

Pulse Modulation

Channel Capacity

The quantity that gives the relative strength of signal to noise is called Signal to Noise Ratio (SNR). It is usually expressed in dB.

$$\text{SNR in dB} = 10 \log_{10} (\text{signal energy}/\text{noise energy})$$

The unit dBW is used to measure signal strength reference to 1 Watt where, 0 dBW = 1 Watt. Similarly dBm gives signal strength referenced to 1 mW where, 0 dBm = 1 mWatt.

The power that can be transmitted is finite and usual noise spectrum is flat (also called white). Bandwidth of a channel is defined as the range of frequencies which can be transmitted through a channel within acceptable degradation.

To transmit any given message if we use larger bandwidth then signal power required to send it for any given quality is usually less. This depend how an input message is encoded into an electrical signal. If channel bandwidth is increased by a factor n, required SNR is approximately reduced to $\frac{1}{n}$ th root. Small increase in bandwidth reduces SNR requirement to a great extent. Thus there is a bandwidth required-SNR tradeoff.

The capacity of a channel to support a particular rate of information is defined as channel capacity.

The upper limit of channel capacity comes from Shannon's equation

$$C = B \log_2 (1 + SNR) \text{ bits/sec} \quad \dots \textcircled{1}$$

The significance of this equation is that digital data theoretically can be sent at any rate less than C without any error. Practical systems are yet to achieve this upper limit and various coding techniques are employed for this purpose. A complex coding technique can come close to this capacity but issues like time and other overhead may limit its use.

It clearly shows that increase in Bandwidth (B) reduces required SNR when information rate is constant and vice-versa. If noise is zero, $C = \infty$, i.e. theoretically, under no noise condition any information rate is supported.

For channel with white noise or flat noise spectrum of strength N_0 the eqn can be modified

to $C = B \log_2 [1 + S/(N_0 B)] \text{ bits/sec} \quad \dots \textcircled{2}$

⊕ Consider, $B = 4 \text{ kHz}, SNR = 20 \text{ dB}$

$$= 10^{\frac{20}{10}} = 100$$

$$\therefore C = 4000 \log_2 (1 + 100) \text{ bits/sec}$$

=

⊕ If $B = 1 \text{ MHz}, C \text{ required is } 32 \text{ kbits/sec}$

$$\text{required SNR} = 2^{\frac{32000/10^6}{1}} - 1 =$$

$$SNR \text{ in dB} = 10 \log_{10} (\text{SNR}) =$$

Need of Analog to Digital Conversion :-

The basic problem is associated with the transmission of a signal over a noisy communication channel. Suppose a telephone conversation is to be transmitted from midnatorre to Kolkata. If the signal is transmitted by radio, the signal arrives at destination with great attenuation and combined with noise. The receiver noise is often the noise source of largest power.

Again if we expect a telephone cable to produce an attenuation of the order of 0.5 dB per Km, for a

1000 km run, to get receive a signal of 1 mv, the voltage at the transmitting end would have to be 10^{22} volts.

An amplifier at the receiver end will not help the above situation. Suppose a repeater is introduced at the 1st half of the communication path. This repeater will raise the signal level, in addition, it will raise the level of the noise.

Such a midway repeater with an amplifier at the receiver has the advantage of improving signal to noise ratio. Many such repeaters have to be imposed for a long communication channel.

If we were to transmit a digital signal over the same channel, we find that digital signal ~~sig~~ need significantly less power. In practice we find that SNR of 40-60 dB are required for analog signals while 10-20 dB are required for digital signals.

Sampling theorem and Low pass signal:-

The fundamental principle of digital communications
the Sampling theorem:

Let $m(t)$ be a signal which is bandlimited such that its highest frequency spectral component is f_m . Let the values of $m(t)$ be determined at regular intervals separated by times $T_s \leq \frac{1}{2f_m}$, that is the signal is periodically sampled every T_s seconds. Then these ~~signals~~ samples $m(nT_s)$, where n is an integer, uniquely determine the signal, and the signal may be reconstructed from these samples with no distortion.

The time T_s is called Sampling time. The theorem requires that the sampling rate be rapid enough so that atleast two samples are taken during the course of the period corresponding to the highest frequency spectral component.

The base band signal $m(t)$ which is to be sampled is shown in fig 4.1a. A periodic train of pulses $s(t)$ of unit amplitude and of period T_s is shown in fig 4.1b. The pulses are arbitrarily narrow, having

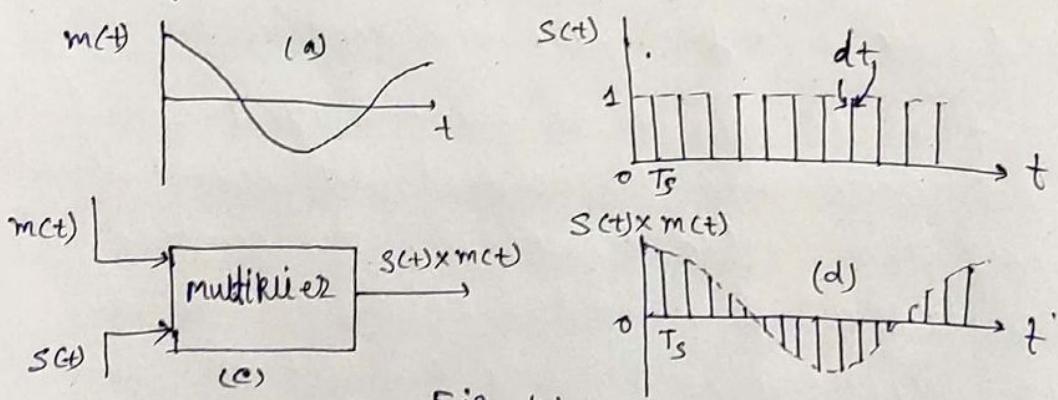


Fig - 4.1

- (a) ~~act~~ signal $m(t)$ to be sampled
- (b) Sampling function $s(t)$
- (c) Sampling operation
- (d) Sampled signal

a width dt . The two signals are applied to a multiplier and shown in fig 4.1c, which then yields as an output the product $s(t)m(t)$ as in the product as in fig.4.1d to be the signal $m(t)$ sampled at occurrence of each pulse. When a pulse occurs, the multiplier output has the same value as $m(t)$ and all other times the multiplier output is zero.

