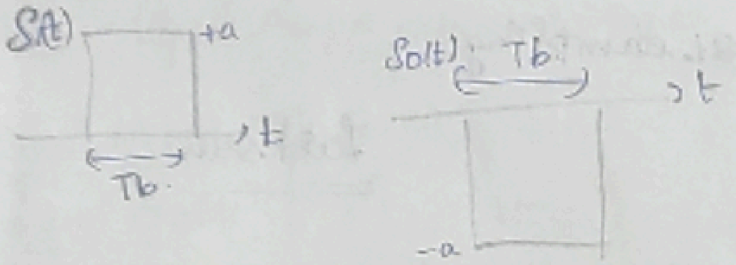
om sri sai ram

12/08/2

Digital Communication:

- We try to map 0,1 into well defined wave forms
- Binary Communication System:

Map (0,1) To $\{S_0(t), S_1(t)\}$



where T_b = Bit Duration

$R_b = \frac{1}{T_b}$ = Bit Rate (bit/sec)

- It is not necessary rectangular wave form but its the convention
- Reciever has to take Decision on the recieved system.
- The reciever must know the $S_0(t)$, $S_1(t)$.
- The distortion may appear, the reciever must have idea of it

Digital modulation = Signal Mapping [Signalling]

- Usually m(t) can be of any shape in AM waves, so reciever has to decode from an infinite set. Since message signal is continuous signal.
- But here, Finite Symbol Set [Symbol waveform]

So only reciever choose from $S_0(t)$ or $S_1(t)$

M-ary Modulation M = 4

Symbols = Waveform

00 → $S_1(t)$ 01 → $S_2(t)$ 10 → $S_3(t)$ 11 → $S_4(t)$

$\{S_1(t), S_2(t), S_3(t), S_4(t)\}$

Any non-binary modulation ⟶ M-array modulation

. Data is represented in binary format.

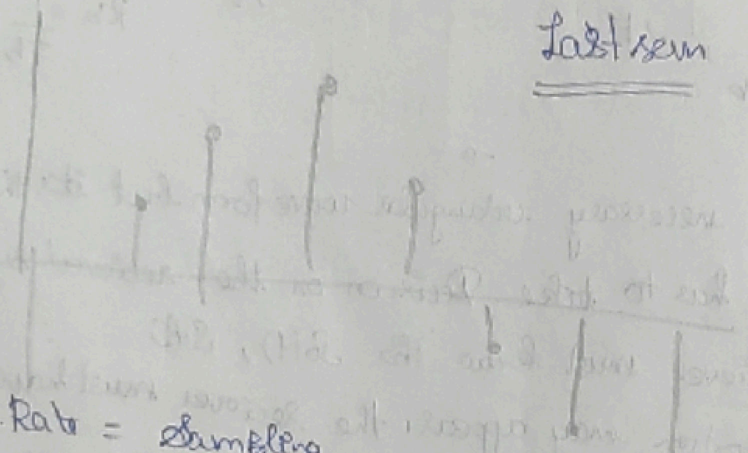
Message m(t) is encoded using Binary set {0,1}

(Waveform encoding)

Waveform encoding:

2. Quantization   ~~2~~ 1. Sampling

Last sem



Bit Rate = Sampling frequency × No. of ~~sampling~~ bits per sample

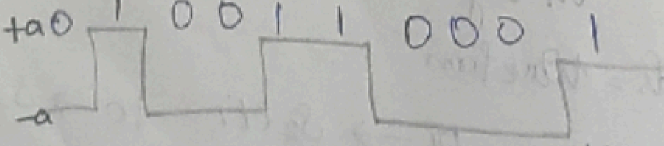
= 3Fs.

- Continuous Signal → Sampling → Quantization → Digital Signal

So we go ahead with Digital signal.

