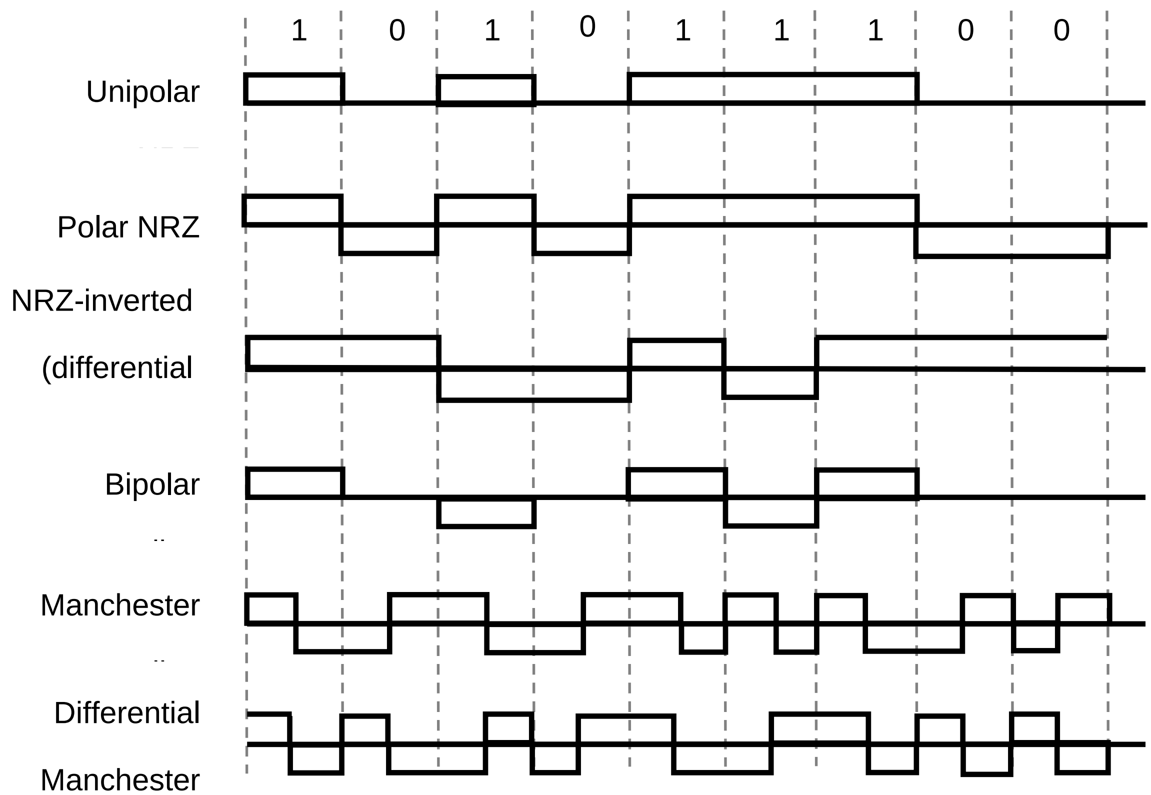
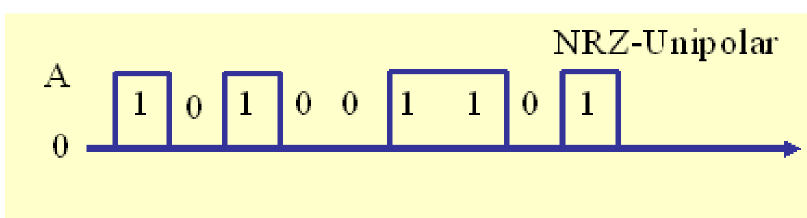
CHAPTER-4
Line Codes

In base band transmission best way is to map digits or symbols into pulse waveform. This waveform is generally termed as Line codes.

RZ: Return to Zero [pulse for half the duration of T_b]
NRZ Return to Zero[pulse for full duration of T_b]



Unipolar (NRZ)



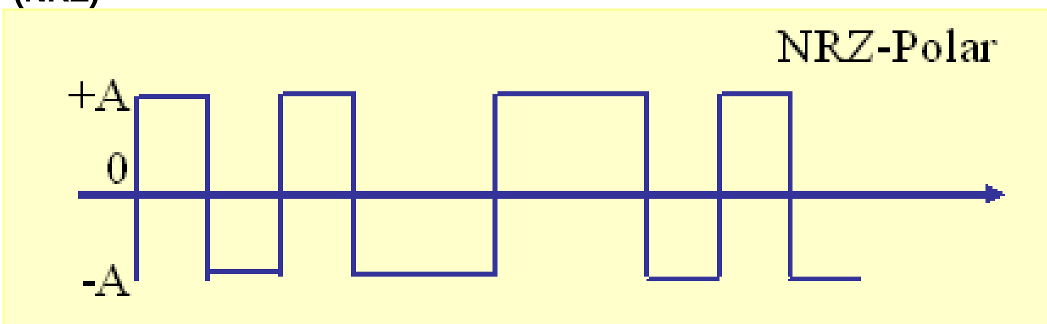
Unipolar NRZ

Unipolar NRZ

“1” maps to +A pulse “0” maps to no pulse

- Poor timing
- Low-frequency content
- Simple
- Long strings of 1s and 0s ,synchronization problem

Polar - (NRZ)

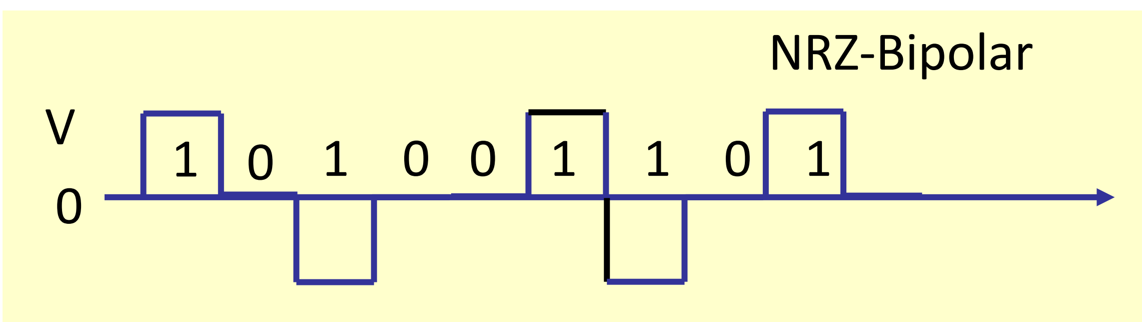


Polar NRZ

“1” maps to +A pulse “0” to –A pulse

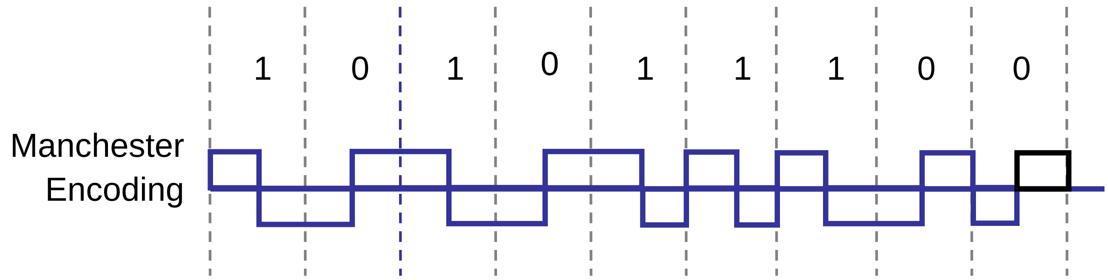
- Better Average Power
- simple to implement
- Long strings of 1s and 0s ,synchronization problem
- Poor timing

Bipolar Code



- Three signal levels: {-A, 0, +A}
- “1” maps to +A or –A in alternation
- “0” maps to no pulse
- Long string of 0’s causes receiver to loose synchronization
- Suitable for telephone systems.

Manchester code



- "1" maps into A/2 first for $T_b/2$, and -A/2 for next $T_b/2$
- "0" maps into -A/2 first for $T_b/2$, and A/2 for $T_b/2$
- Every interval has transition in middle
  - Timing recovery easy
- Simple to implement
- Suitable for satellite telemetry and optical communications

Differential encoding

- It starts with one initial bit .Assume 0 or 1.
- Signal transitions are used for encoding.

Example NRZ - S and NRZ – M

- NRZ –S : symbol 1 by no transition , Symbol 0 by transition.
- NRZ-M : symbol 0 by no transition , Symbol 1 by transition
- Suitable for Magnetic recording systems.

M-ary formats

Bandwidth can be properly utilized by employing M-ary formats. Here grouping of bits is done to form symbols and each symbol is assigned some level.

Example

Polar quaternary format employs four distinct symbols formed by dibits.
Gray and natural codes are employed

Parameters in choosing formats

1. **Ruggedness**
2. **DC Component**
3. **Self Synchronization.**
4. **Error detection**
5. **Bandwidth utilization**
6. **Matched Power Spectrum**

Power Spectra of Discrete PAM Signals:

The discrete PAM signals can be represented by random process

$$X(t)=\sum_{K=-\infty}^{\infty} A_k V(t-KT)$$

Where A_k is discrete random variable, V(t) is basic pulse, T is symbol duration. V(t) normalized so that V(0) = 1.

Coefficient A_k represents amplitude value and takes values for different line codes as

Unipolar $A_k = [$ Symbol 1 = a, Symbol 0 = 0

Polar $A_k = [$ Symbol 1 = +a, Symbol 0 = −a

Bipolar $A_k = [$ Alternate Symbol 1 takes = +a, −a; Symbol 0 = 0

Manchester $A_k = [$ Symbol 1 = a, Symbol 0 = −a

As A_k is discrete random variable, generated by random process X(t),

We can characterize random variable by its ensemble averaged auto correlation function given by

$R_A(n) = E\,[A_k.A_{k-n}]$,

A_k, A_{k-n} = amplitudes of k^{th} and $(k-n)^{th}$ symbol position

PSD & **auto correlation function** form **Fourier Transform pair** & hence auto correlation function tells us something about bandwidth requirement in frequency domain.

Hence PSD $S_x(f)$ of discrete PAM signal X(t).is given by

$$S_X(f)=\frac{1}{T}\left|V(f)\right|^2 \sum_{n=-\infty}^{\infty} R_A(n)\, e^{-j2\pi fnT}$$

Where V(f) is Fourier Transform of basic pulse V(t). V(f) & $R_A(n)$ depends on different line codes.

Power Spectra of NRZ Unipolar Format

Consider unipolar form with symbol 1's and 0's with equal probability i.e.

$P(A_k=0) = ½$ and $P(A_k=1) = ½$

For n=0;

Probable values of $A_k.A_k$ = 0 x 0 & a x a

$=E[A_k.A_{k-0}]$

$= E[A_k^2] = 0^2 \times P[A_k=0] + a^2 \times P[A_k=1]$

$R_A(0) = a^2/2$

If n ≠ 0

$A_k.A_{k-n}$ will have four possibilities (adjacent bits)

0 x 0, 0 x a, a x 0, a x a with probabilities ¼ each.

$E[A_k.A_{k-n}] = 0 \times ¼ + 0 \times ¼ + 0 \times ¼ + a^2 / 4$

$= a^2 / 4$

V(t) is rectangular pulse of unit amplitude, its Fourier Transform will be sinc function.