The signal $s(t)$ is periodic with period T_s and has Fourier expansion

$$s(t) = \frac{dt}{T_s} + \frac{2dt}{T_s} \left(\cos 2\pi \frac{t}{T_s} + \cos 2 \times 2\pi \frac{t}{T_s} + \dots \right)$$

For $T_s = \frac{1}{2f_m}$, the product $s(t)m(t)$ is

$$s(t)m(t) = \frac{dt}{T_s} m(t) + \frac{dt}{T_s} \left\{ 2m(t) \cos 2\pi(2f_m)t + 2m(t) \cos 2\pi(4f_m)t + \dots \right\} \quad (3)$$

The 1st term in the series is aside from a constant factor, the signal $m(t)$ itself. The second term is the product of $m(t)$ and $\cos 2\pi(2f_m)t$ which gives rise to a DSB-SC signal with carrier frequency $2f_m$. Similarly the succeeding terms yield DSB-SC signals with carrier frequencies $4f_m, 6f_m$ etc.

Let the signal $m(t)$ have a spectral density $M(j\omega) = F[m(t)]$, the Fourier transform of $m(t)$ which is shown in fig 4.2a. $m(t)$ is band limited to the frequency range below f_m . The 1st term in eqn (3) extends from 0 to f_m . The spectrum of 2nd term

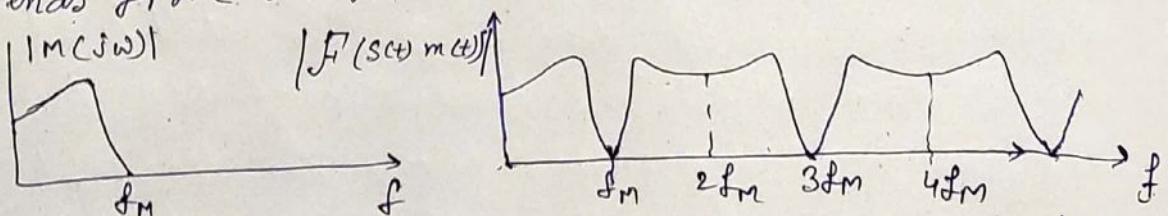


Fig 4.2 (a) magnitude plot of spectral density of a signal
 (b) plot of amplitude of spectrum of sampled signal

is symmetrical about the frequency $2f_m$ and extends from $2f_m$ to $3f_m$ and so on as shown in fig 4.2b. If the sampled signal is passed through an ideal low pass filter with cutoff frequency f_m , the filter passes $m(t)$ and nothing else.

The spectral pattern of 4.2b with $f_s = \frac{1}{T_s}$ larger than $2f_m$ is shown in 4.3a. In this case there is a gap between the upper limit f_m of the spectrum of the baseband signal and the lower limit of DSB-SC signal, centered around the carrier frequency $f_s > 2f_m$. In this case a practical low pass filter can select $m(t)$. The filter attenuation may begin at f_m but need not attain a high value until the frequency $f_s - f_m$. The range from f_m to $f_s - f_m$ is called a guardband and is always required in practice. Typically when sampling is used in connection with voice messages on telephone lines, the voice signal is limited to $f_m = 3.3$ KHz while f_s is selected at 8.0 KHz. The guard band is $8.0 - 2 \times 3.3 = 1.4$ KHz.

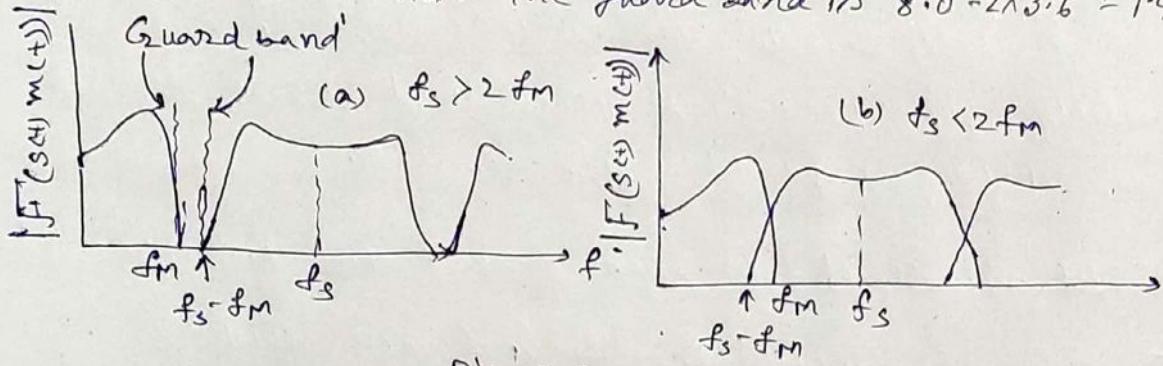


Fig 4.3

(a) guardband appears when $f_s > 2f_m$

(b) overlapping of spectra when $f_s < 2f_m$

The situation depicted in fig 4.3b corresponds to the case $f_s < 2f_m$. Here we find an overlap between the spectrum of $m(t)$ and the spectrum of DSB-SC signal centered around f_s . No filtering operation will allow exact recovery of $m(t)$. This phenomenon is known as aliasing in frequency domain. To avoid

aliasing, an antialiasing filter, a lowpass filter that limits the frequency band of the message $m(t)$ within frequency $f_s/2$ is used.

The minimum sampling rate of a signal $2f_m$ is called Nyquist rate. An increase in sampling rate increases the guard band as well as the band width.

