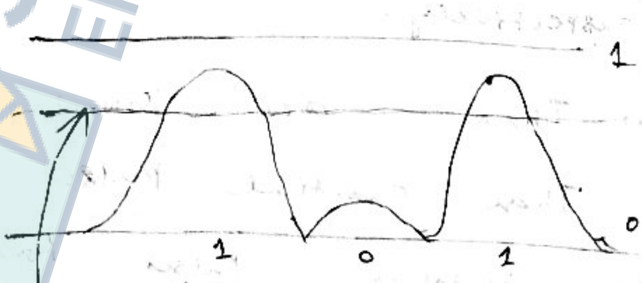
100 001 110

→ The major advantage of the PCM system is that the information does not lie in any property of the pulse, but it lies in the presence or absence of the pulse. Thus, even if noise distorts the pulse it makes no difference as long as the decision regarding the presence or absence of the pulse is correct.

→

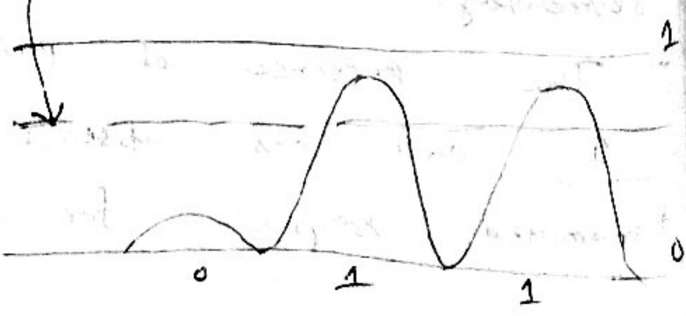


(a) (Transmitted pulses)



(b) Received pulses

Threshold level



(c) Received pulses

(36)

fig (b)
pulses are
no error
(Note:- W
crosses
~~error~~
Transmitted
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the
reducing
→ Th
extreme
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needed
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needed
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effect

fig (b) shows that although the received pulses are distorted due to noise, there is no error of decision.

(Note:- when the voltage in a time slot crosses the threshold it is treated as 1 ~~else~~ else it is treated as 0)

Transmitted → 101, Received → 101 (No error)

In extreme noise, consider fig (c), the received pulse is 011. There is error in first & 2nd bit.

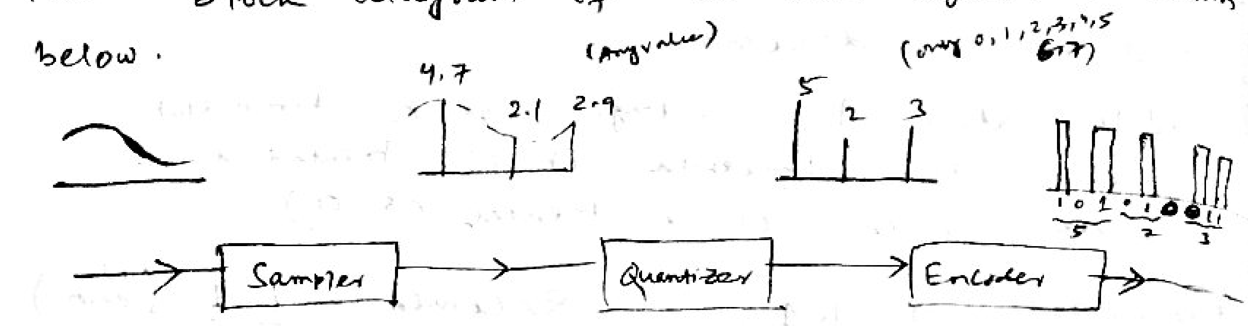
There is possibility of error due to quantization, as there is no way to know whether a 4 V signal is a result of 3.8 V signal or 4.3 V signal. However the quantization error can be minimized by reducing the step size.

→ The repeaters in PCM system are extremely simple as compared to those used in analog communication system. Amplifiers are needed in analog repeaters which not only amplify noise along with signal, but also generate noise internally, thus degrading the quality of communication.

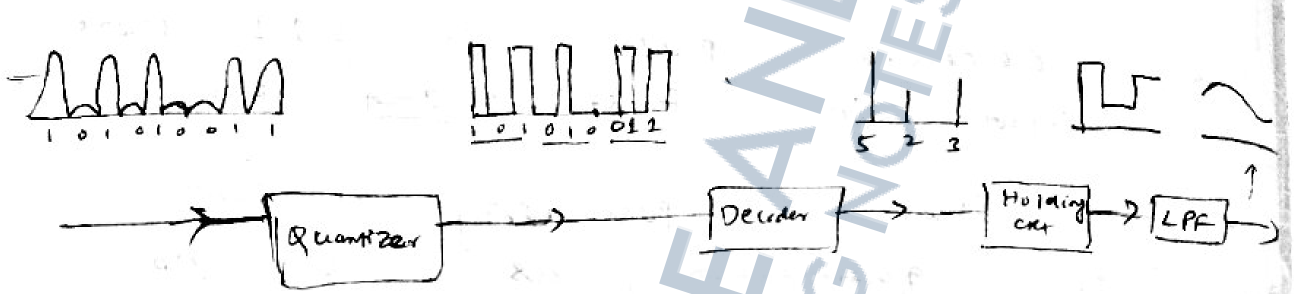
On the other hand, the repeaters needed in PCM system require only regenerators which generate pulses in time slots according to the presence of 0 or 1, thus eliminating the effect of noise till that point.

PCM System

The block diagram of a PCM system is shown below.



(a) Transmitter



(b) Receiver

→ Fig (a) shows a PCM transmitter. The baseband signal is sampled at Nyquist rate by the sampler. The sampled pulses are then quantized in the quantizer.

The encoder (an A/D converter) encodes these quantized pulse into bits which are transmitted over the channel.

→ Fig (b) shows a PCM receiver. The first block is again a quantizer. But this quantizer is different from the transmitter quantizer because it has to take a decision about the presence or (1) absence (0) of a pulse.

The O/P of the quantizer goes to the decoder which is D/A converter, performs the inverse operation of the encoder. The decoder O/P is a sequence of quantized pulses.

(38) The original baseband signal is reconstructed in the holding Ckt & LPF.

Inter Symbol Interference:-

→ [As the PCM Channels are band limited, the received waveforms are distorted and they extend to the next time slot, resulting in error in the determination of received bits.

→ The situation is analogous to cross-talk in PAM system. In PAM, the adjacent time slots are different channels and hence the term cross-talk is appropriate.

→ However, in PCM the adjacent time slots are generally symbols in code representation of a single quantized sample; hence the term interSymbol interference is used for PCM.]

→ Defn :- (In a communication system, when the data is being transmitted in the form of pulses (ie bits), the o/p produced at the receiver due to other bits or symbols interferes with the o/p produced by the desired bit. This is known as Inter Symbol Interference (ISI). The intersymbol interference will introduce error in the detected signal.)

→ In a 8 bit ~~system~~ PCM signal, for first 7 bits, the adjacent bit corresponds to code representation of a single quantized sample, where as for eighth bit, the adjacent bit corresponds to next channel.

①

Eye Patterns:-

→ A CRO can be used to give an indication
of the performance of a PCM system.
→ In this method, the received bit stream
is applied to the vertical deflection plates
and the time base frequency is made equal
to the bit rate so that a sweep lasts
one time slot duration.
→ When the received bit stream is ideal,
as shown in fig (a), the CRO pattern will
be like the one shown in fig (b).
→ If the bit stream is distorted as shown
in fig (c), the CRO pattern will be shown in
fig (d). The pattern of fig (d) is very
much similar to human eye, the central
portion of the pattern being the opening of the
eye.

→ If the signal is further distorted,
as shown in fig (e), the CRO pattern will
be as shown in fig (f).

→ In this case, the eye is further closed.
Thus an observation of the eye
pattern gives an idea about the
distortion of the system. The more is
the opening of the eye, the less is
the distortion and vice versa.