$V(f) = FT [V(t)] = T_b \, Sinc(fT_b)$ PSD is given by

$$S_X(f) = \frac{1}{T}\left|V(f)\right|^2 \sum_{n=-\infty}^{\infty} R_A(n) e^{-j2\pi fnT}$$

substituting the values of V(f) and $R_A(n)$

$$S_X(f) = \frac{1}{T_b}\left[T_b^2 Sinc^2(fT_b)\right] \sum_{n=-\infty}^{\infty} R_A(n) e^{-j2\pi fnT_b}$$

$$= \left[T_b Sinc^2(fT_b)\right]\left[R_A(0) + \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} R_A(n) e^{-j2\pi fnT_b}\right]$$

$$= \left[T_b Sinc^2(fT_b)\right]\left[\frac{a^2}{2} + \frac{a^2}{4} \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} e^{-j2\pi fnT_b}\right]$$

$$= \frac{a^2}{4} T_b \, Sinc^2(fT_b) + \frac{a^2}{4} T_b \, Sinc^2(fT_b) \sum_{n=-\infty}^{\infty} e^{-j2\pi \, fn \, T_b}$$

using Poisson's formula

$$\sum_{n=-\infty}^{\infty} e^{-j2\pi \, fn \, T_b} = \frac{1}{T_b} \sum_{n=-\infty}^{\infty} \delta(f - \frac{n}{T_b})$$

$$S_X(f) = \frac{a^2}{4} T_b \, Sinc^2(fT_b) + \frac{a^2}{4} T_b \, Sinc^2(fT_b) \frac{1}{T_b} \sum_{n=-\infty}^{\infty} \delta(f - \frac{n}{T_b})$$

$\sum_{n=-\infty}^{\infty} \delta(f - \frac{n}{T_b})$ is Dirac delta train which multiplies Sinc function which

has nulls at $\pm\frac{1}{T_b}, \pm\frac{2}{T_b}$ . . . .

As a result, $Sin^2(fT_b) . \sum_{n=-\infty}^{\infty} \delta(f - \frac{n}{T_b}) = \delta(f)$

Where $\delta(f)$ is delta function at f = 0,

Therefore

$$S_X(f) = \frac{a^2 T_b}{4} Sin^2(fT_b) + \frac{a^2}{4} \delta(f)$$

Power Spectra of Bipolar Format

Here **symbol 1** has levels ± a, and **symbol 0** as 0. Totally three levels.
Let 1's and 0's occur with equal probability then

$P(A_K = a) = 1/4$ For Symbol 1
$P(A_K = -a) = 1/4$
$P(A_K = 0) = 1/2$ For Symbol 0

For n=0

$E[A_K^2] = a \times a \, P(A_K = a) + (0 \times 0) \, P[A_K = 0] +$
$(-a \times -a) \, P(A_K = -a)$
$= a^2/4 + 0 + a^2/4 = a^2/2$

For n≠0, i.e. say n=1;

Four possible forms of $A_K.A_{K-1}$

00,01,10,11 i.e. dibits are

0 X 0, 0 X ±a, ±a X 0, ±a X ±a

with equal probabilities ¼.

$E[A_K.A_{K-1}]$ = 0 x ¼ + 0 x ¼ + 0 x ¼ - a^2 x ¼

= $-a^2/4$

For n>1, 3 bits representation 000,001,010 . . . . . . 111. i.e. with each probability of 1/8 which results in

$E[A_K.A_K\text{-}n] = 0$

$$\text{Therefore } R_A(n) = \begin{cases} a2/2 & n = 0 \\ -a2/4 & n = \pm 1 \\ 0 & n > 1 \end{cases}$$

$$S_X(f) = \frac{1}{T}|V(f)|^2 \sum_{n=-\infty}^{\infty} R_A(n) e^{-j2\pi fnT}$$

PSD is given by

$$S_X(f) = \frac{1}{T_b}\left[T_b{}^2 \mathrm{SinC}^2(fT_b)\right]\left[R_A(-1)e^{j2\pi fnTb}\right] + R_A(0) + \left[R_A(1)e^{-j2\pi fnTb}\right]$$

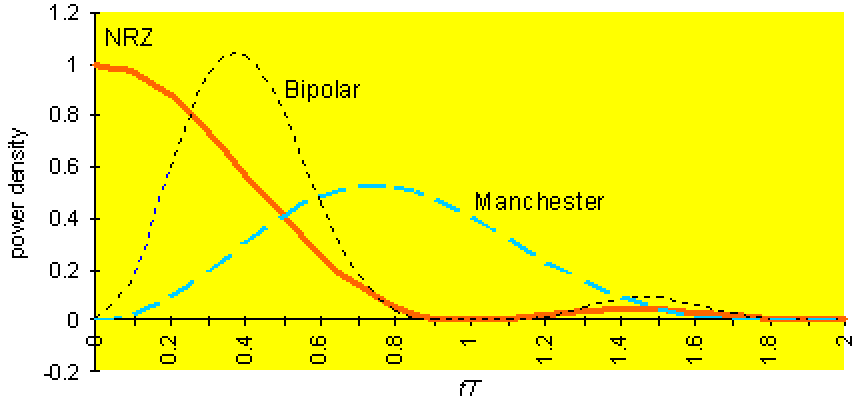
$$S_X(f) = \left[T_b \mathrm{SinC}^2(fT_b)\right]\left[\frac{a^2}{2} - \frac{a^2}{4}(e^{j2\pi fnTb} + e^{-j2\pi fTb})\right]$$

$$S_X(f) = \left[\frac{a^2 T_b}{2}\mathrm{SinC}^2(fT_b)\right]\left[1 - \mathrm{Cos}(2\pi fT_b)\right]$$

$$S_X(f) = \left[\frac{a^2 T_b}{2}\mathrm{SinC}^2(fT_b)\right]\left[2\mathrm{Sin}^2(fTb)\right]$$

$$S_X(f) = \left[a^2 T_b \mathrm{SinC}^2(fT_b)\right]\left[\mathrm{Sin}^2(fTb)\right]$$

Spectrum of Line codes



- Unipolar most of signal power is centered around origin and there is waste of power due to DC component that is present.
- Polar format most of signal power is centered around origin and they are simple to implement.
- Bipolar format does not have DC component and does not demand more bandwidth, but power requirement is double than other formats.
- Manchester format does not have DC component but provides proper clocking.

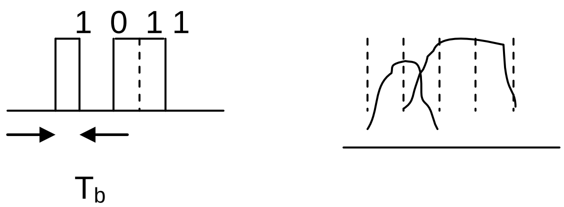
Spectrum suited to the channel.

- The PSD of the transmitted signal should be compatible with the channel frequency response
- Many channels cannot pass dc (zero frequency) owing to ac coupling
- Low pass response limits the ability to carry high frequencies

Inter symbol Interference

Generally, digital data is represented by electrical pulse, communication channel is always band limited. Such a channel disperses or spreads a pulse carrying digitized samples passing through it. When the channel bandwidth is greater than bandwidth of pulse, spreading of pulse is very less. But when channel bandwidth is close to signal bandwidth, i.e. if we transmit digital data which demands more bandwidth which exceeds channel bandwidth, spreading will occur and cause signal pulses to overlap. This overlapping is called **Inter Symbol Interference**. In short it is called ISI. Similar to interference caused by other sources, ISI causes degradations of signal if left uncontrolled. This problem of ISI exists strongly in Telephone channels like coaxial cables and optical fibers.

In this chapter main objective is to study the effect of ISI, when digital data is transmitted through band limited channel and solution to overcome the degradation of waveform by properly shaping pulse.



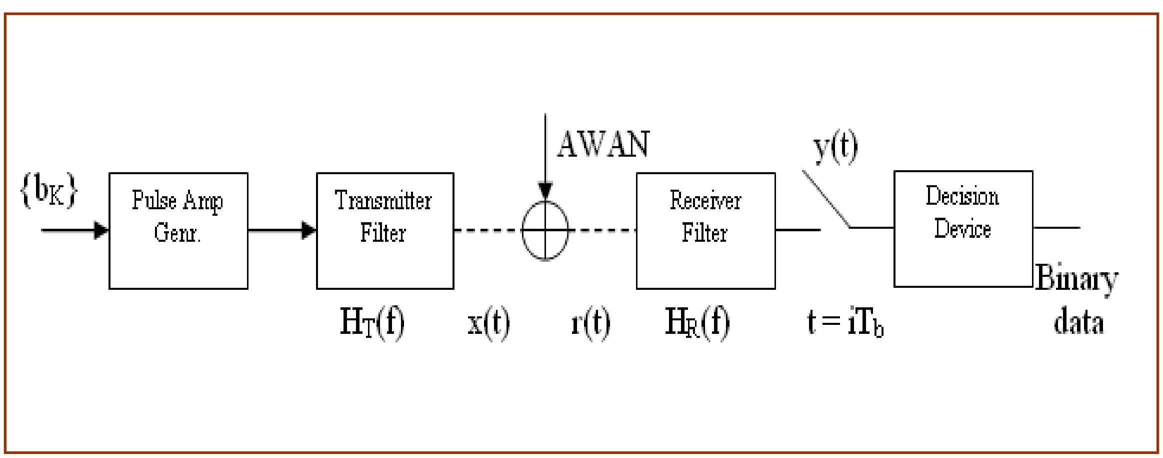
Transmitted Waveform   Pulse Dispersion

The effect of sequence of pulses transmitted through channel is shown in fig. The

Spreading of pulse is greater than symbol duration, as a result adjacent pulses interfere. i.e. pulses get completely smeared, tail of smeared pulse enter into adjacent symbol intervals making it difficult to decide actual transmitted pulse.

First let us have look at different formats of transmitting digital data.In base band transmission best way is to map digits or symbols into pulse waveform. This waveform is generally termed as **Line codes**.

BASEBAND TRANSMISSION:



PAM signal transmitted is given by

$$x(t) = \sum_{K=-\infty}^{\infty} a_K V(t - KT_b) \qquad \text{-------------------------- (1)}$$

V(t) is basic pulse, normalized so that V(0) = 1,

x(t) represents realization of random process X(t) and a_k is sample value of random variable a_k which depends on type of line codes.

The receiving filter output

$$y(t) = \mu \sum_{K=-\infty}^{\infty} a_k P(t - KT_b) \qquad \text{---------------(2)}$$

The output pulse μ P(t) is obtained because input signal **a_k .V**(t) is passed through series of systems with transfer functions $H_T(f)$, $H_C(f)$, $H_R(f)$

Therefore μ P(f) = V(f). $H_T(f).H_C(f).H_R(f)$ --------- (3)

P(f) $\rightleftharpoons$ p(t) and V(f) $\rightleftharpoons$ v(t)

The receiving filter output y(t) is sampled at **t_i** = **iT_b.** where 'i' takes intervals i = ±1, ±2 . . . . .

$$y(iT_b) = \mu \sum_{K=-\infty}^{\infty} a_k P(iT_b - KT_b)$$

$$y(iT_b) = \mu\, a_i P(0) + \mu \sum_{K=-\infty}^{\infty} a_k P(iT_b - KT_b) \qquad \text{---------------------(4)}$$

K = i K≠i

In equation(4) first term μa_i represents the output due to i^{th} transmitted bit. Second term represents residual effect of all other transmitted bits that are obtained while decoding i^{th} bit. This unwanted residual effect indicates ISI. This is due to the fact that when pulse of short duration T_b is transmitted on band limited channel, frequency components of the pulse are differentially attenuated due to frequency response of channel causing dispersion of pulse over the interval greater than T_b.