Sampling of Bandpass Signal :-

We consider the case with the signal $m(t)$ where, highest frequency spectral component is f_m and lowest component is $f_L \neq 0$. It may be that the sample frequency need be no larger than $f_s = 2(f_m - f_L)$.

Let us select a sample frequency $f_s = 2(f_m - f_L)$ and let f_L is integral multiple of f_s , i.e. $f_L = n f_s$ with n an integer. The situation is shown in fig. 4.4 below-

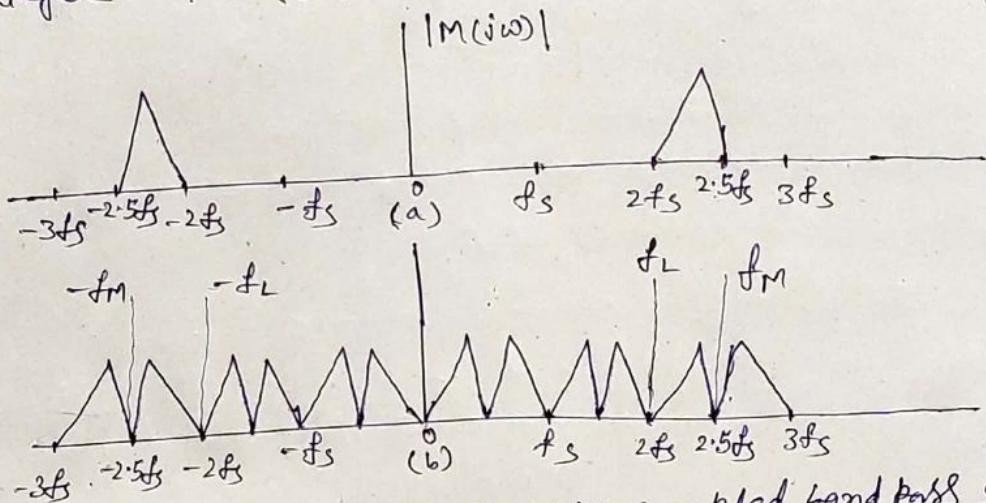


Fig. 4.4
① Spectrum of (a) bandpass signal (b) Sampled band pass signal

In part a, the two sided spectral pattern of signal $m(t)$ with Fourier transform $M(j\omega)$ is shown.

Here $n=2$, f_L coincides with 2nd harmonic of the sampling frequency while the sampling frequency is exactly $f_s = 2(f_m - f_L)$. In part b' spectral pattern of $s(t)m(t)$ is shown. The product of $m(t)$ and dc term of $s(t)$ [eqn ③] leaves it in the same frequency range from f_L to f_m . The other spectral terms of $s(t)$ [$f_s, 2f_s, 3f_s$]

in multiplication with $m(t)$ gives rise to spectral pattern by shifting the pattern in part a to the right and to the left by amount f_s and so on. A band pass filter with sharp cutoffs with pass bands from f_L to f_m can recover the signal $m(t)$.

Since spectrum of $m(t)$ extends over the 1st half of the frequency interval of the sampling frequency, i.e. from $2.0f_s$ to $2.5f_s$, there is no spectral overlap and signal recovery is possible. If $m(t)$ overlaps and signal recovery is possible. If the spectral range of $m(t)$ is extended from $2.5f_s$ to $3.0f_s$, there would similarly be no overlap. If the spectrum of $m(t)$ were not confined within 1st or 2nd half of the interval between sample-frequency harmonics, there would be overlap between the spectrum patterns and signal recovery would not be possible. Hence minimum sampling frequency is $f_s = 2(f_m - f_L)$ when f_m or f_L is a harmonic of f_s .

If f_m or f_L is not a harmonic of f_s , the spectral pattern is shown in Fig 4.5. Let the positive and negative frequency part are called PS and NS.

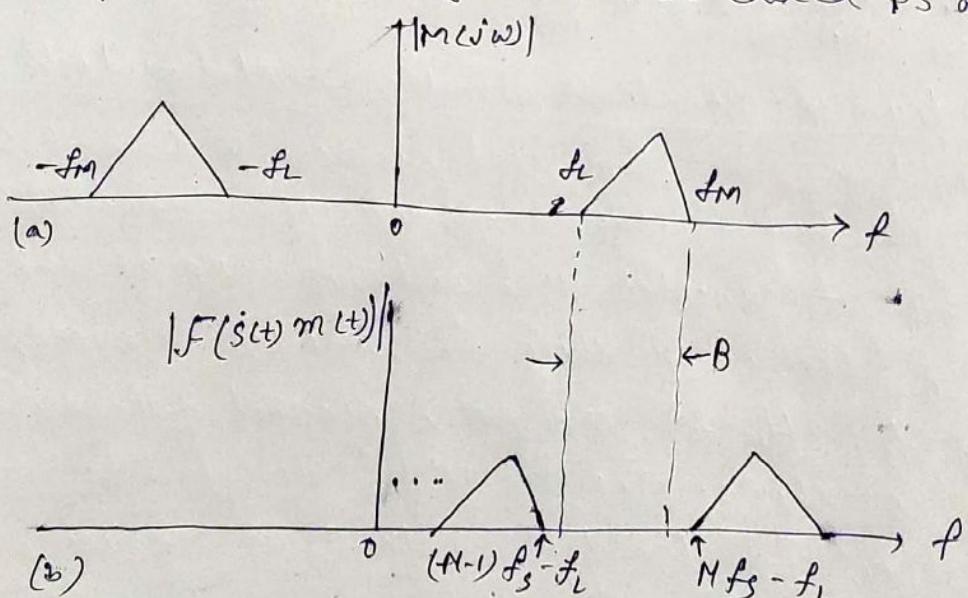


Fig. 4.5 Spectrum of (a) bandpass signal (b) NS shifted by $(N-1)f_s$ and N th harmonic of the sampling waveform