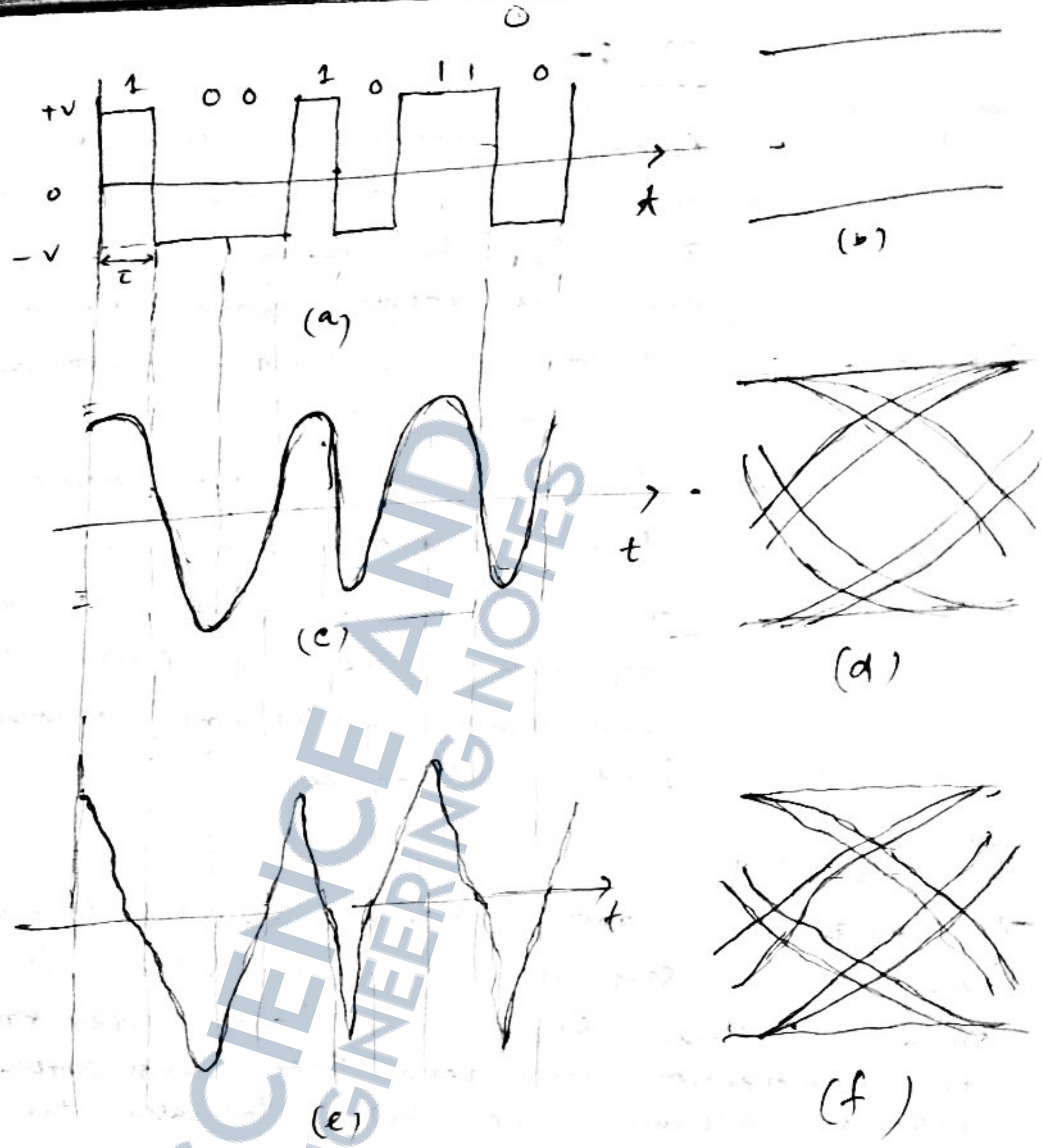


fig:-

Eye patterns:

(a) Received Bit stream (ideal)

(b) CRO pattern for (a)

(c) Received bit stream (less distortion)

(d) CRO pattern for C

(e) Received Bit stream (more distortion)

(f) CRO pattern for (e)

(41)

Equalization :- (Assignment II)

→ The intersymbol interference causes distortion. These distortions can be reduced by designing a proper equalizer. If the frequency response of the channel $H_c(\omega)$ is completely known, then an equalizer is designed whose frequency response $H_e(\omega)$ is the inverse of $H_c(\omega)$.

→ Equalizing filters are inserted between the receiving filter and A/D converter.

→ We have to adjust the equalizer filter manually by observing the eye pattern. In an adaptive equalizer, this is done automatically by using feedback technique.

Companding:-

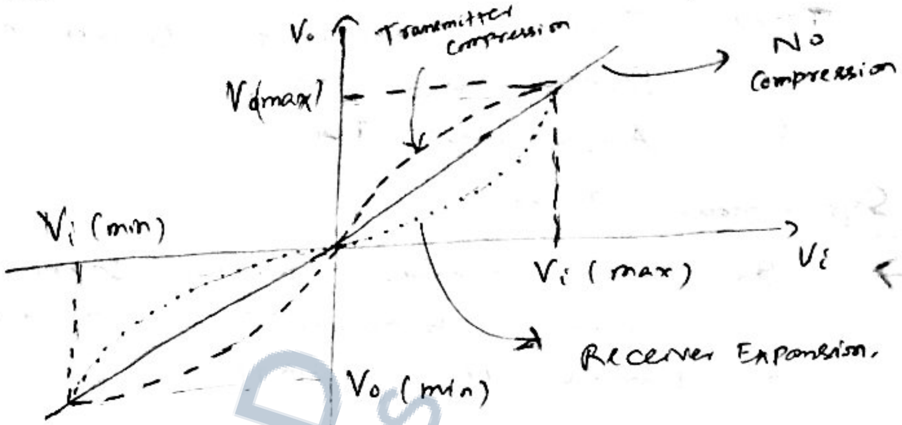
→ Quantization error depends upon the step size. [When the steps are uniform in size, the small amplitude signal will have poorer signal to quantization noise ratio than large amplitude signals, because in both the cases the denominator (quantization noise) is the same; whereas, the numerator ~~in both the cases~~ is small for small amplitude and large for large amplitude.]

→ [Since we have to use fixed number of quantization levels, & we have to adjust the step size so that SNR remains const.]

→ So, the step size should be small for small amplitude signals & large for large amplitude signals.

42

→ This is explained in fig given below.



→ [The o/p is enhanced more at low amplitudes than at high amplitudes (dashed curve).]

→ This o/p is then applied to quantizer.

[Thus, the low amplitude signal will carry more quantization levels than the ~~the~~ undistorted signal (Solid line).]

→ [A signal transmitted through a non-linear n/w with the characteristics shown by the dashed curve will have its extremities compressed. Hence, such a n/w is known as Compressor.]

→ At the receiver side, an inverse operation is to be performed to recover the original signal. This is achieved by an expander connected betn the decoder and holding ckt, whose characteristic is shown by the dotted curve.

→ The combination of compressor & expander is known as Compander, which performs the companding operation.

→ Time division multiplexing of PCM Signals:-

Since PCM is a digital system, various signals

(43) Can share the time scale, giving rise to TDM. The multiplexing concepts are 2 types.

(i) Synchronous TDM

(ii) Asynchronous TDM

Synchronous TDM:-

→ In this method each sample is coded into several bits. The multiplexing can be possible in 2 ways.

(a) Bits are taken, one by one from each channel sample code. After the first bits from all channels are taken, the commutator takes the second bits from all channels samples, and so on. This is 'bit interleaving'.

(b) All code bits of the first channel samples are taken followed by all codes bits of the second channel samples & so on. In this method, the desired commutator speed is less than that required in the first method. This is 'word interleaving'.

→ At the end synchronizing bits are added to each frame of synchronization between commutator and decommutator.

→ The signal that is to be time division multiplexed is band limited to the same frequency which results in the same sampling frequency for all channels and hence the ~~same~~ name synchronous time division multiplexing.

Asynchronous TDM (Pulse Stuffing)

→ When signals to be time division multiplexed are bandlimited to different frequencies, their sampling frequencies are also different, they can't be multiplexed by synchronous TDM.

(44) → Such asynchronously sampled signals are multiplexed by a technique called <u>pulse stuffing</u>.

<u>Procedure</u> :- <u>Different</u> signals are sampled at different frequencies. The samples are stored on different storage devices. The recording rate of each storage device is different due to the different sampling frequencies.

While transmitting these signals each storage device is played back at different speeds in such a way that the O/P sample rate of each device is the same. Hence, these signals can now be synchronously time division multiplexed and then transmitted. At the receiver, this process is reversed to recover each signal.

~~The pulses~~ In this method, the pulses ~~be~~ stuffed into spaces provided for empty time slots, this is called <u>pulse stuffing</u>.

[<u>Note</u> :- why pulse stuffing reqd.?

A : Let 2 signals band limited to 4 KHz & 5 KHz respectively. Corresponding sampling frequencies will be 8 KHz & 10 KHz, respectively and thus the word duration will be 125 μs & 100 μs respectively.

Lets assume the above 2 signals are recorded on 2 separate devices for a duration of 1 sec. The first storage device will then have 8000 words of first signal recorded on it, where as second storage device will have 10,000 words of second signal recorded on it. To time division multiplex these signals

(45) ①

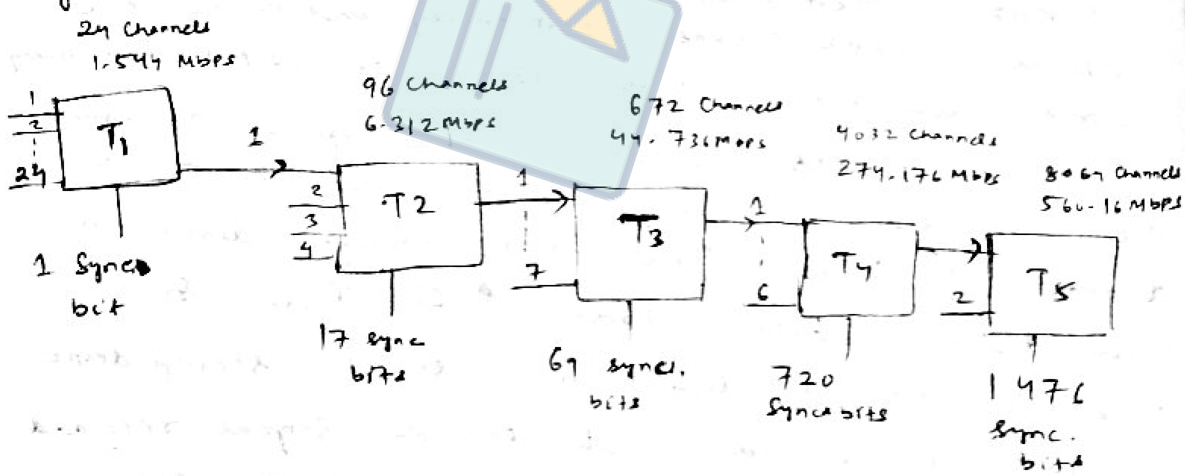
Each storage unit is played back at the same ~~rate~~ word rate. (Note that the playback speed of 2nd device should be slower than the first device by 20% to have the same word rate).

The first 8000 words can now be multiplexed without any trouble. But, during the 2nd multiplexing the last 2000 words of the first signal, there is no contribution from the first signal. Hence, these 2000 (time slots of the ~~are followed by~~ first signal are filled with a sequence of digits (highly improbable (0, 1) to indicate that actually no message is being transmitted during these time slots.

These pulses (the above method requires) be stuffed into spaces provided for empty time slots, it is known as pulse stuffing.]

Transmission Hierarchy (T carrier Systems)

Multiplexed PCM channels are transmitted using various T carrier systems such as T1 carrier system, T2 carrier system ---- etc. (Shown below)



(46)

T1

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Frame

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or 193x
78...