In absence of ISI desired output would have **$y(t_i) = \mu a_i$**

Nyquist Pulse Shaping Criterion

In detection process received pulse stream is detected by sampling at intervals $\pm KT_b$, then in detection process we will get desired output. This demands sample of i^{th} transmitted pulse in pulse stream at K^{th} sampling interval should be

$$P(iT_b - KT_b) = \begin{cases} 1 & K=i \\ 0 & K \neq i \end{cases} \quad \text{------------- (5)}$$

If received pulse P(t) satisfy this condition in time domain, then

$y(t_i) = \mu a_i$

Let us look at this condition by transform eqn(5) into frequency domain.

Consider sequence of samples $\{P(nT_b)\}$ where n=0,±1. . . . . . . by sampling in time domain, we write in frequency domain

$$p_\delta(f) = \frac{1}{T_b}\sum_{n=-\infty}^{\infty} p(f - n/T_b) \text{----------------(6)}$$

Where $p_\delta(f)$ is Fourier transform of an infinite period sequence of delta functions of period T_b but $p_\delta(f)$ can be obtained from its weighted sampled $P(nT_b)$ in time domain

$$p_\delta(f) = \int_{-\infty}^{\infty} \sum_{m=-\infty}^{\infty} p(mT_b)\,\delta(t - mT_b)\,e^{-j2\pi ft}dt \; = \; p(t).\delta(t)$$

Where m = i-k, then i=k, m=0; so

$$p_\delta(f) = \int_{-\infty}^{\infty} p(0)\,\delta(t)\,e^{-j2\pi ft}dt$$

Using property of delta function

i.e $\int_{-\infty}^{\infty} \delta(t)\,dt = 1$

Therefore $p_\delta(f) = p(0) = 1$

$$P_\delta(f) = 1 \quad \text{------------(7)}$$

p(0) =1 ,i.e pulse is normalized (total area in frequency domain is unity)

Comparing (7) and (6)

$$\frac{1}{T_b}\sum_{n=-\infty}^{\infty} p(f - n/T_b) = 1$$

Or $$\sum_{n=-\infty}^{\infty} p(f - n/T_b) = T_b = \frac{1}{R_b}$$ ----------- (8)

Where R_b = Bit Rate

Is desired condition for zero ISI and it is termed Nyquist's first criterion for distortion less base band transmission. It suggests the method for constructing band limited function to overcome effect of ISI.

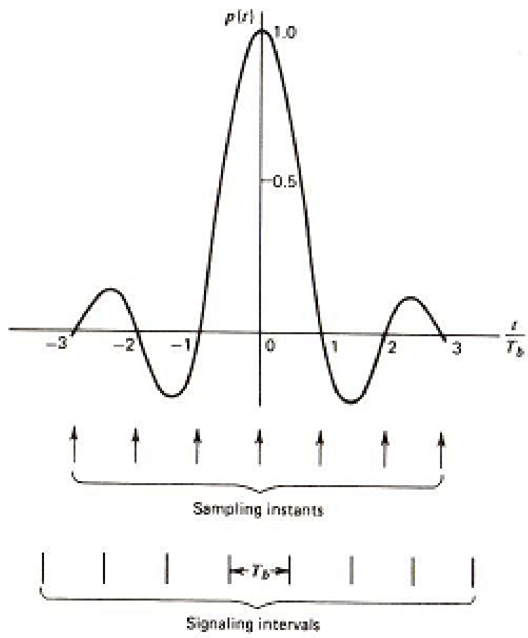
Ideal Solution

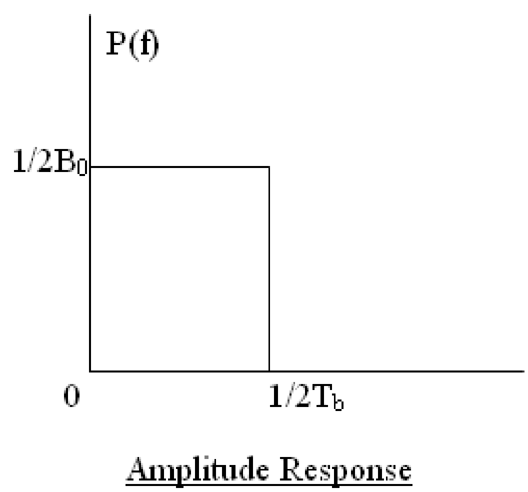
Ideal Nyquist filter that achieves best spectral efficiency and avoids ISI is designed to have bandwidth as suggested

$B_0 = 1/2T_b$ (Nyquist bandwidth) = $R_b/2$

ISI is minimized by controlling P(t) in time domain or P(f) to be rectangular function in frequency domain.

$$\mathbf{P(f) = \frac{1}{2B0}\ rect\left(\frac{f}{2B0}\right)}$$





Amplitude Response

Impulse response in time domain is given by

$$P(t) = \frac{\sin(2\pi B_0 t)}{2\pi B_0 t}$$

$$= \text{sinc}(2B_0 t)$$

Disadvantage of Ideal solution

- P(f) to be flat from $-B_o$ to $+B_o$ and zero else where , abrupt transition is physically not realizable.
- For large values of ' t' , function P(t) decreases as resulting in slower decay of sinc function due to discontinuity of P(f)

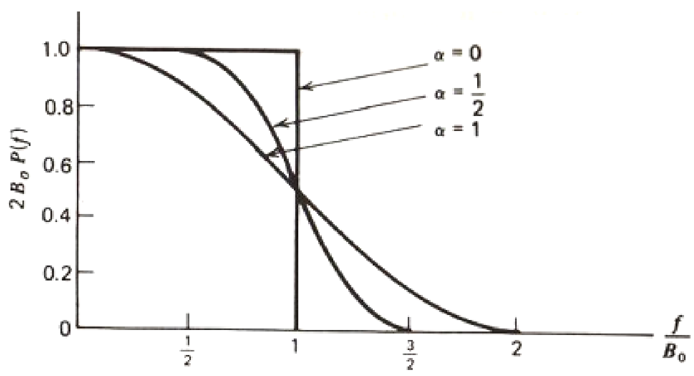
This causes timing error which results in ISI.

<u>Practical solution</u>

Raised Cosine Spectrum

- To design raised cosine filter which has transfer function consists of a flat portion and a roll off portion which is of sinusoidal form
- Bandwidth $B_0 = \frac{1}{2Tb}$ is an adjustable value between B_o and $2B_o$.

$$P(f) = \begin{cases} \frac{1}{2Bo} & |f| < f_1 \\ \frac{1}{4Bo\left\{1+\cos\left[\frac{\pi|f|-f_1}{2Bo-2f_1}\right]\right\}} & f_1 \le |f| < 2Bo - f_1 \\ 0 & |f| \ge 2Bo - f_1 \end{cases}$$



The frequency f_1 and bandwidth B_o are related by

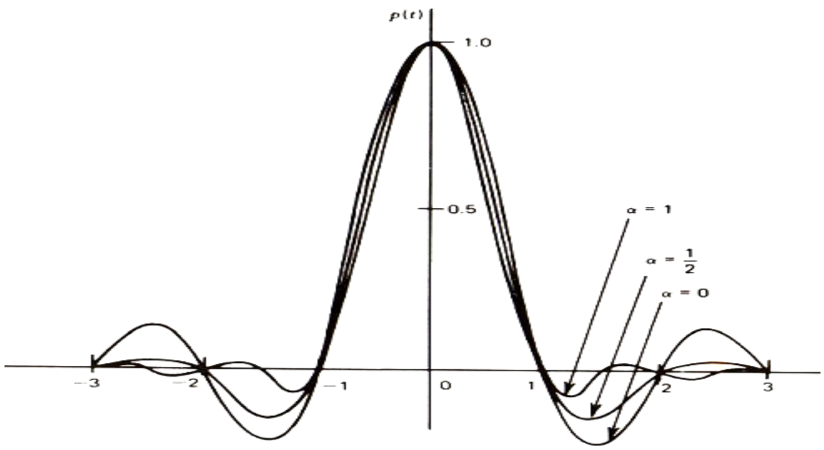
$$\alpha = 1 - \frac{f_1}{B_0}$$ **α** is called the **roll off factor**

for α = 0, $f_1=B_o$ and BW=B_o is the minimum Nyquist bandwidth for the . rectangular spectrum.

- For given Bo , roll off factor **'α'** specifies the required excess bandwidth
- **α =1,**indicates required excess bandwidth is 100% as roll off characteristics of P(f) cuts off gradually as compared with ideal low pass filters. This function is practically realizable.

Impulse response P(t) is given by

$$P(t) = \text{sinc}(2Bot)\frac{\cos(2\pi\ \alpha Bot)}{1-16\ \alpha Bo^2 t^2}$$



P(t) has two factors

- sinc(2Bot) which represents ideal filter - ensures zero crossings
- second factor that decreases as $1/|t|^2$ - helps in reducing tail of sinc pulse i.e. fast decay
- For α =1, $$P(t)=\frac{\text{sinc}(4B_0 t)}{1-16B_0^2 t^2}$$

At $t=\frac{Tb}{2}$ **p(t)=0.5**

Pulse width measured exactly equal to bit duration T_b. Zero crossings occur at t = $\pm 3T_b$, $\pm 5T_b$… In addition to usual crossings at t = ±Tb, ±2Tb… Which helps in time synchronization at receiver at the expense of double the transmission bandwidth

Transmission bandwidth required can be obtained from the relation

$B = 2B_o - f_1$

Where B = Transmission bandwidth

Bo = $\frac{1}{2Tb}$ Nyquist bandwidth

But $$\alpha = 1 - \frac{f1}{B0}$$

using

$f_1 = B_0\,(1- \alpha)$

$B = 2\,B_0 - B_0(1- \alpha)$

therefore **$B = B_0(1+ \alpha)$**

α =0; $B=B_0$, minimum band width

α =1; $B=2B_0$, sufficient bandwidth

<u>Roll-off factor</u>

Smaller roll-off factor:

- Less bandwidth, but
- Larger tails are more sensitive to timing errors

Larger roll-off factor:

- Small tails are less sensitive to timing errors, but
- Larger bandwidth

<u>Example1</u>

A certain telephone line bandwidth is 3.5Khz .calculate data rate in bps that can be transmitted if binary signaling with raised cosine pulses and roll off factor α = 0.25 is employed.

Solution:

α = 0.25 ---- roll off

B = 3.5Khz ---transmission bandwidth

B = Bo(1+ α)

$B_0 = \frac{1}{2Tb} = \frac{Rb}{2}$ Ans: R_b= 5600bps

Example2

A source outputs data at the rate of 50,000 bits/sec. The transmitter uses binary PAM with raised cosine pulse in shaping of optimum pulse width. Determine the bandwidth of the transmitted waveform. Given

a.α = 0 b.α = 0.25 c. α = 0.5 d. α = 0.75 e. α = 1

Solution

B = B0(1+ α) B0=Rb/2

a. Bandwidth = 25,000(1 + 0) = 25 kHz

b. Bandwidth = 25,000(1 + 0.25) = 31.25 kHz

c. Bandwidth = 25,000(1 + 0.5) = 37.5 kHz

d. Bandwidth = 25,000(1 + 0.75) = 43.75 kHz

e. Bandwidth = 25,000(1 + 1) = 50 kHz

Example 3

A communication channel of bandwidth 75 KHz is required to transmit binary data at a rate of 0.1Mb/s using raised cosine pulses. Determine the roll off factor α.