The product of $m\omega$ and the dc component of sampling waveform leaves PS unmoved and it is this part of spectrum which we propose to selectively draw out to reproduce the original signal. If we select minimum value of f_s to be $f_s = 2(f_m - f_L) = 2B$ then the shifted PS pattern will not overlap PS. Same thing happens to NS. To avoid overlap it is necessary that

$$(N-1)f_s - f_L \leq f_L \quad \dots \textcircled{I}$$

$$\text{and } Nf_s - f_m \geq f_m \quad \dots \textcircled{II}$$

with $B = f_m - f_L$ we have

$$\begin{aligned} (N-1)f_s &\leq 2(f_m - B) \\ \text{and } Nf_s &\geq 2f_m \end{aligned} \quad \left. \begin{array}{l} \dots \textcircled{III} \\ \dots \textcircled{IV} \end{array} \right.$$

Let, $\kappa = f_m/B$ eqs \textcircled{III} and \textcircled{IV} becomes

$$f_s \leq 2B \left(\frac{\kappa-1}{\kappa-1} \right) \quad \textcircled{V}$$

and $f_s \geq 2B \left(\frac{\kappa}{\kappa-1} \right) \dots \textcircled{VI}$ in which $\kappa \geq N$ since

$f_s \geq 2B$. Eqs \textcircled{V} and \textcircled{VI} establish the constraint which must be observed to avoid an overlap on PS. The symmetry says that this constraint assures no overlap on NS.

We can write the Bandpass Sampling theorem as follows—

A bandpass signal with highest frequency f_m and bandwidth B , can be recovered from its samples through bandpass filtering by sampling it with frequency $f_s = \frac{2f_m}{\kappa}$, where κ is the largest integer not exceeding $\frac{f_m}{B}$. All frequencies higher than f_s but below $2f_m$ may or may not be useful for bandpass sampling depending on overlap of shifted spectrums.

Pulse Amplitude Modulation and Concept of Time Division Multiplexing:

The technique of sampling by time division multiplexing is illustrated in fig 4.8 in an idealized manner. On the left of the fig as the rotary switch turns around, it samples each signal sequentially. The rotary switch at the receiver end is in synchronism with the switch at the sending end. With each revolution, one sample is taken of ~~all~~ each input signals and presented to the correspondingly numbered contact of the receiving end switch. If f_m is the highest

The switch which samples the signals is called the commutator. The switching mechanism which performs the function of switch at the receiver is called decommutator.

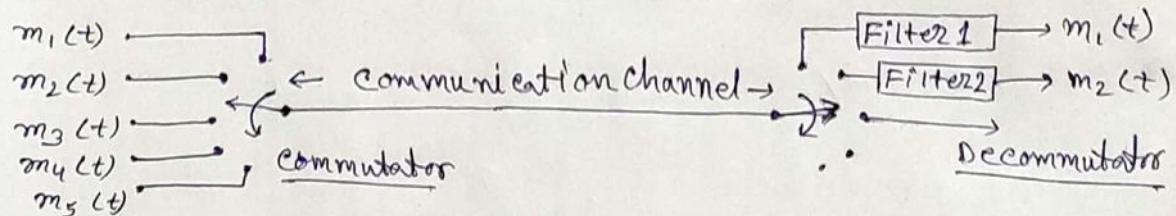


Fig - 4.8

Illustration of sampling

The commutator samples and combines samples while decommutator generates samples belonging to individual signal so that these signals may be reconstructed.

The interlacing of signals that allows multiplexing is shown in fig 4.9. For simplicity, multiplexing of two signals $m_1(t)$ and $m_2(t)$ are considered. The signal $m_1(t)$ is sampled regularly at intervals of T_s and at the times indicated in figure. The sampling of $m_2(t)$ is similarly regular, but the samples are taken at a time different from the sampling time of $m_1(t)$. We shall see how many signals may be multiplexed.

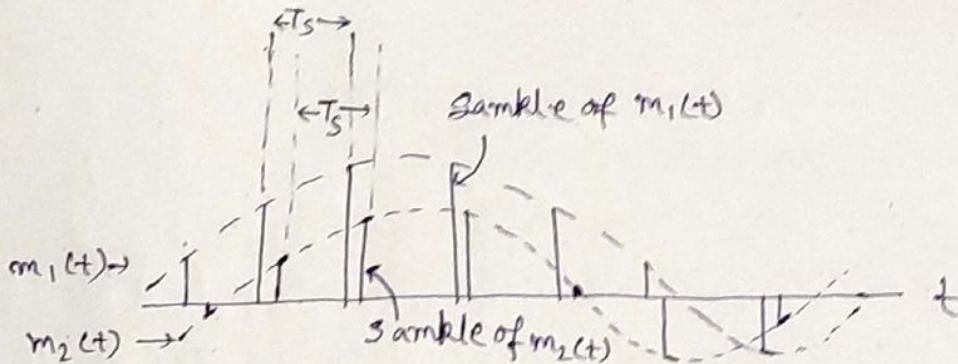


Fig 4.9 The interlacing of two baseband signals

We observe that the train of pulses corresponding to the samples of each signal are modulated in amplitude in accordance with the signal itself. Accordingly the scheme of sampling is called pulse-amplitude modulation and called PAM.

Multiplexing of several PAM signals is possible because the various signals are kept distinct and are separately recoverable by virtue of the fact that they are sampled at different times. Hence the system is an example of time-division multiplex (TDM) system.

If the multiplexed signals are to be transmitted directly, say, over a pair of wires, no further signal processing need be undertaken. Suppose we require to transmit the TDM-PAM signal from one antenna to another. It would then be necessary to amplitude modulate or frequency modulate a high frequency carrier with the TDM-PAM signal; in such a case the overall system would be referred to, respectively, as PAM-AM or PAM-FM.

Natural Sampling:-

Instantaneous sampling is hardly feasible. Instantaneous samples at the transmitting end of the channel have infinitesimal energy, and when transmitted through a band limited channel give rise to signals having a peak value which is infinitesimally small. Such infinitesimal signals will inevitably be lost in background noise.