Map {0,1} To {S0(t), S1(t)}

⟹ Transmitted signal



⇒ Signal distorted? by channel

[Bandwidth limitation]

• In case of ideal channel, ~~distort~~ uniform attenuation of 0.5 is given to every freq so only amplitude changes, the signal shape remains same

⇒ Signal distorted? by chanel + noise

This noise is usually white noise

⇒ Signal detected by reciever

Analog Vs Digital

• Repeater in Analog Modulation AM

* As the signal is transmitted, Recieved signal loses

As power, as power $\frac{A_c^2}{2}$

⇒ As Recieved signal is weak

⇒ Change in shape of signal

⇒ Addition of noise

- Inorder to overcome these issues, Reciever uses modulation and amplification.

Amplification of signal leads to unintentional amplification of the $\boxed{\text{in band noise}}$

This is because the noise is white noise so it has all frequency.

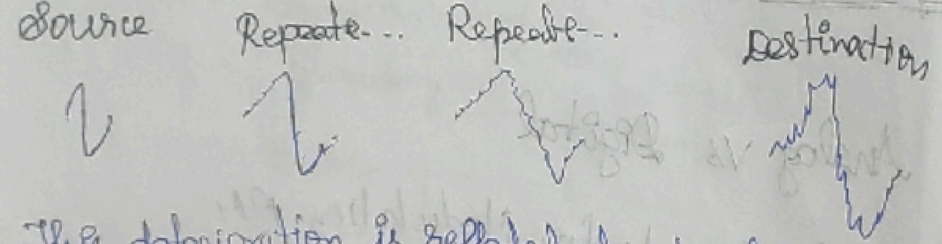
Let My required freq be 200H-200.5H Hz

then white noise has 200H-200.5H Hz also

So, the ~~bar~~ noise gets amplified

This is called in band noise.

Analog Long-Distance Communication.



The deterioration is reflected due to long distance transmission

- But as distance is large, the recieved signal is very weak, noise is too loud, so decision can be wrong.
- So Reduce distances,

This can be done by Regenerative Repeaters

Source ⟶ Regenerator ... Regenerator ⟶ Destination

- Decision on recieved symbol
- Regeneration of signal at the repeater.

⇒ Performance dependent of the signal on distance,

regenerative repeaters

Advantages of Digi Com:-

- Regenerative Repeaters
- Flexibility of implementation
- Privacy & Security through encryption
- Error detection & Correction
- Multiplexing of heterogenous signals

Disadvantages:-

- Higher Bandwidth
- Complex Synchronization Technique

⇒ Bandwidth $\propto$ Bitrate

$\propto$ Kind of Modulation

13/08/2020

• 0, 1 are abstract, so they need a physical form.

So these waveform should match spectra of receiving

How do we transmit bits over wires/wireless:

Maps the bits to waveform → Signalling

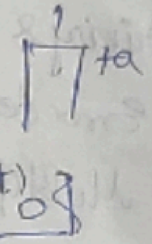
→ Baseband signalling [Linecodes]

→ Bandpass

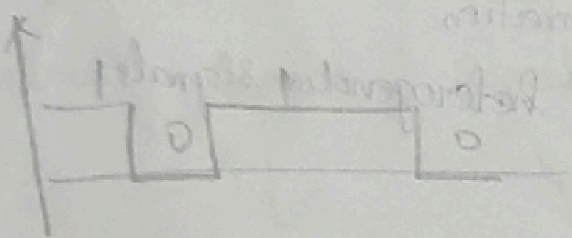
Both require Mapping

Baseband signalling (Mapping)

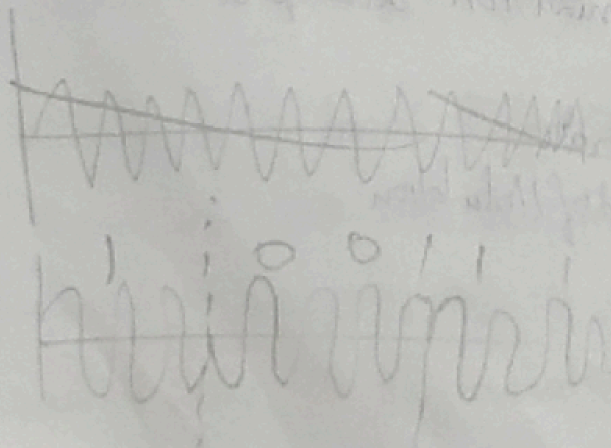
$S_1(t)$



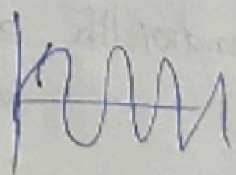
$S_2(t)$



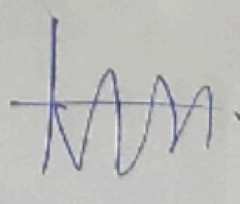
Bandpass signalling (Mapping):