Rb = 0.1Mbps

B=75Khz

α = ?

B = Bo(1+ α)

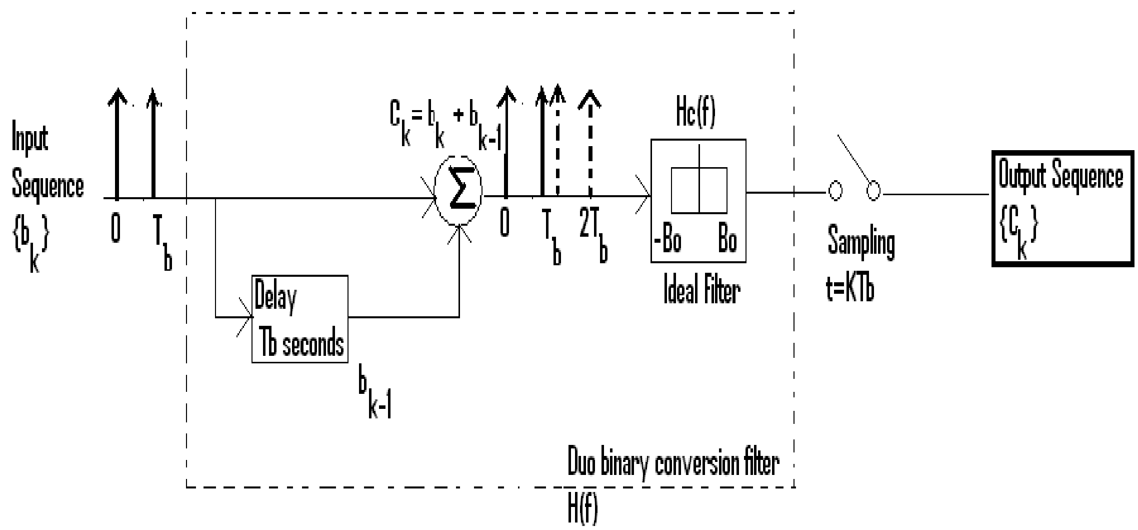
B_0 = R_b/2   **Ans : α =0.5**

Correlative coding :

So far we treated ISI as an undesirable phenomenon that produces a degradation in system performance, but by adding ISI to the transmitted signal in a controlled manner, it is possible to achieve a bit rate of 2Bo bits per second in a channel of bandwidth Bo Hz. Such a scheme is **correlative coding** or **partial- response signaling** scheme. One such example is **Duo binary signaling**.

Duo means transmission capacity of system is doubled.

Duo binary coding



Consider binary sequence $\{b_k\}$ with uncorrelated samples transmitted at the rate of R_b bps. Polar format with bit duration T_b sec is applied to duo binary conversion filter. when this sequence is applied to a duobinary encoder, it is converted into three level output, namely -2, 0 and +2.To produce this transformation we use the scheme as shown in fig.The binary sequence $\{b_k\}$ is first passed through a simple filter involving a single delay elements. For every unit impulse applied to the input of this filter, we get two unit impulses spaced T_b seconds apart at the filter output. Digit C_k at the output of the duobinary encoder is the sum of the present binary digit b_k and its previous value b_{k-1}

$$C_k = b_k + b_{k-1}$$

The correlation between the pulse amplitude C_k comes from b_k and previous b_{k-1} digit, can be thought of as introducing ISI in controlled manner., i.e., the interference in determining $\{b_k\}$ comes only from the preceding symbol $\{b_{k-1}\}$ The symbol $\{b_k\}$ takes ±1 level thus C_k takes one of three possible values -2,0,+2 . The duo binary code results in a three level output. in general, for M-ary transmission, we get ***2M-1*** levels

<u>Transfer function of Duo-binary Filter</u>

The ideal delay element used produce delay of T_b seconds for impulse will have transfer function $e^{-j 2\pi f T_b}$.

Overall transfer function of the filter H(f)

$$H(f) = H_c(f) + H_c(f) e^{-j 2\pi f T_b}$$

$$H(f) = H_c(f) \left[1 + e^{-j 2\pi f T_b} \right]$$

$$= 2H_c(f) \left[\frac{e^{j\pi f T_b} + e^{-j\pi f T_b}}{2} \right] e^{-j\pi f T_b}$$

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

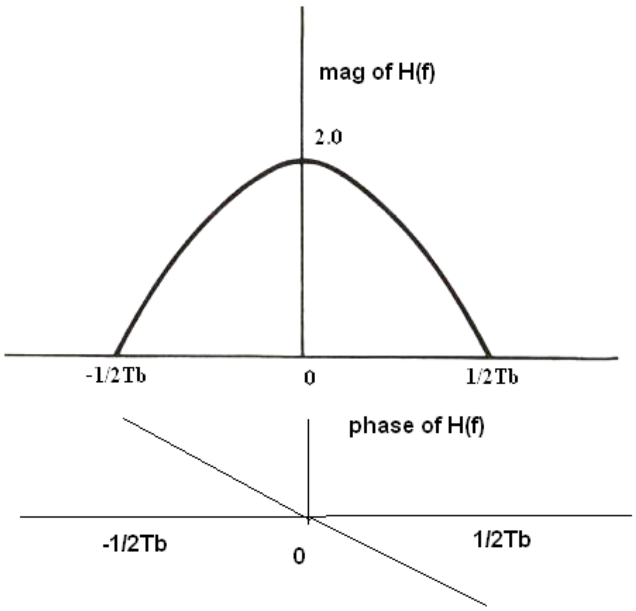
As ideal channel transfer function

$$H_c(f) = \begin{cases} 1 & |f| \le \frac{1}{2T_b} \\ 0 & \text{otherwise} \end{cases}$$

Thus overall transfer function

$$H(f) = \begin{cases} 2\cos(\pi f T_b) e^{-j\pi f T_b} & |f| \le \frac{1}{2T_b} \\ 0 & \text{otherwise} \end{cases}$$

H(f) which has a gradual roll off to the band edge, can also be implemented by practical and realizable analog filtering Fig shows Magnitude and phase plot of Transfer function



Advantage of obtaining this transfer function H(f) is that practical implementation is easy

<u>Impulse response</u>

Impulse response h(t) is obtained by taking inverse Fourier transformation of H(f)

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

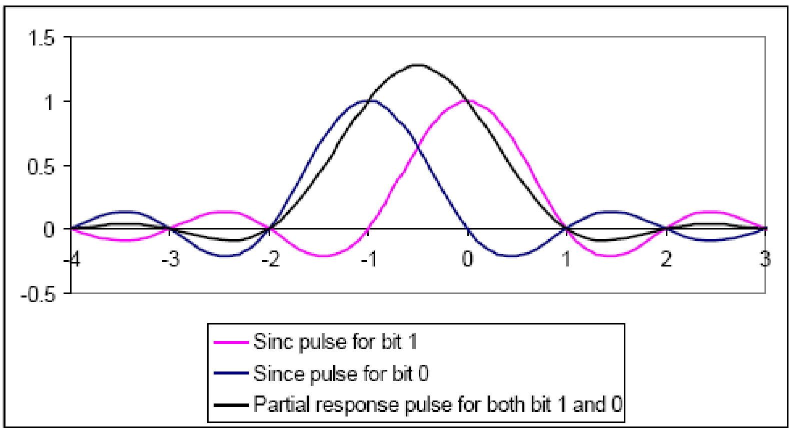
$$= \int_{-1/2T_b}^{\frac{1}{2Tb}} 2\cos(\pi f T_b) e^{-j\pi f T_b} [e^{j2\pi f t}] df$$

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

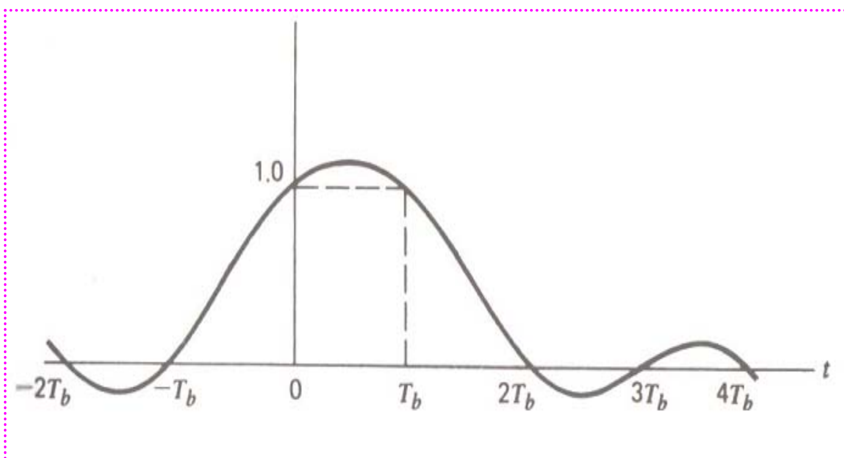
$$= \frac{\sin\left(\frac{\pi t}{T_b}\right)}{\left(\frac{\pi t}{T_b}\right)} - \frac{\sin\left[\frac{\pi t}{T_b}\right]}{\left[\frac{\pi(t - T_b)}{T_b}\right]}$$

$$h(t) = \frac{T_b{}^2 \sin\left[\frac{\pi t}{T_b}\right]}{\pi t(T_b - 1)}$$

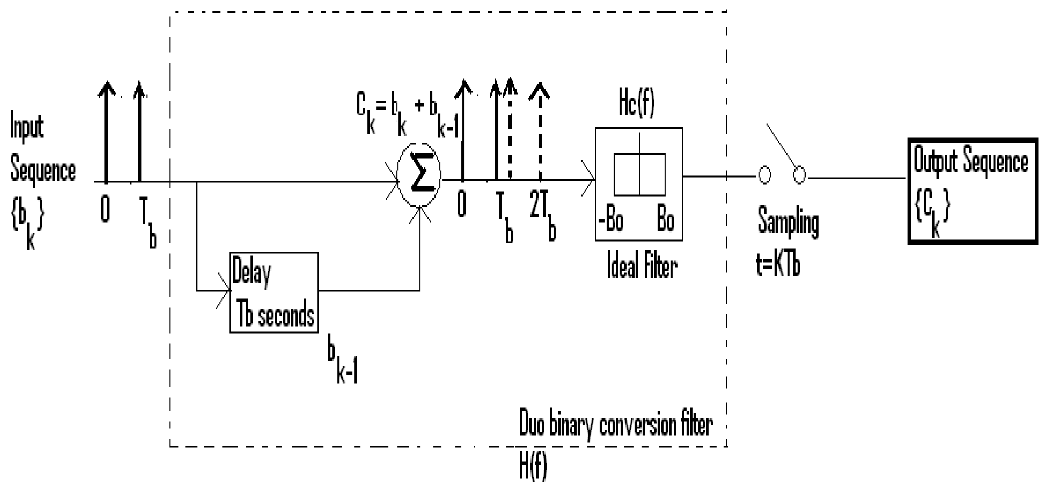
Impulse response has two sinc pulses displaced by T_b sec. Hence overall impulse response has two distinguishable values at sampling instants t = 0 and t = T_b.