A much more reasonable manner of sampling referred to as natural sampling, shown in Fig. 4.11.

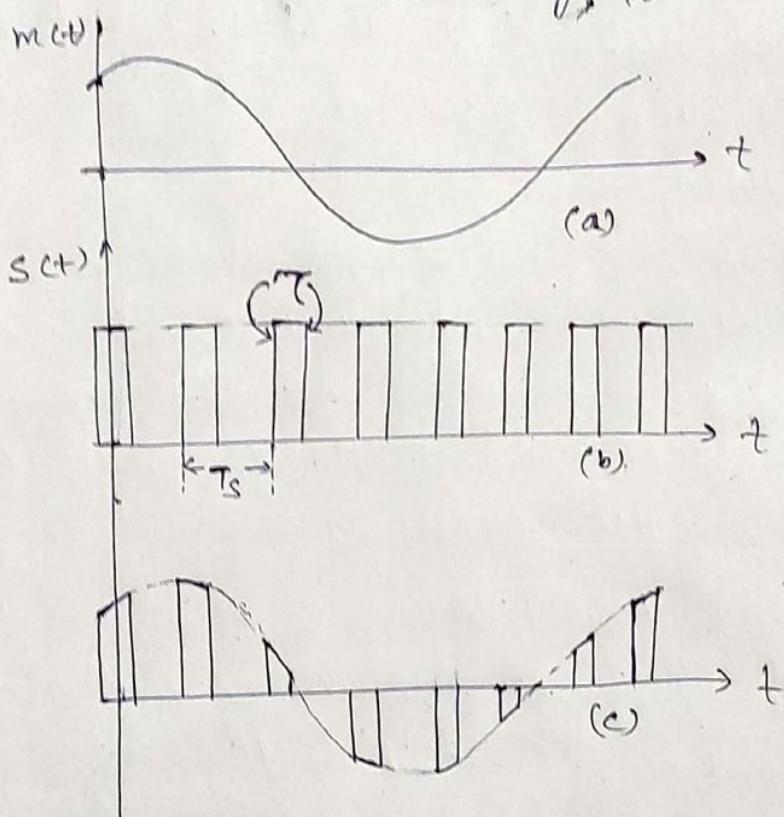


Fig-4.11 (a) baseband signal $m(t)$ (b) sampling signal $s(t)$ with finite duration T (c) naturally sampled signal $s(t)m(t)$

Here the sampling waveform $s(t)$ consists of a train of pulses having duration T and separated by a sampling time T_s . The baseband signal is $m(t)$, and the sampled signal $s(t)m(t)$ is shown in Fig. 4.11C. The sampled signal consists of a sequence of pulses of varying amplitude whose tops are not

flat but follow the waveform of the signal $m(t)$.

A signal sampled at the Nyquist rate may be reconstructed exactly by passing the samples through an ideal lowpass filter with cutoff at the frequency f_m , where f_m is the highest frequency spectral component of the signal.

The Sampling waveform $s(t)$ is given by

$$s(t) = \frac{2}{T_s} + \frac{2}{T_s} \left(C_1 \cos 2\pi \frac{t}{T_s} + C_2 \cos 2\pi 2\frac{t}{T_s} + \dots \right) \quad \dots \textcircled{i}$$

where, $C_n = \frac{\sin(n\pi t/T_s)}{n\pi t/T_s}$

The Sampled baseband signal $s(t)m(t)$ is, for $T_s = \frac{1}{2f_m}$,

$$s(t)m(t) = \frac{2}{T_s} m(t) \left\{ + \frac{2}{T_s} \left[m(t)C_1 \cos 2\pi(2f_m)t + m(t)C_2 \cos 2\pi(4f_m)t + \dots \right] \right\} \dots \textcircled{iii}$$

So, as in instantaneous sampling, a lowpass filter with cutoff at f_m will deliver an output signal $s_o(t)$

given by $s_o(t) = \frac{2}{T_s} m(t)$

With samples of finite duration, it is not possible to completely eliminate crosstalk generated in a channel, sharply bandlimited to a bandwidth h_f . If N signals are to be multiplexed, then the maximum sample duration is $\tau = \frac{T_s}{N}$. It is advantageous for the purpose of increasing the level of the output signal, to make τ as large as possible. $s_o(t)$ increases with τ . To help suppress crosstalk, it is ordinarily required that the samples be limited to a duration much less than $\frac{T_s}{N}$. The result is a large ~~gap~~

Guard time between the end of one sample and the beginning of the next.

Flat-Top Sampling:

Pulses of the type shown in fig. 4.11, with tops contoured to follow the waveform of the signal, are actually not frequently employed. Instead flat-topped pulses are customarily used, as shown in fig. 4.12a. A flat-topped pulse has a constant amplitude established by the sample value of the signal at some point within the pulse interval. In sampling of this type of the baseband signal $m(t)$ cannot be recovered exactly by simply passing the samples through an ideal low-pass filter. However the distortion need not be large. Flat-top sampling simplifies the design of the electronic circuitry.

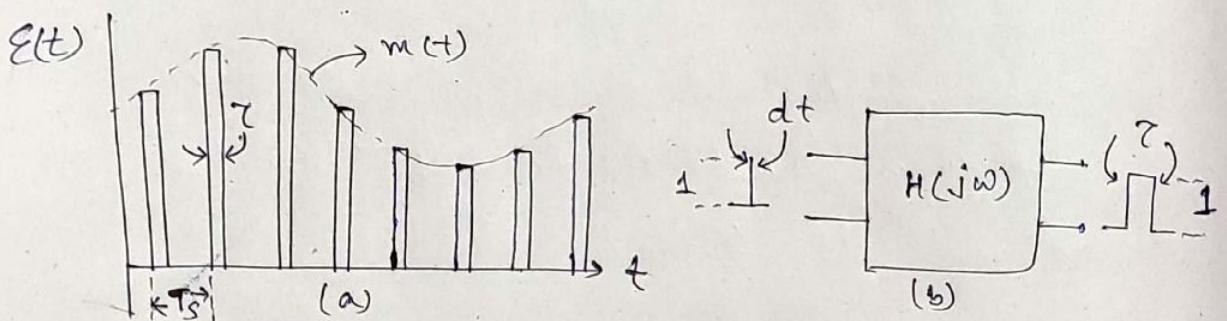


Fig-4.12 (a) Flat topped sampling

(b) Network ~~of~~ with transform $H(j\omega)$ which converts a pulse of width dt into a pulse of duration τ .