T1 Carrier System:-

In this system, 24 voice freq. signals, each sampled at 8000 samples per second, encoded into an 8-bit word and transmitted along with one synchronizing bit per frame using BRZ - AMI code. T1 carrier system uses word interleaving:

$$\text{Frame duration} = \frac{1}{8000}\ \text{sec} = 125\ \mu\text{sec.}$$

$$\text{No. of bits per frame} = (24 \times 8) + 1 = 193\ \text{bits}$$

$$\text{Hence, the transmission rate} = \frac{193}{125 \times 10^{-6}} = 1.544\ \text{Mbps.}$$

[In addition to voice signals, special supervisory signals are also to be sent to the receiving end. These are needed to transmit off-hook / on-hook signals, dial pulses and telephone ... For this purpose, the least significant bit of each voice channel of every sixth frame is deleted and a signalling bit is inserted in its place. The rate of transmission of these signalling bits is thus are sixth of the sampling rate, i.e $\frac{8000}{6} = 1333\ \text{bps} = 1.33\ \text{kbps}$]

T2 Carrier System:-

→ 96 voice freq. signals are multiplexed.

→ Achieved by multiplexing O/P of 4 T_1 carrier.

→ Additional synchronization bits used 17.

$$\text{Transmission rate} = 4 \times 1.544\ \text{Mbps} + (17 \times 8)\ \text{kbps}$$

$$= 6.312\ \text{Mbps.}$$

17 synch bits → 1 rev
1 rev → 17 bits
1 sec = 8000 rev → 8000 × 17
= 17 × 8 kbps

or $193 \times 4 + 17 = 789$ bits/frame.

$$789\ \frac{\text{bits}}{\text{frame}} \times 8000\ \frac{\text{frame}}{\text{sec}} = 6.312\ \text{Mb/s}$$

(47)

T_3 Carrier System:-

→ 672 voice freq channels are multiplexed.
→ Achieved by multiplexing o/p of 7, T2 Carriers.
→ 69 synchronization bits are used in T3 Carrier system.

→ Transmission rate

$$= (7 \times 6.312)\,\text{Mbps} + (69 \times 8)\,\text{Kbps}$$

$$= 44.736\ \text{Mbps}.$$

T4 Carrier System:-

→ In this system 4032 voice freq signals are multiplexed. Achieved by multiplexing six T3 carriers. 720 synchronization bits are used.

$$\text{Transmission rate} = (6 \times 44.736)\,\text{Mbps} + (720 \times 8)\,\text{Kbps}$$

$$= 274.176\ \text{MBPS}.$$

T5 Carrier System:-

→ In this system, 8064 voice frequency signals are multiplexed. This achieved by multiplexing o/p of 2 T4 Carriers. 1476 synchronizing bits are used in T5 Carrier system.

$$\text{Transmission rate} = (2 \times 274.176)\,\text{Mbps} + (1476 \times 8)\,\text{Kbps}$$

$$= 560.16\ \text{Mbps}.$$

Line Code :-

The digital data (0's and 1's) are transmitted

...nels are multiplexed.
o/p of 7, T2 Carriers.
are used in T3 Carrier

(69 x 8) kbps

s.

voice freq signals
y multiplexing
nchronization bits

736) Mbps + (720 x 8) kbps

Mbps.

voice frequency
els achieved by
T4 Carriers.
used in T5

176) Mbps + (1476 x 8) kbps

Mbps.

's.) are transmitted

(48) Over the line by means of 'Line code' (also known as 'Data Transmission Codes' or 'Modulation Codes'). They give electrical representation of symbol 0 and 1.

Types of Line codes

1. UNRZ (unipolar Non-Return to Zero) Code:

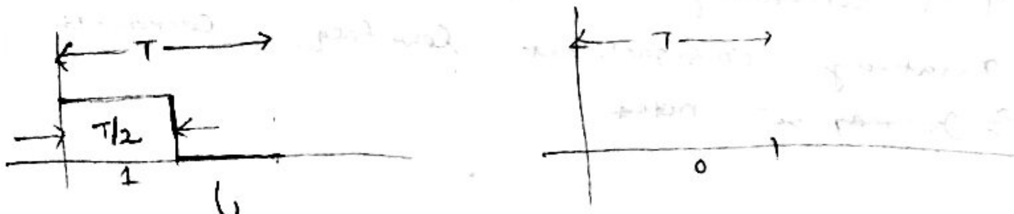
In this Code, a '1' is represented by a +ve pulse and '0' is represented by no pulse. This is also known as 'On-Off' code.

2. BNRZ (Bipolar Non-Return to Zero) Code:-

In this code, a '1' is represented by a +ve pulse and '0' is represented by a -ve pulse.

3. URZ (Unipolar Return to Zero) Code:

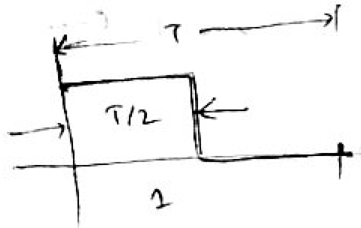
In this Code, a '1' is represented by a +ve pulse of half symbol width & a '0' is represented by no pulse.

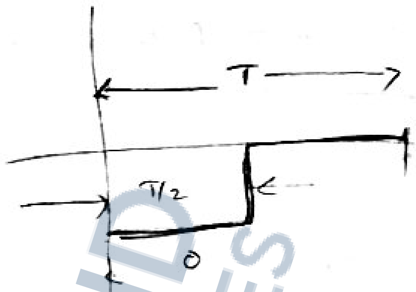


Returned to Zero

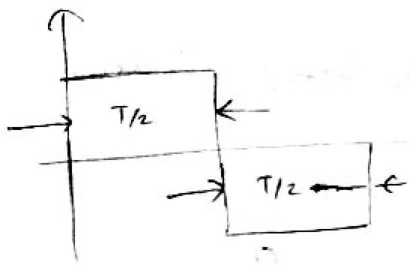
(49) 4. BRZ (Bipolar Return to Zero) Code:-

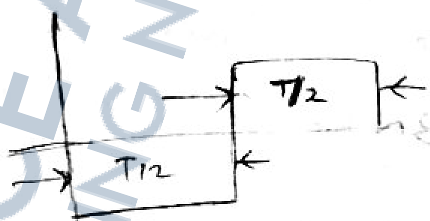
In this code, a '1' is represented by a +ve pulse of half-symbol-width, and '0' is represented by a -ve pulse of half symbol width.





5. Split-Phase Code (Manchester Code):-





In this code, a '1' is represented by a +ve half-symbol width pulse followed by a -ve half symbol width pulse.

'0' is represented by a -ve half symbol width pulse followed by a +ve half-symbol width pulse.

→ This code has a zero dc component because for both symbols, the dc component is zero.

→ Moreover, the maxm half-width duration (+ve as well as -ve pulse) in this code is T (corresponding to 01 or 10) and hence, it has relatively insignificant low-freq. components.

~~→ It may be noted~~

...to Zero) Code:-

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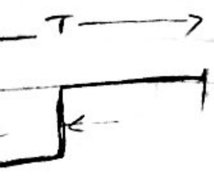
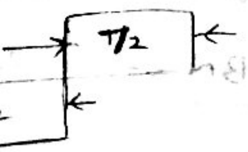


0

...Code):-



0

...sented by a +ve

... -ve half symbol

... -ve half symbol

...half-symbol width

...component because

...component is zero.

...width duration

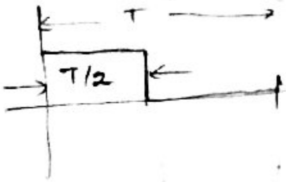
...n this code is

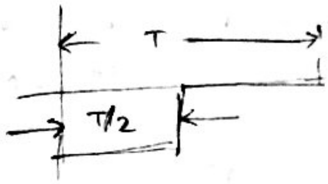
...d hence, it has

...components.

(50)

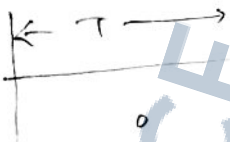
6. Differential Code or BRZ-AMI (Bipolar Return to Zero - Alternate Mark Inversion) Code).





(Symbols alternately occure)

'1'



0

In this Code, a '1' is represented alternately by a +ve pulse of half-width and a -ve pulse of half width whereas '0' is represented by no pulse.

→ The, d.c component of this code is zero (Because of alternate +ve & -ve pulses)

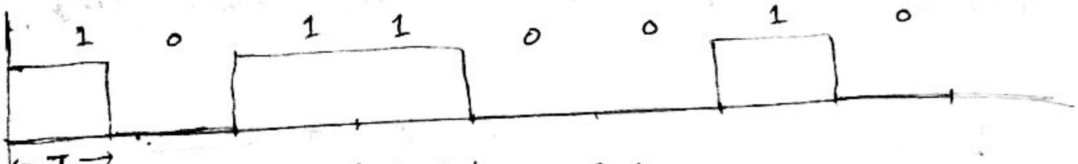
→ In telegraphy, the words 'Mark' and 'Space' are used for symbol '1' and '0' respectively.

→ In differential code, the symbol '1' i.e 'Mark' is alternately represented by waveforms which are inverse with each other, that is why it is called Alternate Mark Inversion (AMI).

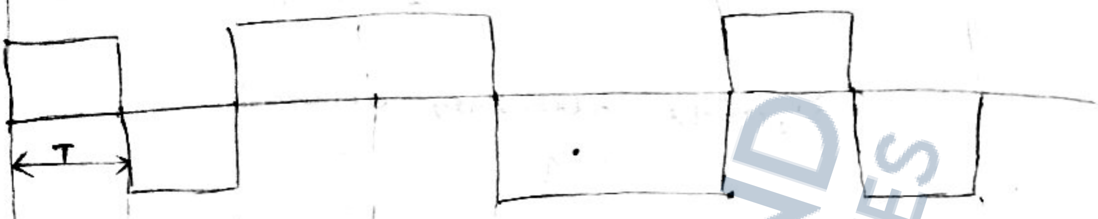
→ Since differential code is basically a BRZ code, another name for this code is 'BRZAMI' code.