Overall Impulse response



Duo binary decoding



Encoding : During encoding the encoded bits are given by

Ck = bk + bk-1

Decoding:

At the receiver original sequence $\{b_k\}$ may be detected by subtracting the previous decoded binary digit from the presently received digit C_k This demodulation technique (known as nonlinear decision feedback equalization) is essentially an inverse of the operation of the digital filter at the transmitter

if $\hat{b}_k$ is estimate of original sequence b_k then

$$\hat{b}_k = C_k - \hat{b}_{k-1}$$

Disadvantage

$$\hat{b}_{k-1}$$

If C_k and previous estimate is received properly without error then we get correct decision and current estimate Otherwise once error made it tends to propagate because of decision feed back. current $\{b_k\}$ depends on previous $b_{k-1.}$

Example consider sequence **0010110**

Transmitter							
Binary Sequence $\{b_k\}$	0	0	1	0	1	1	0
Polar Amplitudes	-1	-1	1	-1	1	1	-1
Coding Rule $c_k=b_k+b_{k-1}$ (transmitted signal)		-2	0	0	0	2	0
Receiver							
Decoding Decision Rule	If $c_k = 2$ decide that $b_k= 1$ (symbol 1) If $c_k = -2$ decide that $b_k= -1$ (symbol 0) If $c_k = 0$ decide opposite of the previous decision						
Received Sequence $\{c_k\}$		-2	0	0	0	2	0
Decoded sequence in bipolar form		-1	+1	-1	+1	1	-1
Decoded sequence $\hat{b}_k$		0	1	0	1	1	0

Precoding

In case of duo binary coding if error occurs in a single bit it reflects as multiple errors because the present decision depends on previous decision also. To make each decision independent we use a precoder at the receiver before performing duo binary operation.

The precoding operation performed on the input binary sequence $\{b_k\}$ converts it into another binary sequence $\{a_k\}$ given by

$$a_k = b_k \oplus a_{k-1}$$

a modulo 2 logical addition

Unlike the linear operation of duo binary operation, the precoding is a non linear operation.

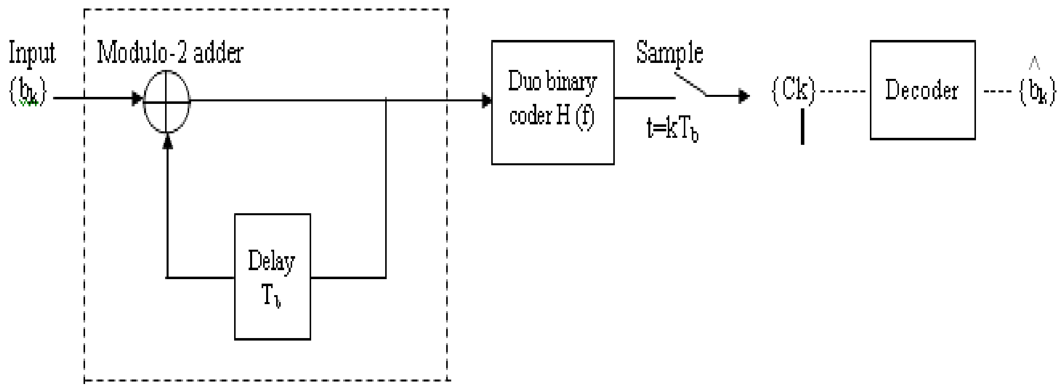


Fig. A precoded duo binary scheme.

$\{a_k\}$ is then applied to duobinary coder, which produce sequence $\{C_k\}$

$$C_k = a_k + a_{k-1}$$

If that symbol at precoder is in polar format $\mathbf{C_k}$ takes three levels,

$$C_k = \begin{cases} \pm 2v & \text{if } b_k = \text{symbol } 0 \\ 0v & \text{if } b_k = \text{symbol } 1 \end{cases}$$

The decision rule for detecting the original input binary sequence $\{b_k\}$ from $\{c_k\}$ is

$$\hat{b}_k = \begin{cases} \text{symbol } 0 & \text{if } |C_k| > 1v \\ \text{symbol } 1 & \text{if } |C_k| \le 1v \end{cases}$$

Example: with start bit as 0, reference bit 1

Transmitter Binary Sequence $\{b_k\}$		0	0	1	0	1	1	0
Precoded Sequence ak $a_k = b_k \oplus a_{k-1}$ (assume start bit as 1 or 0)	1 (a_{k-1})	1	1	0	0	1	0	0
polar Representation of Precoded Sequence a_k	+1	+1	+1	-1	-1	+1	-1	-1
Duo binary coder output $C_k = a_k + a_{k-1}$		2	2	0	-2	0	0	-2
Decoding decision Rule	If $c_k = \pm 2$ decide b_k = Symbol 0 If $c_k = 0$ decide b_k = Symbol 1							
Receiver Received sequence c_k		2	2	0	-2	0	0	-2
Decoded binary sequence b_k		0	0	1	0	1	1	0

Example: with start bit as 0, reference bit 0

Transmitter Binary Sequence $\{b_k\}$		0	0	1	0	1	1	0
Precoded Sequence ak $a_k = b_k \oplus a_{k-1}$ (assume start bit as 1 or 0)	0 (a_{k-1})	0	0	1	1	0	1	1
polar Representation of Precoded Sequence a_k	-1	-1	-1	+1	+1	-1	+1	+1
Duo binary coder output $C_k = a_k + a_{k-1}$		-2	-2	0	+2	0	0	+2
Decoding decision Rule	If $c_k = \pm 2$ decide b_k = Symbol 0 If $c_k = 0$ decide b_k = Symbol 1							
Receiver Received sequence c_k		-2	-2	0	+2	0	0	+2
Decoded binary sequence b_k		0	0	1	0	1	1	0

<u>Example</u>: with start bit as 1, reference bit 1

<u>Transmitter</u> Binary Sequence $\{b_k\}$		1	1	0	1	0	0	1
Precoded Sequence ak $a_k = b_k \oplus a_{k-1}$ (assume start bit as 1 or 0)	1 (a_{k-1})	0	1	1	0	0	0	1
polar Representation of Precoded Sequence a_k	+1	-1	+1	+1	-1	-1	-1	+1
Duo binary coder output $C_k = a_k + a_{k-1}$		0	0	+2	0	-2	-2	0
Decoding decision Rule	If $c_k = \pm 2$ decide b_k = Symbol 0 If $c_k = 0$ decide b_k = Symbol 1							
Receiver Received sequence c_k		0	0	+2	0	-2	-2	0
Decoded binary sequence b_k		1	1	0	1	0	0	1

<u>Example</u>: with start bit as 1, reference bit 0

<u>Transmitter</u> Binary Sequence $\{b_k\}$		1	1	0	1	0	0	1
Precoded Sequence ak $a_k = b_k \oplus a_{k-1}$ (assume start bit as 1 or 0)	0 (a_{k-1})	1	0	0	1	1	1	0
polar Representation of Precoded Sequence a_k	-1	+1	-1	-1	+1	+1	+1	-1
Duo binary coder output $C_k = a_k + a_{k-1}$		0	0	-2	0	2	2	0
Decoding decision Rule	If $c_k = \pm 2$ decide b_k = Symbol 0 If $c_k = 0$ decide b_k = Symbol 1							
Receiver Received sequence c_k		0	0	-2	0	2	2	0
Decoded binary sequence b_k		1	1	0	1	0	0	1

Today the duo-binary techniques are widely applied throughout the world.
While all current applications in digital communications such as data transmission, digital radio, and PCM cable transmission, and other new possibilities are being explored.
This technique has been applied to fiber optics and to high density disk recording which have given excellent results

Example

The binary data **001101001** are applied to the input of a duo binary system.

a)Construct the duo binary coder output and corresponding receiver output, without a precoder.
b) Suppose that due to error during transmission, the level at the receiver input produced by the second digit is reduced to zero. Construct the new receiver output.
c) Repeat above two cases with use of precoder

without a precoder

Input Sequence $\{b_k\}$	0	0	1	1	0	1	0	0	1
Polar Voltage Representation	-1	-1	+1	+1	-1	+1	-1	-1	+1
$c_k = b_k + b_{k-1}$		-2	0	2	0	0	0	-2	0
$\hat{b}_k = c_k - \hat{b}_{k-1}$	-1	-1	+1	+1	-1	+1	-1	-1	+1
Decoded $\hat{b}_k$	0	0	1	1	0	1	0	0	1
If error occurs in second position, c_k received is 0 instead of -2V									
Received C_k		0	0	2	0	0	0	-2	0
polar form $\hat{b}_k$ = ck – $\hat{b}_{k-1}$	-1 $\hat{b}_{k-1}$	1	-1	3	-3	+3	+3	1	-1
Decoded $\hat{b}_k$		1	0	1	0	1	0	1	0

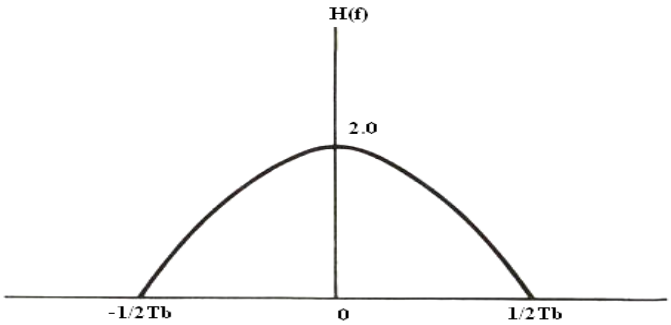
errors errors

With a precoder (start bit 1)

Input Sequence {b_k}		**0**	**0**	**1**	**1**	**0**	**1**	**0**	**0**	**1**
Precodeed sequence {a_k} = $b_k \oplus a_k$-1	**1**	**1**	**1**	**0**	**1**	**1**	**0**	**0**	**0**	**1**
Polar Representation	**+1**	**+1**	**+1**	**-1**	**+1**	**+1**	**-1**	**-1**	**-1**	**+1**
Duobinary coded sequence $c_k = a_k + a_{k-1}$		**2**	**2**	**0**	**0**	**2**	**0**	**-2**	**-2**	**0**
Decision b_k **c_k > 1 symbol 0** **c_k < 1 symbol 1**		**0**	**0**	**1**	**1**	**0**	**1**	**0**	**0**	**1**
If error occurs in 2nd position then voltage level of c_k = 0 , then										
Received c_k		**2**	**0**	**0**	**0**	**2**	**0**	**-2**	**-2**	**0**
Decision for b_k **c_k > 1 symbol 0** **c_k < 1 symbol 1**		**0**	**1** ↑	**1**	**1**	**0**	**1**	**0**	**0**	**1**

With a precoder (start bit 0)