To show the extent of distortion, consider the signal $m(t)$ having a Fourier transform $M(j\omega)$. Flat-top pulse can be generated by passing the instantaneously sampled signal through a network that broadens a pulse of duration dt (an impulse) into a pulse of duration τ . The Fourier transform of a pulse of unit

Signal Recovery through Holding:

amplitude and width $d\tau$ is -

$$\mathcal{F}[\text{impulse of strength } d\tau \text{ at } t=0] = d\tau \quad \dots \textcircled{i}$$

Fourier transform of unit amplitude and width τ is

$$\mathcal{F}[\text{pulse of amp. 1, extending from } t = -\frac{\tau}{2} \text{ to } +\frac{\tau}{2}]$$

$$= \tau \frac{\sin(\omega \tau/2)}{(\omega \tau/2)} \quad \dots \textcircled{ii}$$

The transfer function of the network shown below is

$$H(j\omega) = \frac{\tau}{d\tau} \frac{\sin(\omega \tau/2)}{(\omega \tau/2)} \quad \dots \textcircled{iii}$$

Let the signal $m(t)$ with transform $M(j\omega)$, be bandlimited to f_m and be sampled at the Nyquist rate or faster. In the range 0 to f_m the transform of flat-topped signal is given by $H(j\omega) M(j\omega)$

$$\mathcal{F}[\text{flat-topped sampled } m(t)] = \frac{\tau}{T_s} \frac{\sin(\omega \tau/2)}{(\omega \tau/2)} M(j\omega) \quad 0 \leq f \leq f_m \quad \dots \textcircled{iv}$$

To illustrate the effect, we consider the signal $m(t)$ has a flat spectral density equal to M_0 over its entire range from 0 to f_m as shown in fig 4.13a. The sampling frequency $f_s = \frac{1}{T_s}$ is assumed large enough to allow for a guardband between the spectrum of baseband signal and the DSB-SC signal with carrier f_s . The spectrum of flat-topped sampled signal is shown in 4.13d. If in the range 0 to f_m the spectra of sampled signal and

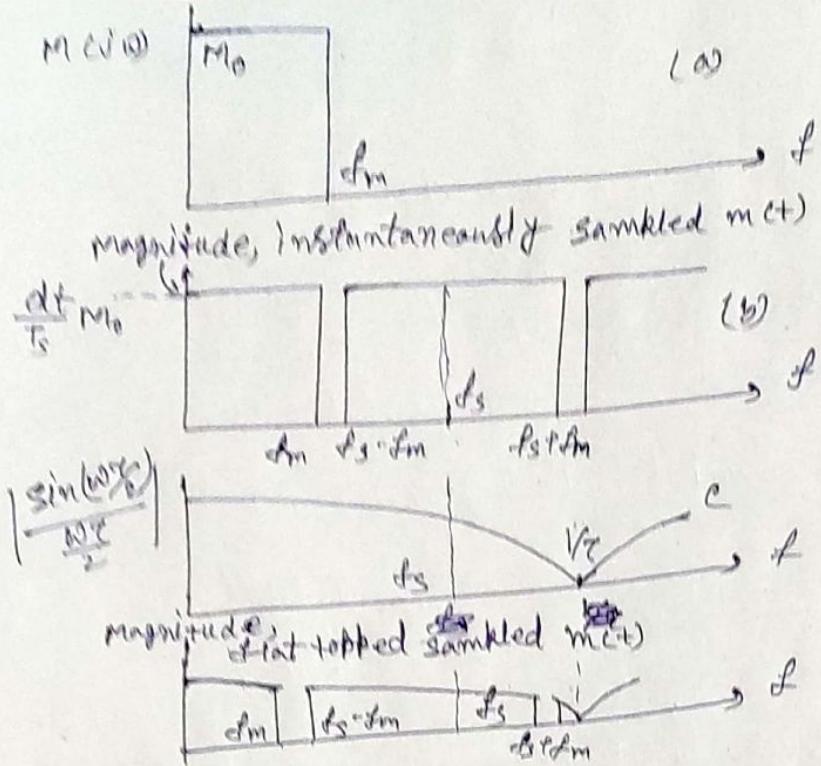


Fig-4.13
original stored are identical, original signal may be recovered by a low pass filter. Since, they are not identical, distortion will result. The distortion occurs since, the original signal was 'observed' through a finite rather than an infinitesimal time 'aperture' and hence the distortion is known as aperture effect distortion.

The distortion results as the spectrum is multiplied by the sampling function $Sa(x) = \frac{\sin x}{x}$ (with $x = \frac{\omega t}{2}$). The magnitude of sampling function with increasing x in the neighbourhood of $x=0$ and does not fall off sharply until we approach $x=\pi$, at which $Sa(x)=0$. To minimize this distortion, corresponding to $x=\pi$ must be very large compared to fm . Since, $x=\pi f T$, the frequency corresponding to $x=\pi$ is $f_0 = \frac{1}{T}$. If $f_0 \gg fm$, i.e. $T \ll \frac{1}{fm}$ the aperture distortion will be small.

Equalization:

It is advantageous to make τ as large as possible, practically for the sake of increasing amplitude of the output signal. If in a particular case, it should happen that the consequent distortion is not acceptable and may be corrected by including an equalizer in cascade with the output low pass filter. An equalizer is a passive network whose transfer function has a frequency dependence of the form $\frac{1}{\sin \omega}$.