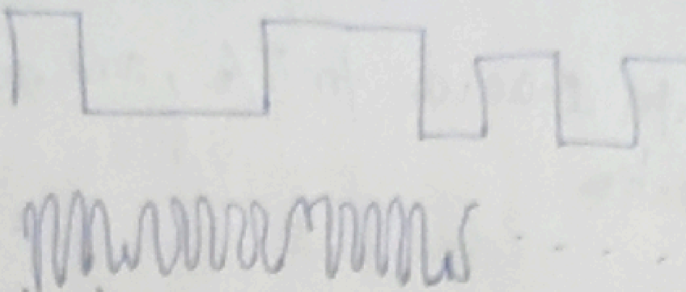
$S_1(t)$



$S_2(t)$



Basepass Vs Bandpass

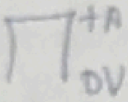


Baseband

Passband

$1 \rightarrow S_1(t)$

$0 \rightarrow S_2(t)$



$S_1(t)$

Tb.

$S_2(t)$

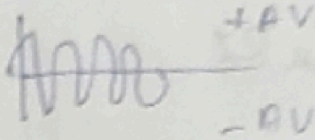
$S_1(t)$ = Infinite passi but mostly 0

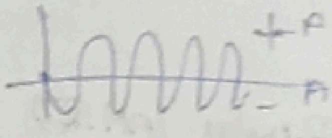
$S_2(t) = 0$.

Fourier transfer.

$1 \rightarrow S_1(t)$

$0 \rightarrow S_2(t)$





By fourier transform

Sinc function

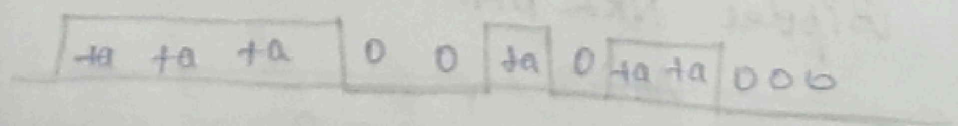
F(c) have much much ~~bett~~ more than bandwidth

Carrier freq is going to be high

Line codes [Baseband signalling)

Unipolar non return to zero.

1 1 1 0 0 1 0 1 1 0 0 0

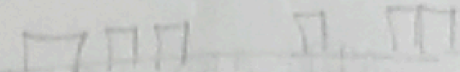


→ It will be +a for the whole duration.

+ On off NRZ

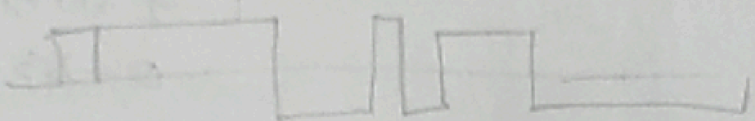
Unipola RZ.

+a increases for T/2, and decreases for T/2 duration.

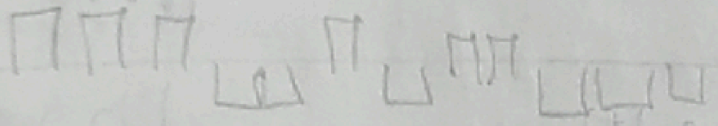


Polar NRZ

It contains negative

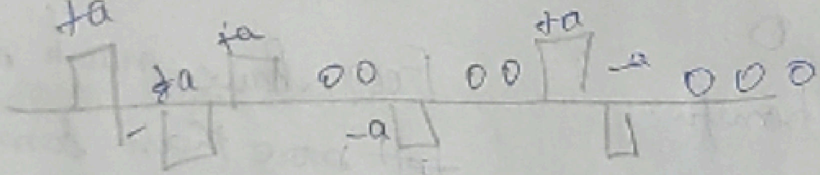


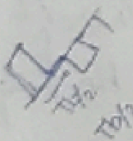
Polar RZ



Bipolar (AMI RZ)