Input Sequence $\{b_k\}$		**0**	**0**	**1**	**1**	**0**	**1**	**0**	**0**	**1**
Precodeed sequence $\{a_k\} = b_k \oplus a_k\text{-}1$	**0**	**0**	**0**	**1**	**0**	**0**	**1**	**1**	**1**	**0**
Polar Representation	**-1**	**-1**	**-1**	**+1**	**-1**	**-1**	**+1**	**+1**	**+1**	**-1**
Duobinary coded sequence $c_k = a_k + a_{k-1}$		**-2**	**-2**	**0**	**0**	**-2**	**0**	**+2**	**+2**	**0**
Decision b_k $c_k > 1$ symbol 0 $c_k < 1$ symbol 1		**0**	**0**	**1**	**1**	**0**	**1**	**0**	**0**	**1**
If error occurs in 2nd position then voltage level of $c_k = 0$, then										
Received c_k		**-2**	**0**	**0**	**0**	**-2**	**0**	**+2**	**+2**	**0**
Decision for b_k $c_k > 1$ symbol 0 $c_k < 1$ symbol 1		**0**	**1** ↑	**1**	**1**	**0**	**1**	**0**	**0**	**1**

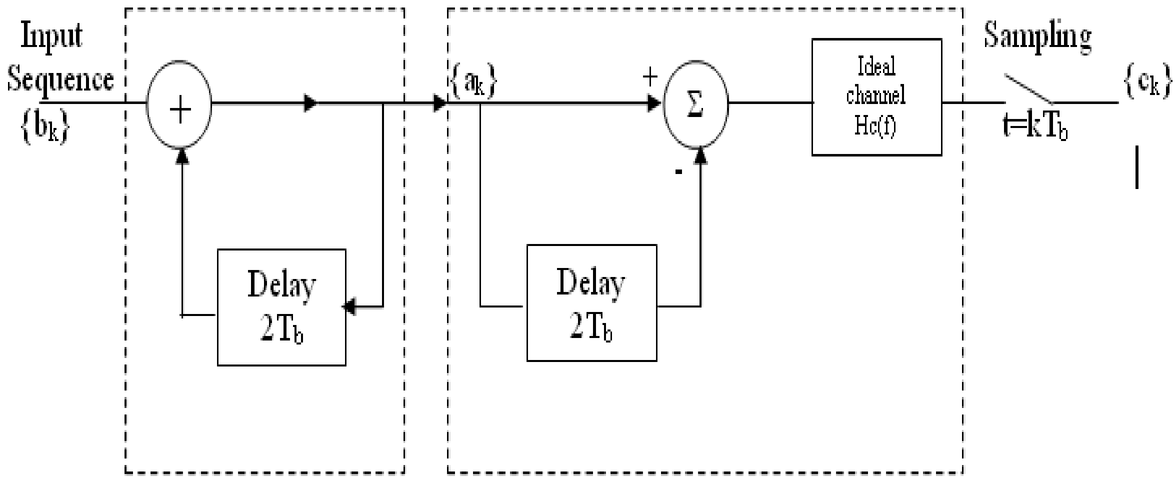


The Transfer function H(f) of Duo binary signalling has non zero spectral value at origin, hence not suitable for channel with Poor DC response. This drawback is corrected by Modified Duobinary scheme.

Modified Duobinary scheme.

It is an extension of the duo-binary signaling. The modified duo binary technique involves a correlation span of two binary digits. Two-bit delay causes the ISI to spread over two symbols. This is achieved by subtracting input binary digits spaced $2T_b$ secs apart.

Modified Duobinary scheme.



Transmitter

The precoded output sequence is given by $\mathbf{a_k = b_k \oplus a_{k-2}}$
a modulo 2 logical addition

If $a_k = \pm 1$v , C_k takes one of three values 2,0,-2.

Output sequence of modified duo binary filter is given by $\mathbf{C_k}$

$\mathbf{C_k = a_k - a_{k-2}}$

C_k takes one of three values 2,0,-2

Ck = 0V, if b_k is represented by symbol 0
Ck = $\pm$2V, if b_k is represented by symbol 1

Receiver

At the receiver we may extract the original sequence $\{b_k\}$ using the decision rule

$$b_k = \begin{cases} \text{symbol 0 if } |C_k| > 1\text{v} \\ \text{symbol 1 if } |C_k| \le 1\text{v} \end{cases}$$

The Transfer function of the filter is given by

$$H(f) = H_c(f) - H_c(f)e^{-j4\pi f T_b}$$

$$= H_c(f)\left[1 - e^{-j4\pi f T_b}\right]$$

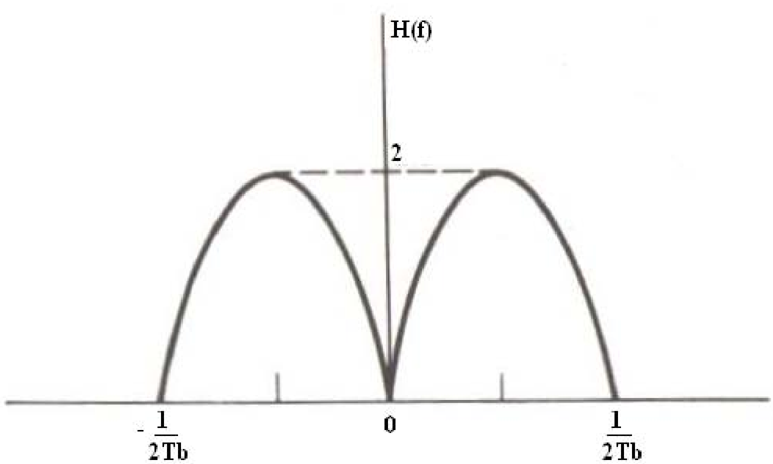
$$= 2_j H_c(f)\ e^{-j2\pi f T_b}\left[\frac{e^{+j2\pi f T_b} - e^{-j2\pi f T_b}}{2j}\right]$$

$$= 2_j H_c(f)\sin(2\pi f T_b)e^{-j2\pi f T_b}$$

Where $H_c(f)$ is $\begin{cases} 1 & |f| \le \frac{1}{2T_b} \\ 0 & \text{otherwise} \end{cases}$

$$H(f) = \begin{cases} 2j\sin(2\pi f T_b)e^{-j2\pi f T_b} & |f| \le \frac{1}{2T_b} \\ 0 & \text{Otherwise} \end{cases}$$

The Transfer function has zero value at origin, hence suitable for poor dc channels



Impulse response

Impulse response h(t) is obtained by taking Inverse Fourier transformation of H(f)

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

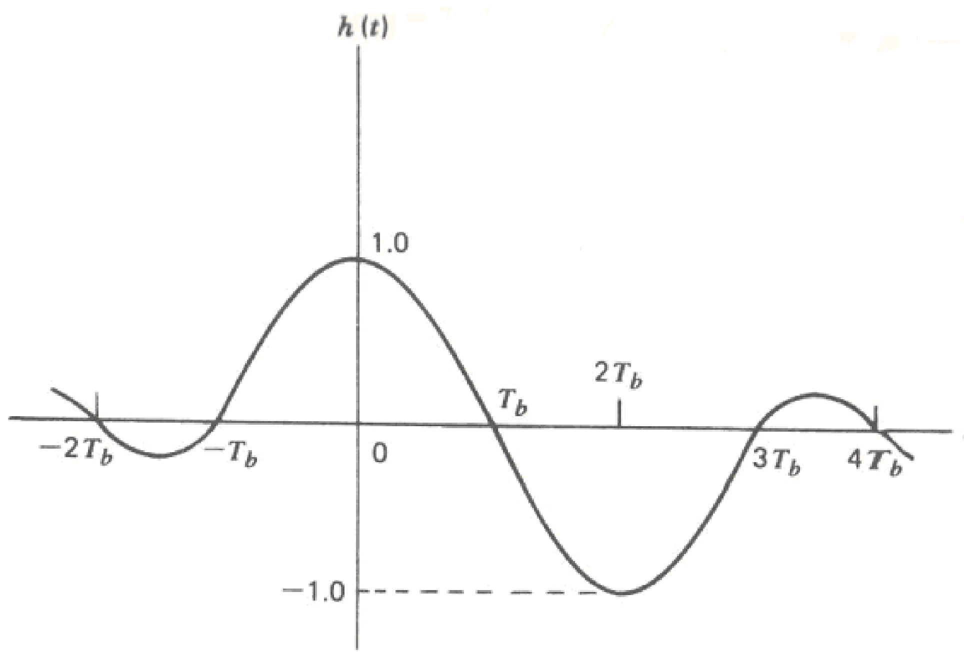
$$= \int_{-1/2T_b}^{1/2T_b} 2j\sin(2\pi f T_b)e^{-j\pi f T_b}[e^{j2\pi f t}]df$$

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

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

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

Impulse response has three distinguishable levels at the sampling instants.



To eliminate error propagation modified duo binary employs Precoding option same as previous case.

Prior to duo binary encoder precoding is done using modulo-2 adder on signals spaced $2T_b$ apart

$$a_k = b_k \oplus a_{k-2}$$

Example : Consider binary sequence $\{b_k\}$= {01101101} applied to input of a precoded modified duobinary filter. Determine receiver output and compare with original $\{b_k\}$.

Binary sequence $\{b_k\}$			0	1	1	0	1	1	0	1
Precoded sequence $a_k = b_k \oplus a_{k-2}$	1 (a_{k-2})	1 (a_{k-1})	1	0	0	0	1	1	1	0
Polar Representation	+1	+1	+1	-1	-1	-1	+1	+1	+1	-1
Transmitted output $c_k = a_k - a_{k-2}$			0	-2	-2	0	+2	+2	0	-2
Received Sequence decision $\lvert C_k \rvert < 1V \rightarrow 0$, $\lvert C_k \rvert > 1V \rightarrow 1$			0	-2	-2	0	2	2	0	-2
Decoded $\hat{b}_k$			0	1	1	0	1	1	0	1

Consider binary sequence {bk}={01101101}

Binary sequence $\{b_k\}$			0	1	1	0	1	1	0	1
Precoded sequence $a_k = b_k \oplus a_{k-2}$	0 (a_{k-2})	0 (a_{k-1})	0	1	1	1	0	0	0	1
Polar Representation	-1	-1	-1	+1	+1	+1	-1	-1	-1	+1
Transmitted output $c_k = a_k - a_{k-2}$			0	+2	2	0	-2	-2	0	2
Received Sequence decision $\lvert C_k \rvert < 1V \rightarrow 0$, $\lvert C_k \rvert > 1V \rightarrow 1$			0	+2	2	0	-2	-2	0	2
Decoded $\hat{b}_k$			0	1	1	0	1	1	0	1

Example

The binary data 011100101 are applied to the input of a modified duo binary system.

a) Construct the modified duobinary coder output and corresponding receiver output, without a precoder.

b) Suppose that due to error during transmission, the level at the receiver input produced by the third digit is reduced to zero. Construct the new receiver output.

c) Repeat above two cases with use of precoder

Modified duobinary coder output and corresponding receiver output, without a precoder

Binary sequence $\{b_k\}$			0	1	1	1	0	0	1	0	1
Polar Representation	+1	+1	-1	+1	+1	+1	-1	-1	+1	-1	+1
Transmitted output $c_k = b_k - b_{k-2}$			-2	0	+2	0	-2	-2	2	0	0
Received Sequence			-2	0	+2	0	-2	-2	2	0	0
Decision $b_k = c_k + b_{k-2}$	+1	+1	-1	+1	+1	+1	-1	-1	+1	-1	+1
Decoded $\hat{b}_k$			0	1	1	1	0	0	1	0	1