Signal Recovery Through Holding:

Maximum ratio τ/T_s , of the sample duration to the sampling interval is $1/N$, N being the number of signals to be multiplexed. We discuss now an alternative method of recovery of the baseband signal which raises the level of the output signal. The method has additional advantage that that some distortion must be accepted. The method is illustrated in fig 4.14 below where the baseband signal $m(t)$ and its flat-topped samples are shown.

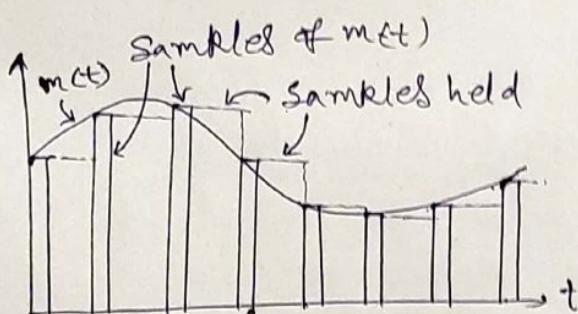


Fig-4.14 Illustrating the operation of holding

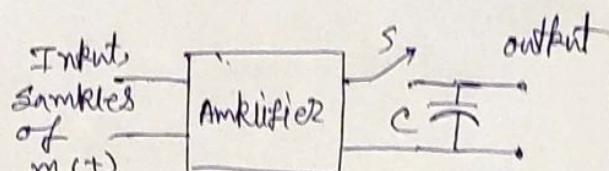


Fig-4.15 Illustrating a method of performing the operation of holding

At the receiving end and after demultiplexing, the sample pulses are extended, i.e. the sample value of each individual baseband signal is held until the occurrence of the next sample of that same baseband signal. The output waveform is then an up and down staircase without blank intervals.

A method of performing the holding operation is shown in fig 4.15. The switch S operates in synchronism with the occurrence of input samples. The switch ordinarily open, closes after the leading edge of a sample pulse and opens before the trailing edge. The amplifier has a low output impedance. Hence at the closing of the switch, capacitor C charges abruptly to a voltage proportional to the sample value and the capacitor holds this voltage until the operation is repeated for the next sample. In practical situation, the capacitor voltage is not flat, it charges or discharges exponentially. If the received sampled pulses are natural samples rather than flat-topped, there is a ~~voltage~~ departure from constant voltage within the sampling time. In practical, sampling time is & very small than sampling interval and the departure is small enough to be neglected.

Pulse Width Modulation and Pulse Position

Modulation :- (PWM and PPM)

In Pulse Width Modulation (PWM) the message modulates the width of the pulse and in Pulse Position Modulation (PPM), the position of the arrival of a fixed width pulse in each sample period is modulated by the message signal. These are not suitable for time division multiplexing and contribution is limited. PWM finds application in motor control, in delivery of power. Together PWM and PPM are known as Pulse Time Modulation or PTM. The modulation and de-modulation of these two are closely connected.