(51)

:- Waveform of 10110010 for different line code :-



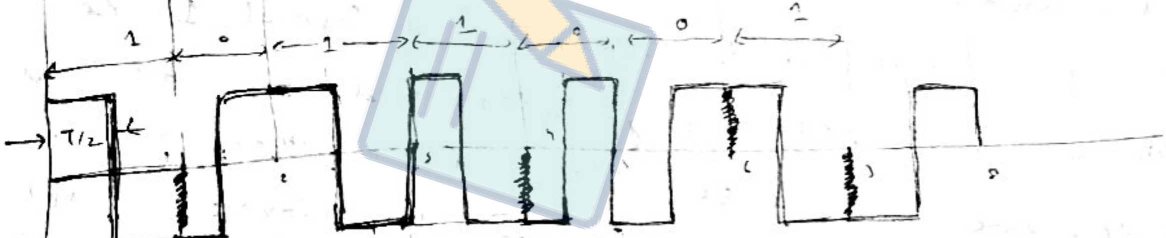
(a) UNRZ Code



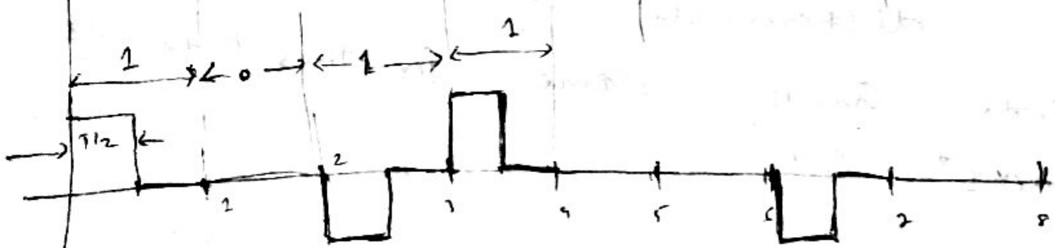
(b) BNRZ Code

(c) URZ Code

(d) BRZ Code



(e) Split Phase Code



(f) Differential (BRZ - AMI) Code

(52)

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Properties

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7) Pow

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Properties of Line Code :-

1) Transmission Bandwidth:-

The minm bandwidth required depends on the highest fundamental frequency of the waveform. This should be as small as possible.

2) Favourable Power Spectral Density:-

The signal spectrum should be matched to the channel frequency response. Zero d.c component is preferable.

3.) Timing (clock) Recovery:-

It should be possible to extract timing or clock information from the signal.

4) Error Detection & Correction Capability.

5) Ease detection and decoding.

6) Transparency:- It should be possible to correctly transmit a digital signal regardless of 0's and 1's.

7) Power efficiency:-

For a given BW and a specified error probability, the transmitted power for a line code should be as small as possible.

Properties of Line Codes:-

Line Code	Minimum Bandwidth	Average DC	Clock Recovery	Error Detection
UNRZ	$1/2T$	$V/2$	Poor	No
BNRZ	$1/2T$	0	Poor	No
URZ	$1/T$	$V/2$	Good	No
BRZ	$1/T$	0	very Good	No
Manchester	$1/T$	0	Best	No
BRZ_AMI	$1/2T$	0	Good	Yes

→ In Manchester Code, a transition occurs in the center of every time slot. Thus, it produces a strong timing component for clock recovery. Hence its clock recovery is the best.

In BRZ-AMI code, positive and -ve pulses ~~are~~ occurs alternately. An error in any bit reception will disturb this property and either two or more consecutive positive pulses or two more consecutive -ve pulses will be received. Thus, this code has built-in error detection capability.

From the above table, it is seen that BRZ-AMI code has the best overall characteristics among all the six codes.

Bandwidth of the PCM System:-

→ Assume that there are n channels, each bandlimited to f_m, to be time division multiplexed.

→ Let N be the length of PCM code so that there are $2^N = M$ quantization levels.

The BW of the PCM system depends on the bit duration (bit time slot).

Sampling frequency $= 2f_m$ and

Sampling period $= \frac{1}{2f_m}$.

There are 'n' channels and N bits per sample and one synchronizing bit, the total number of bits / sampling period (or frame)

$$= nN + 1.$$

$\therefore$ Bit duration $(T_b) = \frac{\text{Sampling period}}{\text{total number of bits}}$

Hence $T_b = \frac{\frac{1}{2f_m}}{(nN+1)} = \frac{1}{2(nN+1)f_m}$ sec — (1)

→ For evaluating the BW, it is assumed that 1's & 0's occure alternatively. Hence, the bit stream in PCM is equivalent to a square wave of pulse width T_b.

→ The practical BW of such a signal is

$$BW = \frac{1}{T_b} \quad \text{—— (2)}$$

Using eq^n (1) in eq^n (2), we have

$$BW = \frac{1}{\left(\frac{1}{2(nN+1)f_m}\right)} = 2(nN+1)f_m \text{ HZ}.$$

If $N >> 1$, & $n >> 1$ (in practical situation)

Then $\boxed{BW = 2nNf_m \text{ HZ}}$

55

Ex:-1) 24 telephone channels, each band limited to 3.4 kHz are to be time division multiplexed by using PCM. Calculate the BW of the PCM system for 128 quantized levels and an 8 kHz sampling frequency.

Ans: Given $\boxed{n = \text{no of channels} = 24}$

$M = \text{quantization levels} = 128$

$$M = 2^N$$

$$\Rightarrow 128 = 2^N$$

$$\Rightarrow \boxed{N = 7} \quad (\text{no of bits})$$

Sampling frequency $(2f_m) = 8\ \text{kHz}$.

using formula,

$$BW = (nN+1)\,2f_m$$

$$= [24 \times 7 + 1] \times 8{,}000$$

$$\boxed{BW = 1.352\ \text{MHz}}$$

using approximate BW formula.

$$BW = 2nNf_m$$

$$= nN(2f_m)$$

$$= nN\,(\text{Sampling freq})$$

$$= 24 \times 7 \times 8000$$

$$\boxed{BW \simeq 1.344\ \text{MHz}}$$

Note: If same no. of channels are FDM by using an SSB modulation, the requ.

annels, each band
time division multiplexed
BW of the
zed levels and

nels = 24

on levels = 128

bits)

= 8 KHZ.

FDM by
tion, the requi.

(56)

BW, assuming 4 KHZ per channel, will be

$$BW = 24 \times 4\ KHz = 96 KHz.$$

∴ This clearly shows BW requirement for PCM system is more.

Noise in PCM system:-

There are two major sources of noise in a PCM system:

(a) Transmission noise introduced outside the transmitter

(b) Quantization noise introduced in the transmitter.

Quantization noise:-

In the PCM transmitter, a quantized value of the sample is encoded instead of actual value. Hence, an error occures. As the difference between actual value and the quantized value of the sample is random, this difference or error may be viewed as noise due to quantization.

Let the message signal be m(t) and let there be M equal voltage intervals, each having magnitude of 'S' volt.

At the center of each voltage interval there are quantization levels $m_1, m_2, \ldots m_m$ as shown in fig (3.7a). The dashed level represents the actual sample value of the message signal m(t) at a time t.

Let m(t) be closest to the quantization level m_k. Then the quantized o/p will be m_k. The quantization error is then

$$e = m(t) - m_k$$

(57)

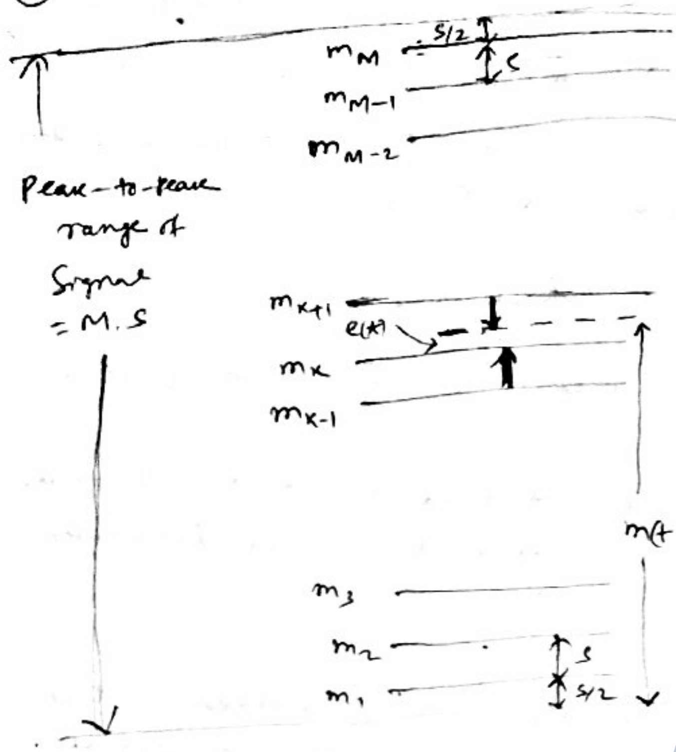

fig 3(a)

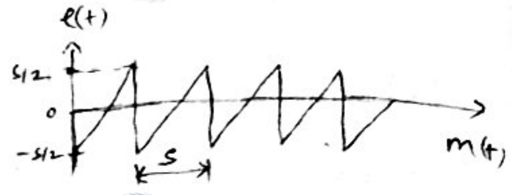

fig 3(b)

Fig 3(b) gives the error voltage e(t) as a function of the instantaneous value of the signal m(t).

Let $\underline{f(m)\,dm}$ be the probability that m(t) lies in the voltage range $\left(m - \frac{dm}{2}\right)$ to $\left(m + \frac{dm}{2}\right)$.

Then mean square quantization error or quantisation noise, is

$\overline{e^2}$ or N_q

Mean square value of a random variable, x is $\overline{x^2} = E[x^2] = \int_{-\infty}^{\infty} x^2 f(m)\,dx$, f(m) = PDF

$$N_q = \int_{m_1 - \frac{S}{2}}^{m_1 + \frac{S}{2}} f(m)\,(m - m_1)^2\,dm + \int_{m_2 - S/2}^{m_2 + S/2} f(m)\,(m - m_2)^2\,dm + \cdots$$

$$\cdots + \int_{m_M - \frac{S}{2}}^{m_M + \frac{S}{2}} f(m)\,(m - m_M)^2\,dm \quad (1)$$

If the number of quantization levels M is large (as is the usual case), it can be assumed that the probability density function f(m) is

(58)

Constant within each quantization range, because in this case the step size 's' is very small as compared to the peak to peak rang M.S of the message signal.