If error occurs in 3rd position then voltage level of $c_k = 0$, then											
Received c_k			**-2**	**0**	**0**	**0**	**-2**	**-2**	**2**	**0**	**0**
Decision for b_k $\hat{b}_k$ $b_k = c_k + b_{k-2}$	**+1**	**+1**	**-1**	**+1**	**-1**	**+1**	**-3**	**-1**	**-1**	**-1**	**-1**
Decoded			**0**	**1**	**0**	**1**	**0**	**0**	**0**	**0**	**0**

Binary sequence $\{b_k\}$			0	1	1	1	0	0	1	0	1
Precoded sequence $a_k = b_k \oplus a_{k-2}$	1 (a_{k-2})	1 (a_{k-1})	1	0	0	1	0	1	1	1	0
Polar Representation	+1	+1	+1	-1	-1	+1	-1	+1	+1	+1	-1
Transmitted output $c_k = a_k - a_{k-2}$			0	-2	-2	2	0	0	2	0	0
Received Sequence decision $\lvert C_k \rvert < 1V \rightarrow 0$, $\lvert C_k \rvert > 1V \rightarrow 1$			0	-2	-2	2	0	0	2	0	0
Decoded			0	1	1	1	0	0	1	0	1

Modified duo binary coder output and corresponding receiver output, with a precoder

If error occurs in 3rd position then voltage level of $c_k = 0$, then										
Received c_k		0	-2	0	2	0	0	2	0	0
Decision for b_k $c_k > 1$ symbol 0 $c_k < 1$ symbol 1		0	1	0	1	0	0	1	0	1

Generalized form of correlative coding scheme

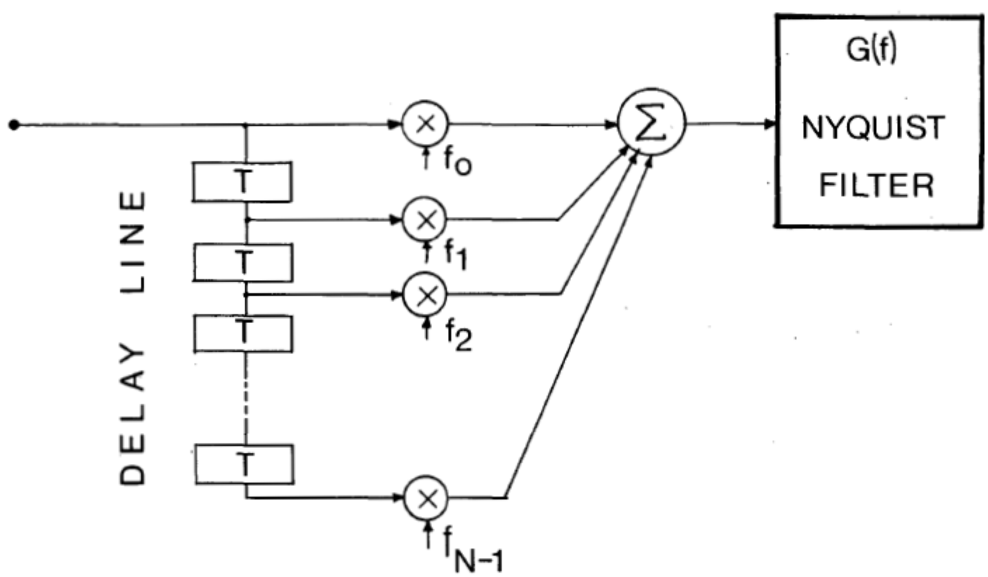


Fig. Generalized Correlative Coding

The Duo binary and modified Duo binary scheme have correlation spans of one binary digit and two binary digits respectively. This generalisation scheme involves the use of a tapped delay line filter with tap weights f0, f1, f2,… fn-1. A correlative samples C_k is obtained from a superposition of 'N' successive input sample values b_K

$$\mathbf{Ck} = \sum_{n=0}^{N-1} f_n b_{k-n}$$

By choosing various combination values for f_n, different correlative coding schemes can be obtained from simple duo binary.

Base band Transmission of M-ary data

In base band M-ary PAM, output of the pulse generator may take on any one of the M-possible amplitude levels with M>2 for each symbol
The blocks of n- message bits are represented by M-level waveforms with

$$\mathbf{M=2^n.}$$

Ex: **M=4** has 4 levels. possible combination are 00, 10, 11, 01

$\mathbf{T= 2T_b}$ is termed **symbol duration** .
In general symbol duration $\mathbf{T = T_b \log_2 M}$.

M-ary PAM system is able to transmit information at a rate of $\mathbf{\log_2 M}$ faster than binary PAM for given channel bandwidth.

$$\mathbf{R} = \frac{\mathbf{Rb}}{\mathbf{\log_2 M}}$$

R_b = bit rate for binary system
R = symbol rate for binary system

M-ary PAM system requires more power which is increased by factor equal to

$\mathbf{M^2 / \log_2 M}$ for same average probability of symbol error.

M-ary Modulation is well suited for the transmission of digital data over channels that offer a **limited bandwidth and high SNR**

Example
An analog signal is sampled, quantised and encoded into a binary PCM wave. The number of representation levels used is 128. A synchronizing pulse is added at

the end of each code word representing a sample of the analog signal. The resulting PCM wave is transmitted over a channel of bandwidth 12kHz using binary PAM system with a raised cosine spectrum. The roll off factor is unity.

a)Find the rate (in BPS) at which information is transmitted through the channel.

b) Find the rate at which the analog signal is sampled. What is the maximum possible value for the highest frequency component of the analog signal.

Solution

Given Channel with transmission BW B=12kHz.
Number of representation levels L = 128
Roll off α = 1

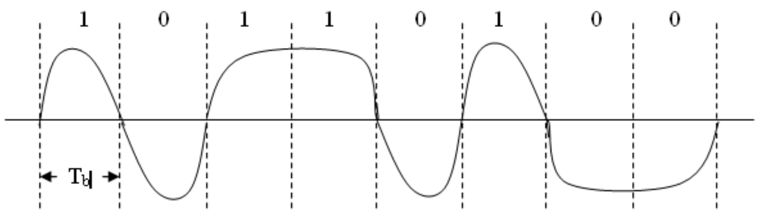
a) B = Bo(1+ α),
Hence Bo =6kHz.
$Bo=R_b/2$ therefore R_b = 12kbps.

b) For L=128, $L = 2^n$, n = 7
symbol duration $T = T_b \log_2 M = nT_b$
sampling rate $f_s = R_b/n$ = 12/7 = 1.714kHz.
And maximum frequency component of analog signal is
From LP sampling theorem $w = f_s/2$ = 857Hz.

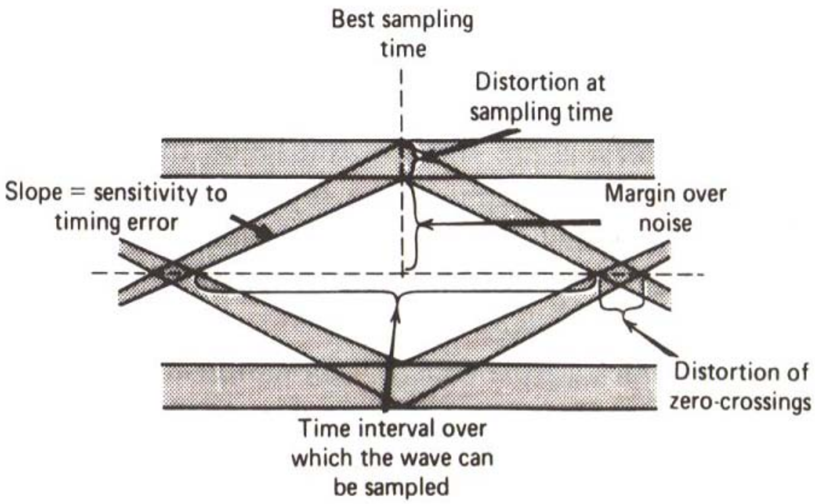
<u>Eye pattern</u>

The quality of digital transmission systems are evaluated using the bit error rate. Degradation of quality occurs in each process modulation, transmission, and detection. The eye pattern is experimental method that contains all the information concerning the degradation of quality. Therefore, careful analysis of the eye pattern is important in analyzing the degradation mechanism.

- Eye patterns can be observed using an oscilloscope. The received wave is applied to the vertical deflection plates of an oscilloscope and the sawtooth wave at a rate equal to transmitted symbol rate is applied to the horizontal deflection plates, resulting display is <u>eye pattern</u> as it resembles human eye.
- The interior region of eye pattern is called eye opening



We get superposition of successive symbol intervals to produce eye pattern as shown below.



Interpretation of eye pattern

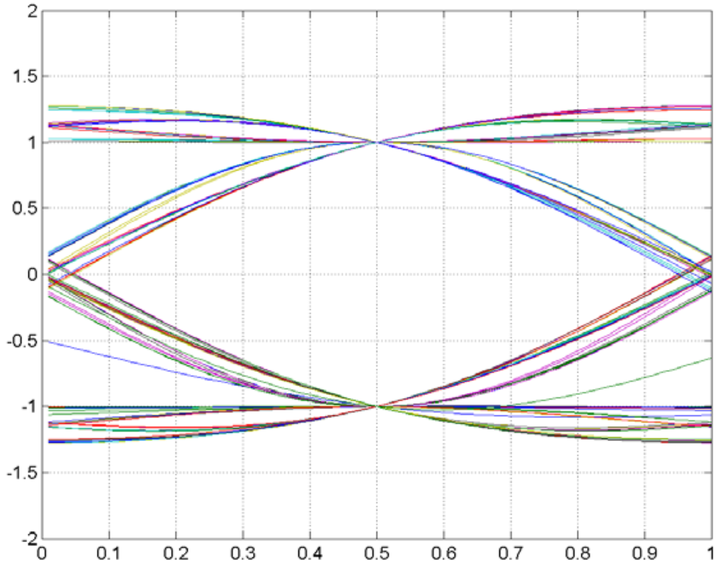
- The width of the eye opening defines the time interval over which the received wave can be sampled without error from ISI
- The optimum sampling time corresponds to the maximum eye opening
- The height of the eye opening at a specified sampling time is a measure of the margin over channel noise.

The sensitivity of the system to timing error is determined by the rate of closure of the eye as the sampling time is varied.

Any non linear transmission distortion would reveal itself in an asymmetric or squinted eye. When the effected of ISI is excessive, traces from the upper portion of the eye pattern cross traces from lower portion with the result that the eye is completely closed.