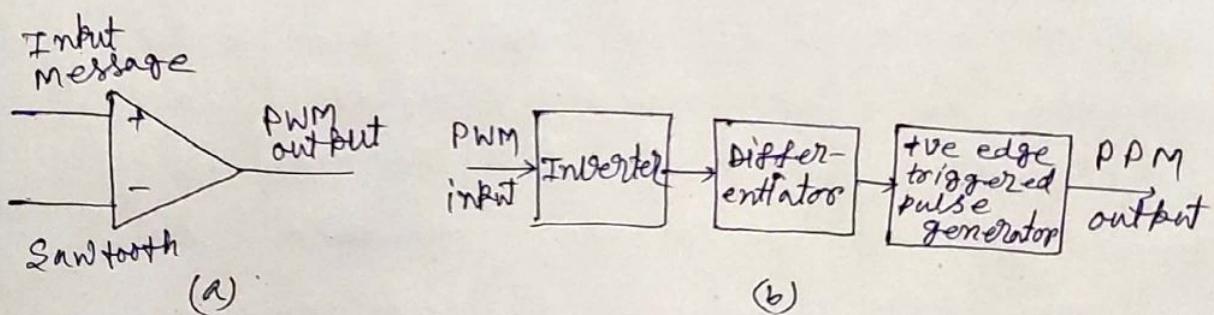


Fig - 4.17 (a) PWM generation by comparator

(b) PPM generation from PWM

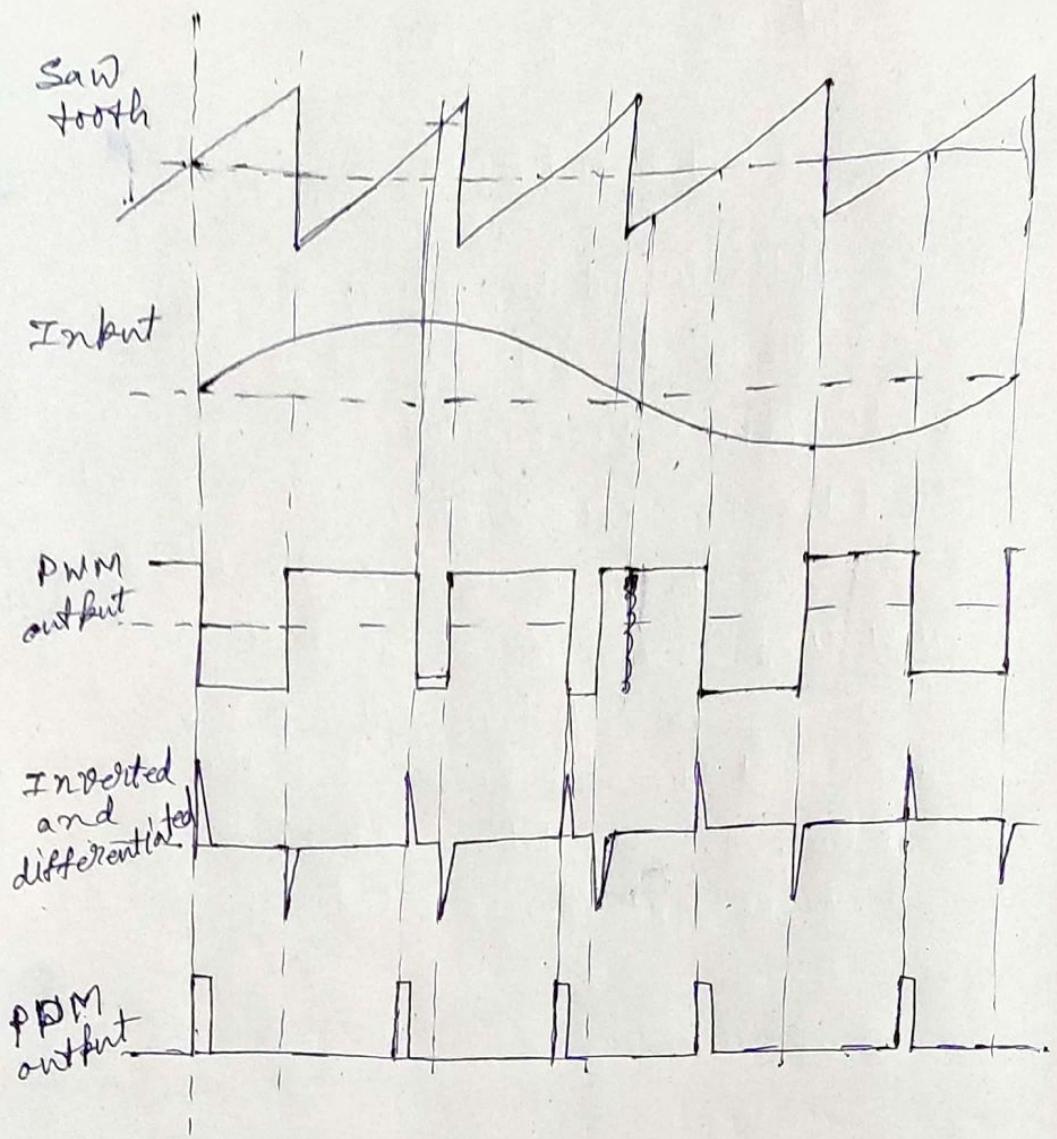


Fig - 4.18 principle of PWM and PPM generation

Let us begin with the simplest method for PWM generation as shown in Fig 4.17a. We have a comparator, one input of which is fed by input message signal and the other by a sawtooth signal (Fig. 4.18). We will get a PWM signal at the output. The rising edge of PWM coincide with the falling edge of the sawtooth signal.

The PPM generation is a post processing of PWM as shown in Fig 4.17b. The PWM signal generated is sent to an inverter which reverses the polarity of

the pulses. If it is followed by a differentiator we'll have +ve spikes where original PWM signal was going High to Low and -ve spikes where Low to High. These spikes are then fed to a +ve edge triggered fixed width pulse generator as shown in figure. They generate pulses of fixed width when +ve spikes occur. These are PPM outputs where positions of the pulses in a sample period carry input message information. This method is known as direct method. The other direct method uses addition of message signal with sawtooth like signal (the signal first shows a step and then linear fall) and setting a Comparator level in such a manner that it always cuts in that triangular region in every cycle. In indirect method a PAM signal from input message is generated to which this sawtooth like signal is added and then the Comparator does the slicing. The more the message amplitude, the cut is more towards the base of the triangle and the more is the pulselength and vice-versa.

If we want to convert PPM to PWM we go through the fig. 4.19 shown below.

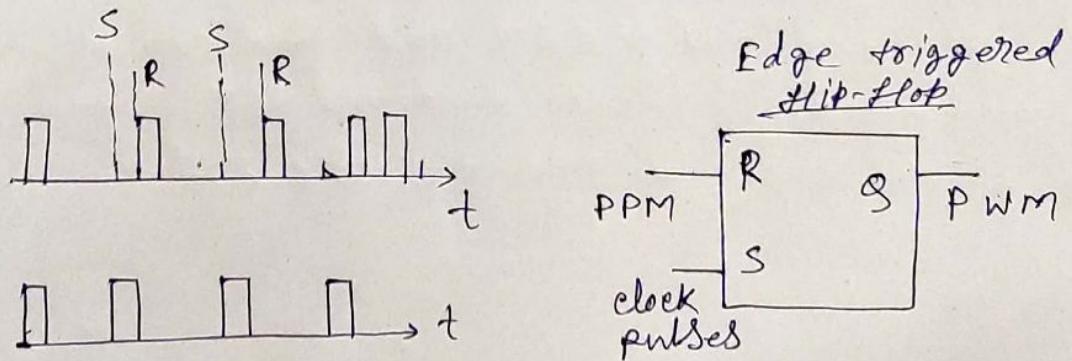


Fig-4.19 PWM from PPM

The SR edge triggered flip flop is set by the edge of the clock. It remains set so that q is high, till a +ve edge from PPM resets it. The more the delay in arrival, the longer the duration q remains high. It again gets set in the next clock pulse by the falling edge of the clock pulse. Thus the output of the flip flop is a train of pulses, the width of which is decided by how late PPM pulses arrive in a particular clock period in which again the message information is contained. Thus we get a PWM output at the flip flop output.

To discuss demodulation of PWM and PPM, first for PWM demodulation, we start a ramp up at the +ve edge and stop it when the -ve edge comes. Since the widths are different, the ramps reach different highest heights in each cycle which is directly proportional to the pulse width and in turn the amplitude of the modulating message. This when passed through a low pass filter will follow the envelope, ie. the demodulation is done. Transistor and RC combination can be used both for ramp generation and filtering to implement a demodulator circuit.

Between PWM

If a synchronous clock is available, a PPM can be converted to PWM and then can be demodulated. Between PWM and PPM, the later gives better performance in a noisy system.