Alternate Mark Inversion (RZ)



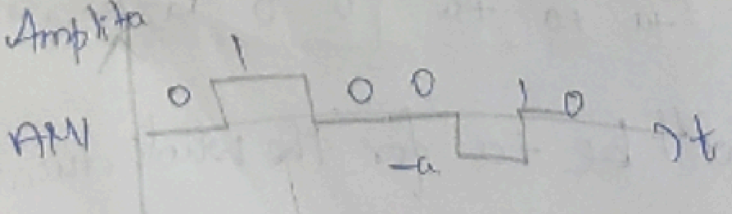


0 ⇒ always 0

1 → is inverting +a then -a, +a, -a.

B Forbidden Component = +a, +a, ~~Foo~~ -a, -a.

Bipolar NRZ AMI



Split Fade code

Manchester:-

100110

1 ⇒ 1 positive, then negative

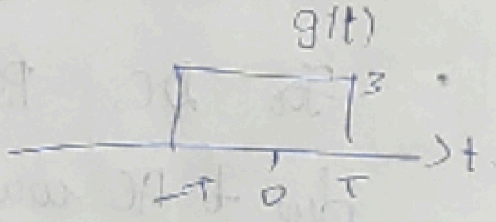
0 → 1 negative to positive

Desirable Properties of Linecodes

- Low Transmission Bandwidth
- Favourable Power Spectral Density
- High power Efficiency
- Error detection & Correction capability
- Adequate Timing Content
- Transparency

⇒ Deterministic Signal:

ex: $x(t) = e^{-2t}$



⇒ Random Process:

Ex: Statistical Class of Speech Signals

Frequency Domain Characterisation

$x(t)$

$$X(f) = \int_{-\infty}^{\infty} x(t)\, e^{2\pi i f t}\, dt.$$

Random Process $X(t)$

Autocorrelation Function $R_x(t)$

and Power Spectral Density $Sx(f)$
form a Fourier Transform Pair

$$Sx(f) = \int_{-\infty}^{\infty} Rx(\tau)\, e^{-j2\pi f\tau}\, d\tau$$

$Sx(f)$ = Watts per Hertz

Square of the magnitude of Fourier transform

Power Spectral Density:
How A power of signal is distributed over various frequencies.

Favourable Power Spectral Density:

- Low transmission Bandwidth
- For DC, PSD = 0.

due to AC coupling of transformers used in repeaters. Significant power in low-freq components causes dc to wander in pulse ~~sep~~ stream when ac

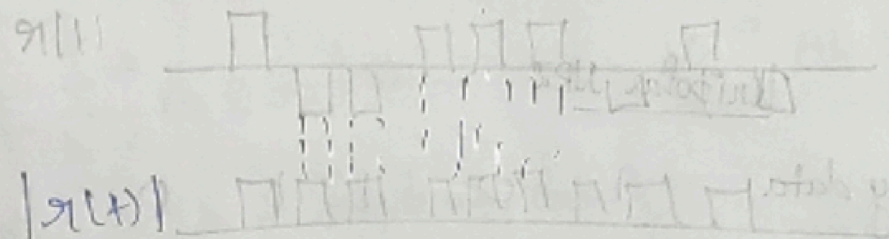
- AC coupling is required because dc paths provide by cable pairs between repeaters ~~site~~ sites are used to transmit power required for repeaters

Power Efficiency:
- For a particular data rate or bit rate

Adequate Timing Content :-

The line code should have extractable clock information from signal

Timing Signal :
- We want to be able to easily extract the timing information from the signal
- Consider the polar RZ $r(t)$.