Example of eye pattern:

Binary-PAM Perfect channel (no noise and no ISI)



Example of eye pattern: Binary-PAM with noise no ISI

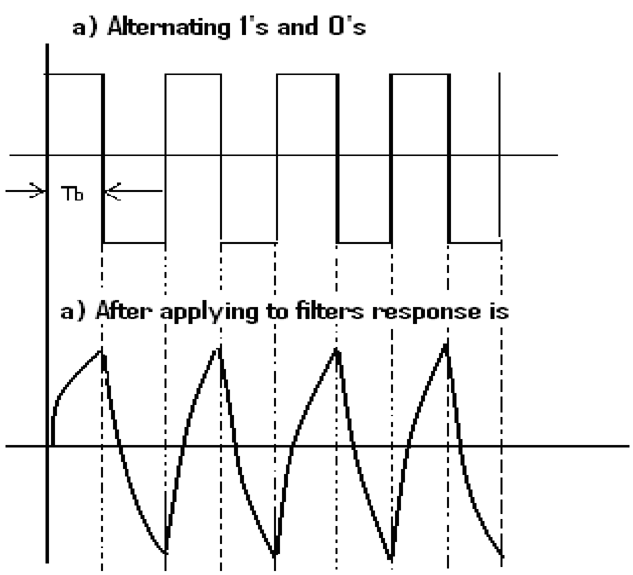
Example 1

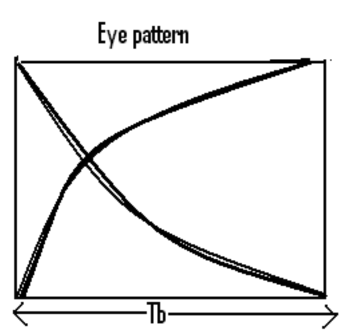
A binary wave using polar signaling is generated by representing symbol 1 by a pulse of amplitude -1v; in both cases the pulse duration equals the bit duration. The signal is applied to a low pass RC filter with transfer function

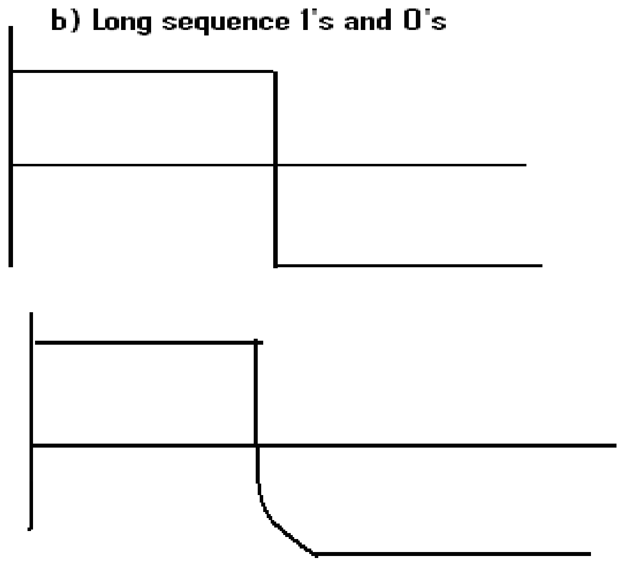
$$\mathbf{H(f)} = \frac{\mathbf{1}}{\mathbf{1} + \mathbf{jf/f_0}}$$

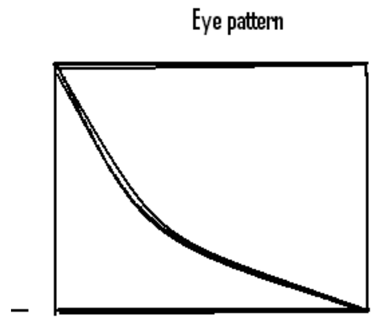
Construct eye pattern for the filter for

1. Alternating 1s and 0s
2. A long sequence of 1s followed by long sequence of zeros.



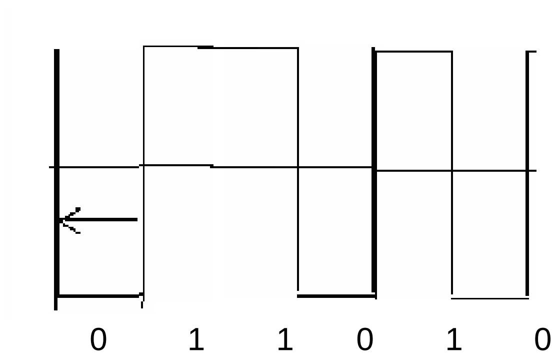


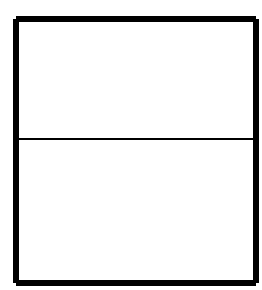




Example 2

The binary sequence 011010 is transmitted through channel having a raised cosine characteristics with roll off factor unity. Assume the use of polar signaling, format. construct the Eye pattern





Eye Pattern

Adaptive equalization for data transmission

This technique is another approach to minimize signal distortion in the base band data transmission. This is Nyquist third method for controlling ISI.

Equalization is essential for high speed data transmission over voice grade telephone channel which is essentially linear and band limited.

High speed data transmission involves two basic operations:

i) **Discrete pulse amplitude modulation**:

The amplitudes of successive pulses in a periodic pulse train are varied in a discrete fashion in accordance with incoming data stream.

ii) **Linear modulation**:

Which offers band width conservation to transmit the encoded pulse train over telephone channel.

At the receiving end of the systems, the received waves is demodulated and then synchronously sampled and quantized. As a result of dispersion of the pulse shape by the channel the number of detectable amplitude levels is limited by ISI rather than by additive noise. If the channel is known , then it is possible to make ISI arbitrarily small by designing suitable pair of transmitting and receiving filters for pulse shaping.

In switched telephone networks we find that two factors contribute to pulse distortion.

1. Differences in the transmission characteristics of individual links that may be switched together.

2. Differences in number of links in a connection

Because of these two characteristics, telephone channel is random in nature. To realize the full transmission capability of a telephone channel we need adaptive equalization.

<u>Adaptive equalization</u>

- An equalizer is a filter that compensates for the dispersion effects of a channel. Adaptive equalizer can adjust its coefficients continuously during the transmission of data.

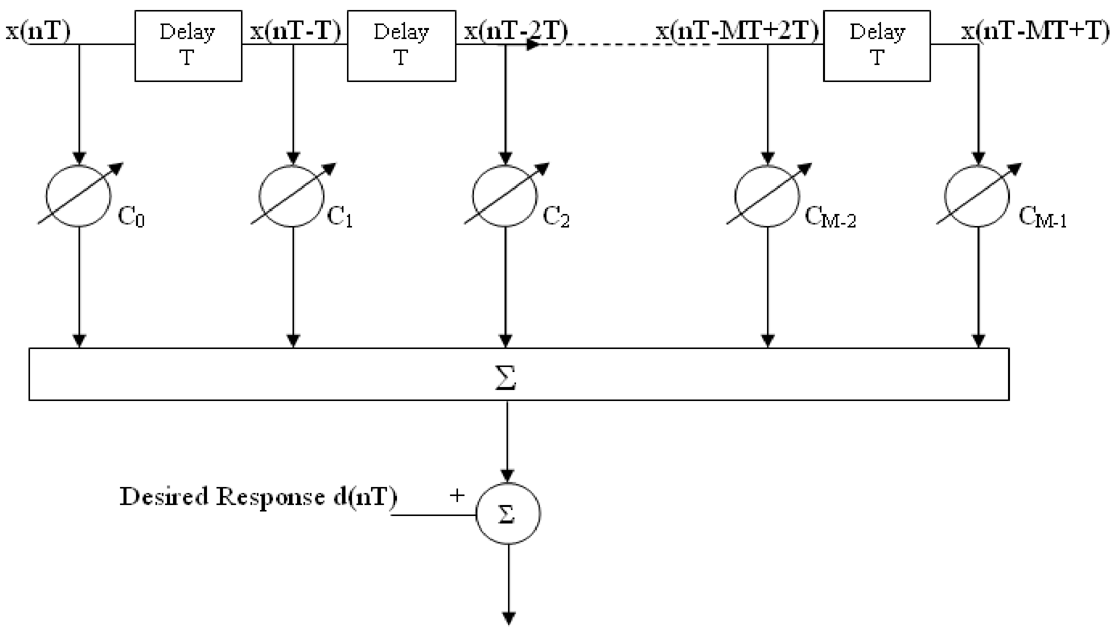
Pre channel equalization

- requires feed back channel
- causes burden on transmission.

Post channel equalization

Achieved prior to data transmission by training the filter with the guidance of a training sequence transmitted through the channel so as to adjust the filter parameters to optimum values.

Adaptive equalization – It consists of tapped delay line filter with set of delay elements, set of adjustable multipliers connected to the delay line taps and a summer for adding multiplier outputs.



The output of the Adaptive equalizer is given by

$$\mathbf{y(nt)} = \sum_{i=-0}^{M-1} \mathbf{c_i\, x(nT - iT)}$$

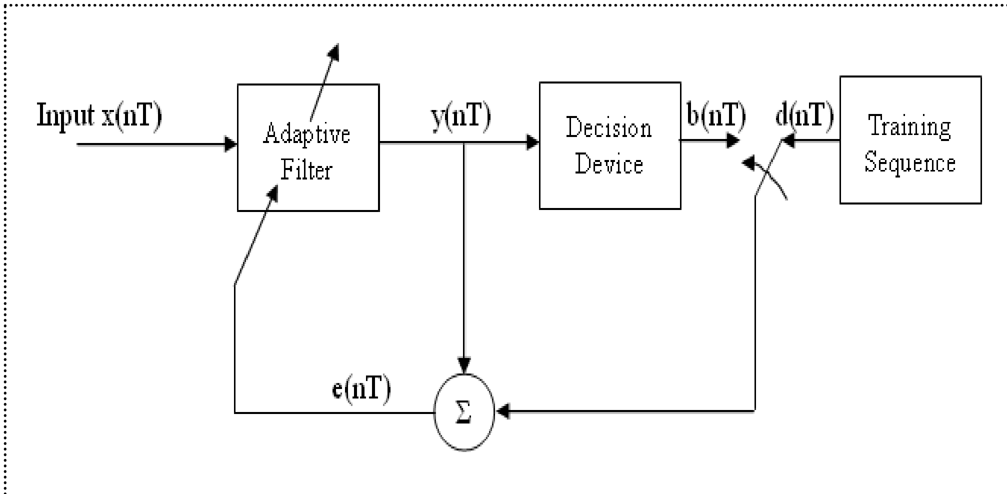
C_i is weight of the i^{th} tap Total number of taps are M .Tap spacing is equal to symbol duration T of transmitted signal

In a conventional FIR filter the tap weights are constant and particular designed response is obtained. In the adaptive equaliser the $\mathbf{C_i}$'s are variable and are adjusted by an algorithm

Two modes of operation

1. Training mode 2 . Decision directed mode

Mechanism of adaptation



Training mode

A known sequence d(nT) is transmitted and synchronized version of it is generated in the receiver applied to adaptive equalizer.

This training sequence has maximal length PN Sequence, because it has large average power and large SNR, resulting response sequence (Impulse) is observed by measuring the filter outputs at the sampling instants.

The difference between resulting response y(nT) and desired response d(nT)is error signal which is used to estimate the direction in which the coefficients of filter are to be optimized using algorithms

Methods of implementing adaptive equalizer

i) Analog
ii) Hard wired digital
iii) Programmable digital

Analog method

- **Charge coupled devices** [CCD's] are used.
- CCD- FET's are connected in series with drains capacitively coupled to gates.
- The set of adjustable tap widths are stored in digital memory locations, and the multiplications of the analog sample values by the digitized tap weights done in analog manner.
- Suitable where symbol rate is too high for digital implementation.

Hard wired digital technique

- Equalizer input is first sampled and then quantized in to form that is suitable for storage in shift registers.
- Set of adjustable lap weights are also stored in shift registers. Logic circuits are used for required digital arithmetic operations.
- widely used technique of equalization

Programmable method

- Digital processor is used which provide more flexibility in adaptation by programming.

Advantage of this technique is same hardware may be timeshared to perform a multiplicity of signal processing functions such as filtering, modulation and demodulation in modem.