Let $f(m) = f^{(1)}$, in the first term of eqⁿ (1)

$f(m) = f^{(2)}$, in the second term of eqⁿ (1)

⋮

$f(m) = f^{(M)}$, in the last term of eqⁿ (1).

So eqⁿ (1) becomes,

$$N_q = f^{(1)} \int_{m_1 - s/2}^{m_1 + s/2} (m - m_1)^2\, dm + f^{(2)} \int_{m_2 - s/2}^{m_2 + s/2} (m - m_2)^2\, dm + \cdots$$

$$\cdots + f^{(M)} \int_{m_M - \frac{s}{2}}^{m_M + s/2} (m - m_M)^2\, dm \quad \text{—(2)}$$

Let $x = m - m_k$, $\Rightarrow dx = dm$, and in all term in eqⁿ (2) the range of integration becomes $-\frac{s}{2}$ to $\frac{s}{2}$.

(Note :- e.g. $x = m - m_1 \Rightarrow m_1 = m - x$

$m_1 - \frac{s}{2} = m - x - \frac{s}{2} = -\frac{s}{2}$

$m_1 + \frac{s}{2} = m - x + \frac{s}{2} = +\frac{s}{2}$)

∴ Eqⁿ (2) becomes,

$$N_q = f^{(1)} \int_{-s/2}^{s/2} x^2\, dx + f^{(2)} \int_{-s/2}^{s/2} x^2\, dx \cdots\cdots + f^{(M)} \int_{-s/2}^{s/2} x^2\, dx$$

P. T. O

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$$N_q = \left[\int_{-S/2}^{S/2} x^2\,dx\right]\left[f^{(1)} + f^{(2)} \cdots f^{(m)}\right]$$

$$= \left.\frac{x^3}{3}\right]_{-S/2}^{S/2}\left[f^{(1)} + f^{(2)} \cdots f^{(m)}\right]$$

$$= \frac{1}{3}\left[\frac{S^3}{8} + \frac{S^3}{8}\right]\left[f^{(1)} + f^{(2)} \cdots + f^{(m)}\right]$$

$$= \frac{1}{3} \times \frac{\cancel{2} \times S^3}{\cancel{8}\,4}\left[f^{(1)} + f^{(2)} \cdots f^{(m)}\right]$$

$$= \frac{S^2}{12}\left[f^{(1)}{}_S + f^{(2)}{}_S \cdots + f^{(m)}{}_S\right] \quad (3)$$

Now $f^{(1)}{}_S$ is the probability that m lies in the first quantization range, $f^{(2)}{}_S$ is the probability that m lies in the 2nd quantization range & so on. Hence the terms in bracket of eqn (3), is the probability that m lies in the entire range of the signal. Hence

$$f^{(1)}{}_S + f^{(2)}{}_S + f^{(3)}{}_S \cdots + f^{(m)}{}_S = 1.$$

$\therefore$ eqn (3) becomes,

$$\boxed{N_q = \frac{S^2}{12}} \quad (4)$$

To Calculate Signal Power

The mean square value of the O/P signal is equal to the mean square value of the quantized samples.

1) (a)
(b)
(c)
(d)
(e)
(f)
2)
(b)
3). (a)
(b)
4)
(c)

$$\therefore S_0 = \overline{m_k^2}$$ [$\because$ The mean square value of O/P signal is equal to the mean square value of quantized samples]

$$= \frac{1}{M}\left[\left(\frac{S}{2}\right)^2 + \left(\frac{3S}{2}\right)^2 + \left(\frac{5S}{2}\right)^2 \cdots + \left\{\frac{(2M-1)S}{2}\right\}^2\right]$$

$$= \frac{S^2}{4M}\left[1^2 + 3^2 + 5^2 \cdots (2M-1)^2\right]$$

$$\simeq \frac{S^2}{\cancel{4M}} \cdot \left(\frac{4M^{\cancel{3}2}}{3}\right) \quad (\text{For large } M)$$

$$S_0 = \frac{S^2M^2}{3} \quad \text{———} (5)$$

Using eq^n (4) & (5), we have O/P signal to quantization noise ratio as

$$\frac{S_0}{N_q} = \frac{\left(\frac{S^2M^2}{3}\right)}{\left(\frac{S^2}{12}\right)} = \frac{\cancel{S^2}M^2}{\cancel{3}} \times \frac{\cancel{12}\,4}{\cancel{S^2}} = 4M^2$$

Assuming $N_q = N_0$,

$$\boxed{\frac{S_0}{N_0} = 4M^2} \quad \text{———} (6)$$

To Calculate Noise figure :-

First we have to find the i/p signal to noise ratio S_i/N_i.

Let the mean square value of noise $= \sigma_n^2$

Therefore $N_i = \sigma_n^2$ ——— (7)

Lets assume that '0' is represented by a '0'-volt level and '1' by 'A' volt level. Assuming an equal probability for 0 and 1, the avg. signal power is

P.T.O

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0

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

The value 'A' is chosen in such a way that it is much larger than the noise σ_n. Lets us say

$$A = K \times \sigma_n$$

where K = Constant

$$\therefore \quad S_i = \frac{A^2}{2} = \frac{K^2 \sigma_n^2}{2} \quad \text{—— (8)}$$

using eqn (7), (8), we have

$$\frac{S_i}{N_i} = \frac{K^2 \sigma_n^2}{2 \cdot \cancel{\sigma_n^2}} = \frac{K^2}{2} \quad \text{—— (9)}$$

using eqn (6) & (9), we have

$$\text{Noise figure},(F) = \frac{S_i/N_i}{S_o/N_o} = \frac{\frac{K^2}{2}}{4M^2} = \frac{K^2}{8M^2}$$

$$\boxed{F = \frac{K^2}{8M^2}}$$

Why DM ?
1) PCM BW is high
2) nf_m (n bit encode)
3) 1 bit encode) per sample

Delta Modulation:- (DM)

→ In delta modulation technique, an analog signal can be encoded into bits. Hence, in one sense, Delta modulation (DM) is also PCM.

→ The Block diagram of DM system is shown in fig (4). The Pulse generator produces a pulse train $P_i(t)$ of +ve pulses.

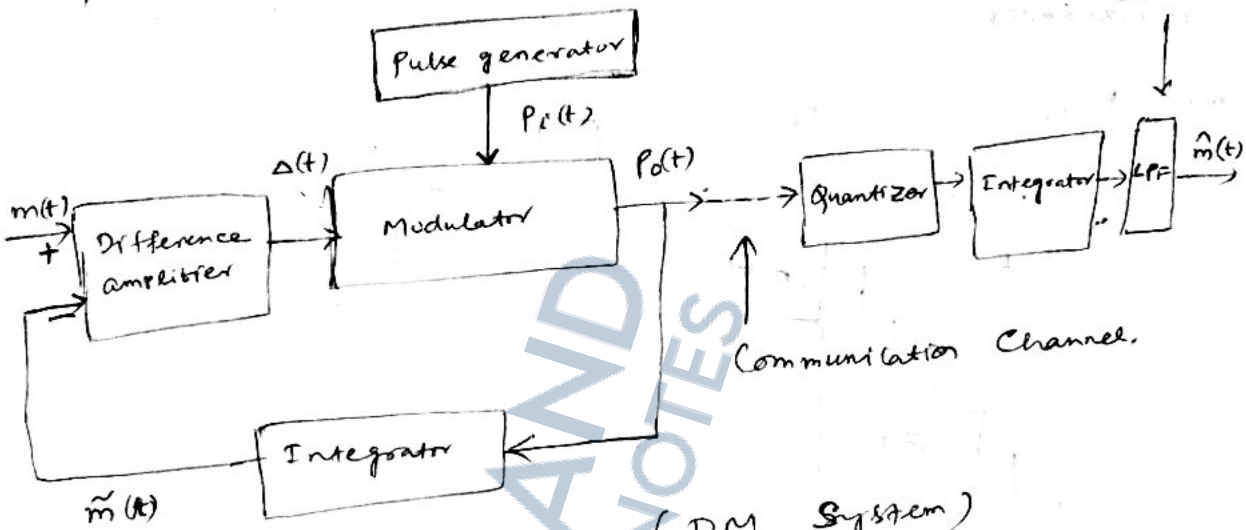


fig:- 4 (DM System)

The modulator receives $P_i(t)$ and $\Delta(t)$.

$\Delta(t)$ = o/p of difference amplifier.

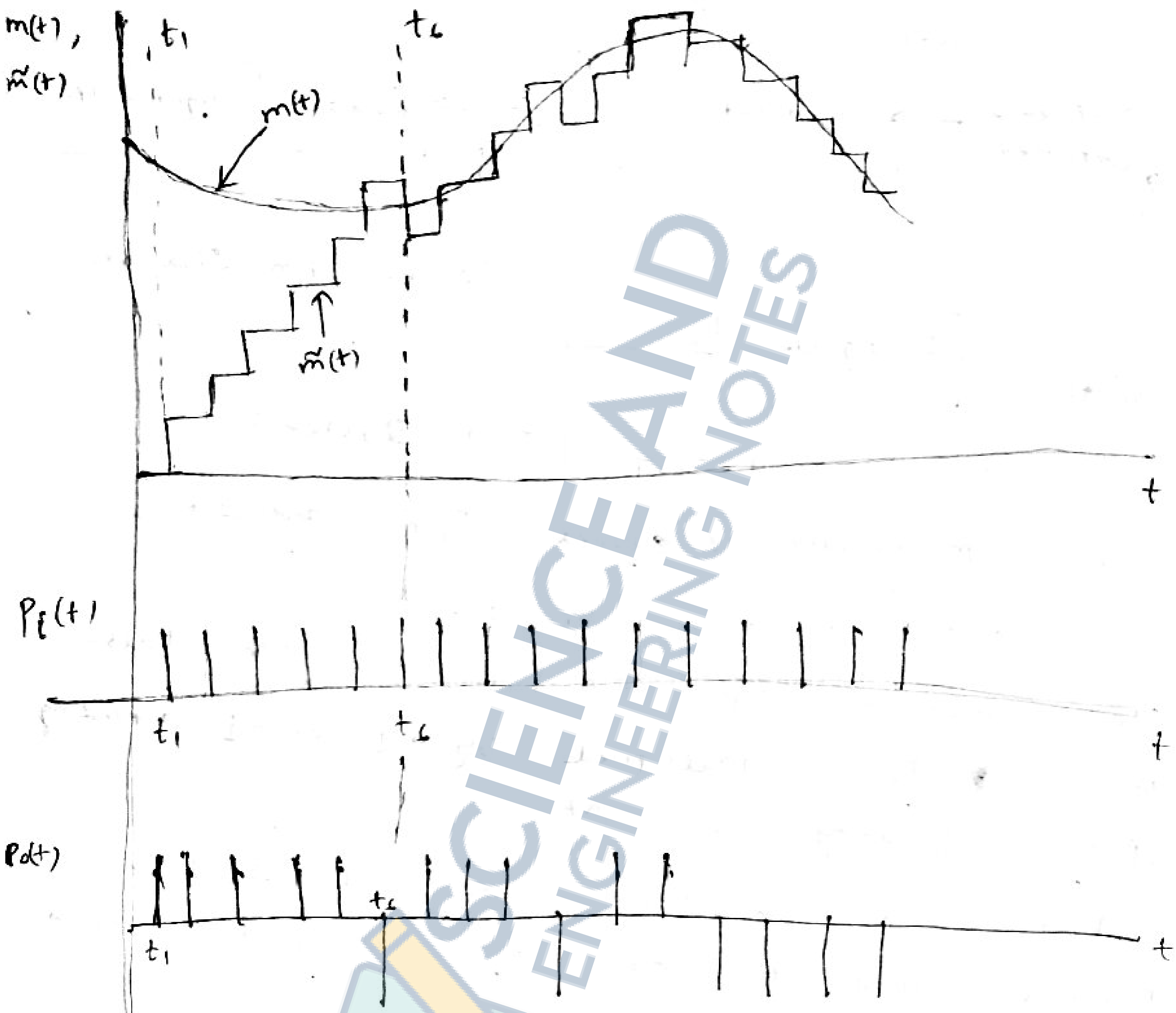
The modulator o/p $P_o(t)$ is the i/p pulse train $P_i(t)$ multiplied by +1 or −1 depending upon polarity of Δt.

→ $P_o(t)$ is a +ve pulse, if $\Delta(t)$ is positive and if is −ve pulse, if $\Delta(t)$ is −ve.

→ The magnitude of $\Delta(t)$ plays no role in deciding $P_o(t)$. Moreover, according to probability theory, the probability of $\Delta(t)$ being exactly zero is zero. and hence $\Delta(t)$ is always either +ve or −ve.

→ The o/p of the modulator $P_o(t)$ is applied to an integrator whose o/p is $\tilde{m}(t)$. The i/p signal $m(t)$ and the integrator o/p $\tilde{m}(t)$ are compared ~~and~~ in a difference amplifier whose o/p is $\Delta(t) = m(t) - \tilde{m}(t)$.

Figure (5) explains the operation of delta modulator.



< Waveform in DM System >

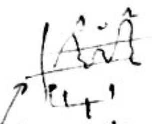
fig - 5

The initial values of $m(t)$ and $\tilde{m}(t)$ have been assumed arbitrarily. At time t_1 of the first pulse in $P_i(t)$, the situation is such that $\Delta(t)$ is positive. Hence, the first pulse in $P_o(t)$ is positive.

In the same way, the pulses in $P_o(t)$ are

(64) either positive or −ve depending upon whether $\Delta(t)$ is positive or −ve. [e.g at time t_6, Δt is −ve and hence $P_o(t)$ is a −ve pulse).

The waveform $\tilde{m}(t)$ approaches $m(t)$ in the form of a staircase and then closely follow it. Thus $\tilde{m}(t)$ is an approximation to the i/p signal $m(t)$.



The waveform $P_o(t)$ is transmitted. At the receiver side, the quantizer takes a decision whether the received pulse is +ve or −ve. Hence, assuming no error, the o/p of the quantizer is same as the waveform $P_o(t)$ and is fed to an integrator, whose o/p takes the form of the waveform $\tilde{m}(t)$.

The LPF then smoothens the o/p of the integrator and gives a waveform $\tilde{m}(t)$ which is similar to the signal $m(t)$.

As the information regarding the difference signal $\Delta(t) = m(t) - \tilde{m}(t)$ is transmitted in this method, it is known as delta modulation.

Limitation of DM:-

The waveform $\tilde{m}(t)$ needs to closely follow the waveform $m(t)$, only then the recovered waveform $\tilde{m}(t)$ resembles $m(t)$.

Fig 6(a) shows a situation where waveform $\tilde{m}(t)$ is unable to follow $m(t)$ because slope of $m(t)$ is greater than slope of $\tilde{m}(t)$.

In fig 6(b), slope of $m(t)$ is more −ve than slope of $\tilde{m}(t)$. In both the cases, the recovered waveform will be distorted. The DM system is, then said to have slope overload.

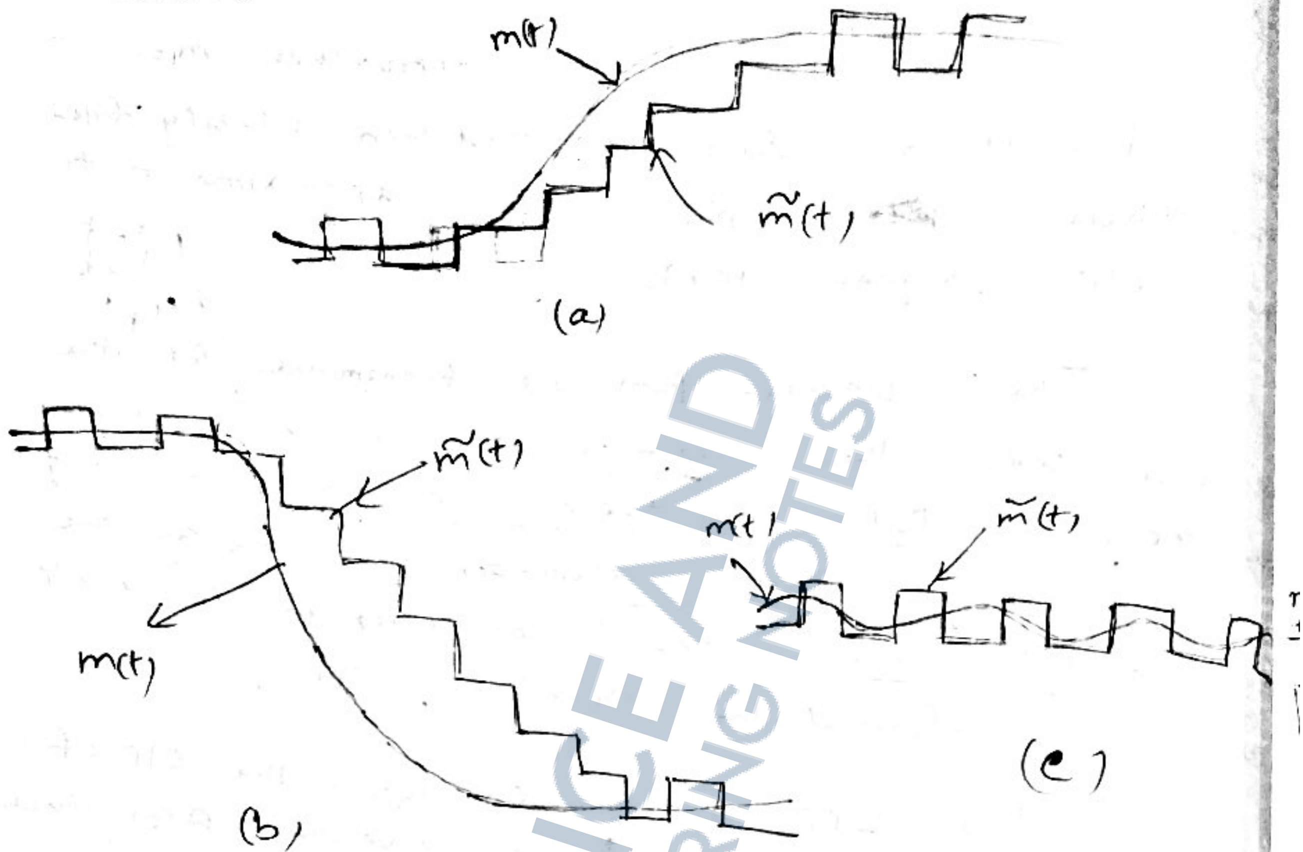


Fig 6 - Limitation of DM (a) Slope overload (+ve) (b) Slope overload (-ve) (c) Slow varying signal

In Fig 6(c), the variation in m(t) are such that they are within the step size. Hence waveform m̃(t) is like a square wave. This will be recovered as d.c, where the original signal m(t) is not d.c. Thus, in this case also, distortion resulted and the noise is known as granular noise.

Adaptive Delta Modulation (ADM)

→ The limitation of DM can be overcome by suitably changing the step size.

→ Slope-overload can be overcome if the step size is increased in such a way that the magnitude of the slope of $\tilde{m}(t)$ becomes greater than the magnitude of the slope of $m(t)$ and when the signal variation are less than the step size, the step size may be reduced to take care of the situation.

A DM system which adjusts its step size is known as the Adaptive Delta Modulation (ADM) system.

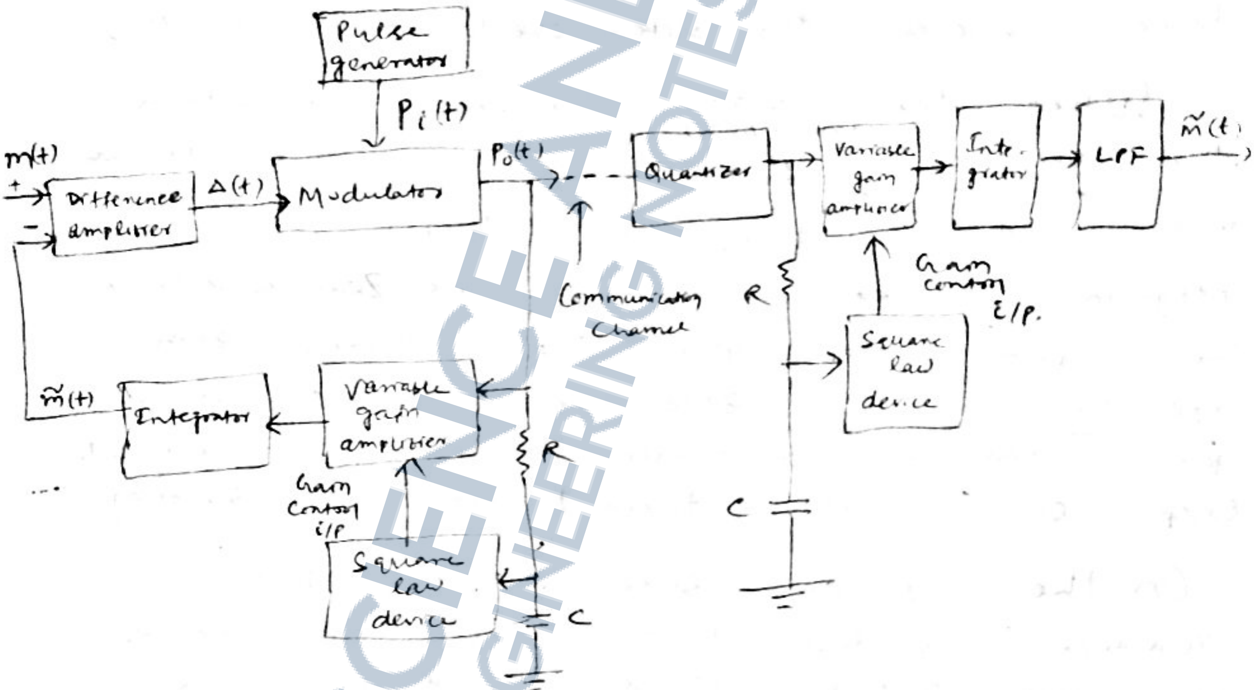


fig (7):- ADM System

→ On the transmitter side, a variable gain amplifier is used before the integrator with $P_o(t)$ as its i/p.

→ The gain of this amplifier depends on the gain control i/p, which is obtained by integrating $P_o(t)$ in an RC-network and then passing the integrator o/p through a square law device.

→ Under slope overload condition $P_o(t)$ is a long sequence of either +ve or −ve pulses. The RC integrator integrates these pulses.

Thus, the o/p of this integrator is either of large +ve or large -ve value.

→ The square law device o/p is of a large +ve value, irrespective of whether the i/p is +ve or -ve. Thus the gain control i/p of the variable gain amplifier is large and its gain increases.

→ Hence the step size increases, which can take care of the slope-overload (+ve or -ve)

→ When the signal variations are within the step size (granular Noise), $P_0(t)$ is a sequence of alternate +ve and -ve pulses. [fig 6(c)]. The RC integrator o/p in this case is zero and hence the gain control i/p of the variable gain amplifier is also zero. The gain of the variable gain amplifier decreases, resulting in a reduced step size, which takes care of the situation.

→ On the receiver side, the o/p of the quantizer is fed to a variable gain amplifier whose gain control i/p is derived from an RC integrator and a square law device. Thus, an adaptive adjustment of the step size is obtained at the receiver, resulting in an undistorted reception of the transmitted signal.

Noise in delta modulation:-

Quantization Noise:-

The quantization error in DM is given by

$$\Delta(t) = m(t) - \tilde{m}(t)$$

The max^m quantization error in DM is S. (In PCM it is $\frac{S}{2}$). If it is assumed that

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is given by

DM is S.
is assumed that

(68) the error $\Delta(t)$ takes on all values between $-S$ to $+S$ with equal likelihood, then the probability density of $\Delta(t)$ is

$$f(\Delta) = \frac{1}{2S}, \quad -S \leq \Delta \leq +S$$

The normalized power of $\Delta(t)$ is then,

$$\overline{[\Delta(t)]^2} = \int_{-S}^{S} \Delta^2 f(\Delta)\, d\Delta$$

$$= \int_{-S}^{S} \Delta^2 \cdot \frac{1}{2S}\, d\Delta \qquad \left(\because f(\Delta) = \frac{1}{2S}\right)$$

$$= \frac{1}{2S} \cdot \left[\frac{\Delta^3}{3}\right]_{-S}^{S}$$

$$= \frac{1}{6S}\left[2S^3\right]$$

$$\overline{[\Delta(t)]^2} = \frac{S^2}{3}.$$

It can be reasonably assumed that the spectrum $\Delta(t)$ is white over the frequency range 0 to f_b where $\boxed{f_b = \frac{1}{\tau}}$ [τ being the step duration.]

The quantization noise power in the frequency range 0 to f_b is $\frac{S^2}{3}$. Hence the o/p noise power in the baseband frequency range 0 to f_m (f_m being the upper limit of baseband frequency range).

$f_b \rightarrow \frac{S^2}{3}$
$1 \rightarrow \frac{S^2}{3f_b}$
$f_m \rightarrow \frac{S^2}{3f_b} \cdot f_m$

$$N_q = \frac{S^2}{3} \cdot \frac{f_m}{f_b} = \frac{S^2 f_m}{3 f_b}$$

$$\boxed{N_q = \frac{S^2 f_m}{3 f_b}} \quad \text{——} \quad (1)$$

①

O/P Signal Power:-

In PCM, the signal excursion limits are $\left(-\frac{MS}{2}\right)$ to $\left(+\frac{MS}{2}\right)$, where 'S' is the step size and 'M' is the number of quantization levels.

On the other hand, in DM there is no such limit on the amplitude of the signal waveform. Rather there is a limitation on the slope of the signal waveform in order to avoid slope overload.

Let the worst case of signal power being concentrated in the upper ~~band~~ end of the baseband be assumed. (When the signal power is concentrated in the lower end of the baseband, i.e when the signal waveform changes slowly, there is nominally no limit to the signal power which may be transmitted.)

i.e Let

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

where

$$A = \text{Amplitude}$$

and $\omega_m = 2\pi f_M$, f_M being the upper limit of the baseband frequency range.

Then, the O/P signal power is

$$S_o = \overline{m^2(t)} = \frac{A^2}{2} \quad \text{——— (2)}$$

The max^m slope of $m(t)$ is $\omega_m A$.

The max^m avg. slope of DM approximation $\tilde{m}(t)$ is

$$\frac{S}{\tau} = S f_b \qquad \left(\because f_b = \frac{1}{\tau}\right)$$

The limiting value of A just before the start

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S' is the step size
quantization levels.
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—— ②

is $\omega_m A$.

approximation $\tilde{m}(t)$

$f_b = \frac{1}{\tau}$)

before the start

---

(71) of the slope overload is then given by the condition

$$\omega_m A = S f_b \quad —— ③$$

$$\Rightarrow A = \frac{S f_b}{\omega_m}$$

From eqn ②,

$$S_o = \frac{A^2}{2} = \frac{\left(\frac{S f_b}{\omega_m}\right)^2}{2} = \frac{S^2 f_b^2}{2\omega_m^2} \quad —— ③$$

~~From eqn~~

O/P signal to Quantization Noise ratio (S/N_q)

From eqn ③, we have $S_o = \frac{S^2 f_b^2}{2\omega_m^2}$.

from eqn ①, we have $N_q = \frac{S^2 f_m}{3 f_b}$

$$\frac{S_o}{N_q} = \frac{\cancel{S^2} f_b^2}{2\omega_m^2} \times \frac{3 f_b}{\cancel{S^2} f_m} = \frac{3 f_b^3}{2 \times (2\pi f_m)^2 \cdot f_m}$$

$$\frac{S_o}{N_q} = \frac{3 f_b^3}{8\pi^2 f_m^3} = \frac{3}{8\pi^2} \cdot \left(\frac{f_b}{f_m}\right)^3$$

$$\boxed{\frac{S_o}{N_q} = \frac{3}{8\pi^2} \cdot \left(\frac{f_b}{f_m}\right)^3} \quad —— (4)$$

$$\boxed{\frac{S_o}{N_q} \simeq \frac{3}{80} \left(\frac{f_b}{f_m}\right)^3}$$

This value of $\frac{S_o}{N_q}$ is the worst case value. The actual value S_o/N_q is greater than the value given by eqn (4). It is found that the value of $\frac{S_o}{N_q}$ comes out to be $\boxed{\frac{3}{64}\left(\frac{f_b}{f_m}\right)^3}$

Comparision between PCM and DM:-

→ DM needs a simple ckt as compared to PCM.

→ But the signal to quantization noise ratio is less in DM than in PCM, because in latter case the max^m possible error due to quantization is $\frac{S}{2}$ where, it is 'S' in the former.

→ Moreover, it has been found experimentally that, for voice transmission, the bit rate needed by PCM, assuming 8kHz sampling rate and 7 quantization level is, 56 kilobit per second, where as, the bit rate needed by DM for same quality of voice transmission is much higher than 56 kilobits per second.

For quality transmission, BW needed by DM is more than PCM.

→ On the otherhand, if BW conservation is the main criterion (at the cost of quality of transmission), then DM is preferable over PCM because in case of slightly substandard quality of transmission, the BW needed by DM is less than that needed by PCM.

→ Therefore, the use of DM can be recommended for the following two situations.

(i) When BW conservation is desirable at the cost of quality of transmission.

(ii) When simple circuitry is utmost importance and allowable BW is large.

1

DM	PCM
1) Simple circuitry than PCM	1) Complex circuitry than DM
2) Signal to quantization noise ratio is less.	2) Signal to quantization noise ratio is more.
3) For quality transmission, BW needed by DM is more than PCM.	3) For quality transmission, BW needed by PCM is less than DM.
4) If quality is compromised & BW conservation is the main criterion, BW needed by DM is less than PCM.	4) If BW conservation is main criterion rather than quality, BW needed by DM is less than by PCM or BW is more in case of PCM than DM.

Delta or Differential PCM (DPCM)

→ In DPCM, instead of quantizing each sample, the difference between 2 successive samples is quantized, encoded and transmitted as in PCM.

→ This is particularly useful in voice transmission, because in this case two successive samples don't differ much in amplitude. Thus, the difference signal is much less in amplitude than actual sample and hence less number of quantization levels are needed.

→ Therefore, the number of bits per code is reduced resulting in a reduced bit rate. Thus the BW required in this case is less than one required in PCM.

→ The disadvantage of DPCM is that the modulator & demodulator ckt are more complicated than those in PCM.

S - ARY System:-

In a binary system, pulses with one of the two possible levels are used.

In S-ary system, the pulses are allowed to take one of the S possible levels $(S > 2)$. Each level corresponds to a distinct i/p symbol.

e.g: in a quaternary system $(S=4)$, the levels may be 0V, 1V, 2V, 3V and respective codes may be 0, 1, 2, 3. If there are M quantization levels, then we need

$\log_S M$ symbols to represent a sample

e.g. If $M = 64$, $S = 4$, we need $\log_4 64 = 3$ symbols to represent a sample. As a comparison in a binary system, we need $\log_2 64 = 6$ symbols to represent a sample. As the number of symbols needed to represent a sample is less, the S-ary system needs less BW than the binary system.

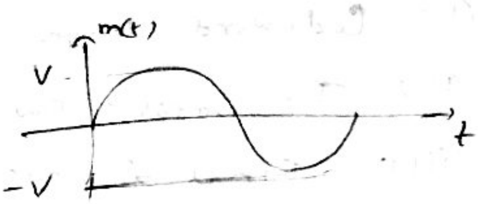
The disadvantage of S-ary system is that for a given probability of error, the needed transmitter power is more because noise is more effective due to a large number of voltage levels.

It can be shown that, for $S >> 2$ and $P_e << 1$, the BW is reduced by $\frac{1}{\log_2 S}$ and transmitter power is to be increased by a factor of $\frac{S^2}{\log_2 S}$.

The circuitry for S-ary system is more complicated because the receiver has to decide on one of the S-levels using S-1 comparator or level slicers.

1

(74) Q/ 1) Derive an expression for a signal to quantization noise ratio for a PCM system which employs linear (uniform) quantization technique. Given that i/p to PCM system is a sinusoidal signal.



Ans :-

$$V_H - V_L = V-(-V) = 2V.$$

$$m = \text{no. of Quantization levels.}$$

$$\text{Step size } (s) = \frac{2V}{m}.$$

$$N_q = \frac{s^2}{12} = \frac{\left(\frac{2V}{m}\right)^2}{12} = \frac{\cancel{4}V^2}{m^2\,\cancel{12}\,3} = \frac{V^2}{3m^2}.$$

$$SNR = \frac{S_o}{N_q} = \frac{\frac{V^2}{2}}{\frac{V^2}{3m^2}} = \frac{V^2}{2} \times \frac{3m^2}{\cancel{V^2}} = \frac{3}{2}m^2.$$

$\because m = 2^n$

$n =$ no of bits.

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

$$= \frac{3}{2}\cdot 2^{2n}.$$

$$(SN_qR)_{dB} = 10\log\left(\frac{3}{2}\cdot 2^{2n}\right)$$

$$= 10\left[\log\frac{3}{2} + \log 2^{2n}\right]$$

$$= 10\left[0.1761 + 2n\log 2\right]$$

$$\boxed{SN_qR = (1.761 + 6.02n)\ dB}$$

2) A T.V signal having BW of 4.2 MHz is transmitted using binary PCM. Given that number of quantization levels is 512. Determine

(i) Codeword length
(ii) Transmission BW
(iii) Final bit rate
(iv) O/P signal to quantization noise ratio.

Ans :- (i) No of quantization level = 512.

$$2^N = 512$$

$$\Rightarrow \boxed{N = 9}$$

$\therefore$ Codelength = 9 bits.

(ii)

$$\boxed{\text{Transmission BW} = N f_m}$$

T.V signal has BW = 4.2 MHz. This means highest freq. component will have freq 4.2 MHz.

$$\therefore f_m = 4.2\ \text{MHz}.$$

$$\text{Transmission BW} \geq 9 \times 4.2 = 37.8\ \text{MHz}.$$

$$\boxed{BW \geq 37.8\ \text{MHz}}$$

(iii)

$$\text{bit rate} = N f_s \ \text{bits/sec}$$

$$\geq N \times 2 f_m$$

$$\geq 9 \times 2 \times 4.2 \times 10^6$$

$$\geq 75.6 \times 10^6\ \text{bits/sec}$$

$$\boxed{\text{bit rate} \geq 75.6 \times 10^6\ \text{bits/sec}}$$

(iv) O/P signal to quantization ratio

$$SN_qR = 1.761 + 6n \text{ dB}$$

$$\approx 1.8 + 6n$$

$$= 1.8 + 6 \times 9$$

$$\boxed{SN_qR = 55.8 \text{ dB}}$$ ✓

3) A PCM system uses a uniform quantizer followed by a 7 bit binary encoder. The bit rate of the system is equal to 50×10^6 bits/sec.

(i) What is the max^m message signal BW for which the system operates satisfactorily?

(ii) Calculate the SN_qR when full load sinusoidal modulating wave of freq 1 MHz is applied to the i/p.

Ans: Given $n = 7$ bit, bit rate $= 50 \times 10^6$ bits/sec.

$$\cancel{Nf_s \geq 50 \times 10^6}$$

$$\cancel{7 \times 2f_m \geq 50 \times 10^6}$$

$$\cancel{\text{bit rate} \geq 50 \times 10^6}$$

$$\text{bit rate} \geq N f_s$$

$$\Rightarrow 50 \times 10^6 \geq N \times 2f_m$$

$$\Rightarrow 50 \times 10^6 \geq 7 \times 2f_m$$

$$\Rightarrow f_m \leq \frac{50}{14} \times 10^6$$

$$\Rightarrow \boxed{f_m \leq 3.57 \text{ MHz}}$$

$\therefore$ max^m signal BW $= 3.57$ MHz.

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(ii) $SN_qR = 1.8 + 6n \text{ dB}$

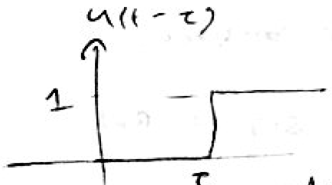
$= 1.8 + 6 \times 7$

$$\boxed{SN_qR = 43.8 \text{ dB}}$$

4. ACF / PSD of $\underline{e^{-at}u(t)}$

Ans:

ACF $R(\tau) = \int_{-\infty}^{\infty} x(t) \cdot x^*(t-\tau)\, dt$



$= \int_{-\infty}^{\infty} e^{-at} u(t)\, e^{-a(t-\tau)} u(t-\tau)\, dt$

$= \int_{\tau}^{\infty} e^{-at} e^{-a(t-\tau)}\, dt$

$= \int_{\tau}^{\infty} e^{-2at} \cdot e^{a\tau}\, dt$

$= e^{a\tau} \left[\frac{e^{-2at}}{-2a} \right]_{\tau}^{\infty}$

$= \frac{e^{a\tau}}{-2a} \left[0 - e^{-2a\tau} \right]$

$= \frac{e^{-a\tau}}{2a}$

$$\boxed{ACF = R(\tau) = \frac{e^{-a|\tau|}}{2a}}$$

ACF $\longleftrightarrow$ ESD (F.T pair)

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①

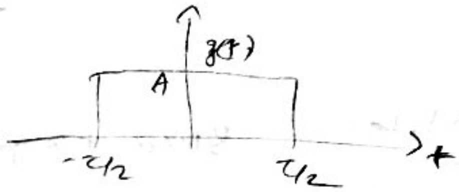
$$ESD = F\left[\frac{1}{2a} e^{-a|t|}\right]$$

$$ESD = F\text{-}T\left[\frac{1}{2a} e^{-a|t|}\right]$$

$$= \frac{1}{2a} \times \frac{2a}{a^2+\omega^2}$$

$$\boxed{ESD = \frac{1}{a^2+\omega^2}}$$

5) ACF, ESD of



$$ESD = |G(\omega)|^2$$

$$G(f) = \int_{-\infty}^{\infty} g(t)\, e^{-j2\pi f t}\, dt$$

$$= \int_{-\tau/2}^{\tau/2} A\, e^{-j2\pi f t}\, dt$$

$$= A \left[\frac{e^{-j2\pi f t}}{-j2\pi f}\right]_{-\tau/2}^{\tau/2}$$

$$= A \cdot \frac{e^{-j2\pi f \frac{\tau}{2}} - e^{j2\pi f \frac{\tau}{2}}}{-2j\pi f}$$

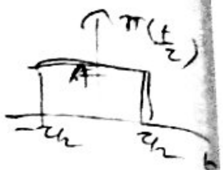
$$= \frac{A}{\pi f} \cdot \cancel{\sin \pi f \tau}\ \sin \pi f \tau$$

$$\cancel{= \frac{2A\tau}{\pi f \tau} \sin \tau} = \frac{A\tau}{\pi f \tau} \sin \pi f \tau$$

$$= A\tau \frac{\sin(\pi \tau f)}{(\pi \tau f)} = A\tau\, \mathrm{sinc}(\tau f)$$

(99)

$$ESD = |g(f)|^2 = (A\tau)^2 \, \text{sinc}^2(f\tau)$$



$$ACF = \int_{-\infty}^{\infty} x(t) \, x^*(t-\tau) \, dt$$

$$= \int_{-\tau/2}^{\tau/2} A \, \Pi\left(\frac{t}{\tau}\right) \cdot A \, \Pi\left(\frac{t-\tau}{\tau}\right) dt$$

$$= A^2 \int_{-\tau/2}^{\tau/2} \Pi\left(\frac{t}{\tau}\right) \cdot \Pi\left(\frac{t-\tau}{\tau}\right) dt$$

Note: Block diagram of a PAM/TDM System:-