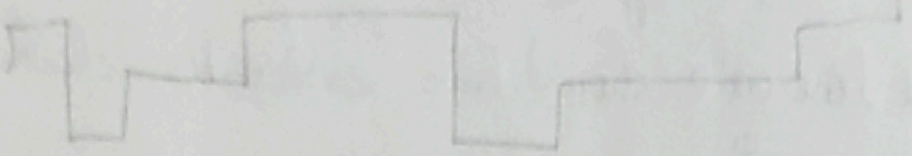


~~Extract~~ The clock extracted must exactly be synchronised with the transmitted clock

⇒ We get that by taking absolute value.
modulus of signal

Manchester codes contain transitions in every bit is very good for clock extraction

- NRZ codes can be more problematic. Long strings of 1's or 0's can cause a loss of synchronization.



Transparency:-

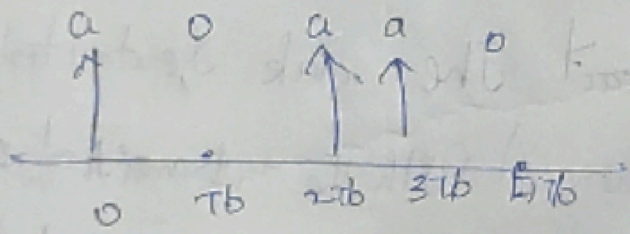
- As a linecode in which the bit pattern doesn't affect accuracy of timing.

(Ex:) NRZ is not transparent to long string of 0's 1s.

Unipolar NRZ

Binary data 1 0 1 1 0

$A_k =$



Impulse Response $v(t)$ NZR Encoder

Then it is mathematically represented as

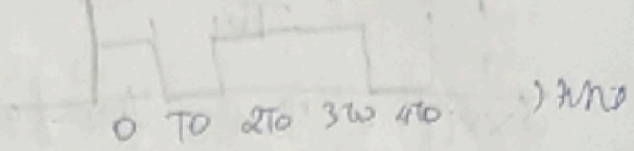
$$a(\delta(t)) + 0\,\delta(t-Tb) + a\,\delta(t-2Tb) + a\,\delta(t-3Tb) + 0\,\delta(t-3Tb).$$

Since amplitude keeps changing let the A_k

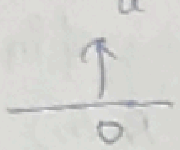
where $A_k = 0/1$

then Input data $x(t) = \sum_{k=-\infty}^{\infty} A_k (t - kT_b)$

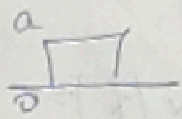
Impulse response $v(t)$ NRZ Encoder. $\longrightarrow$



The impulse response of the is response of a system applied at $t=0$



Impulse response



$\delta(t)$ Impulse Respon $v(t)$

Input $\sum_{k=-\infty}^{\infty} A_k \delta(t-kT_b) \longrightarrow \sum_{k=-\infty}^{\infty} A_k v(t-kT_b)$

This is just a simple ~~conv~~ convolution,

$$= \sum A(kT_b) * v(t-kT_b)$$

It's a convolution random process with deterministic
bit seq. wave for

$$A_k * v(t)$$

Power spectral Density

$$g(t) = S_{gg}(f) = |V(f)|^2 S_{xx}(f)$$

If
$\Rightarrow$ Power spectral density of input,

$S_{xx}(f)$

PSD of output $S_{yy}(f) = |H(f)|^2 S_{xx} f$.

~~Here~~ in $V(t)$ = Frequency Response of Filter,

The input is a random Binary Sequence

$$S_{gg}(f) = \frac{1}{T_b} |V(f)|^2 \sum_{n=-\infty}^{\infty} R_A(n) e^{(-j2\pi n f T_b)}$$

$S_{XX}(f)$ = Fourier Transform of $R_A(n)$

$R_A(n)$ = discrete time signal = ~~Fourier transf of~~ the autocorrelation function of input

$S_{XX}(f)$ = Fourier Transform of $R_A(n)$

$$|V(f)|^2 = T_b^2 \operatorname{sinc}^2(\pi f T_b)$$

Fourier of $v(t)$ = sinc function.

⇒ Now find Autocorrelation function

Autocorrelation Sequence $R_A(n)$

0 a 0 0 a 0 a a a 0 a 0 0

$$R_A(n) = E(A_k A_{k+n})$$

$$R_A(0) = E(A_k^2) = \sum_{k=1}^{2} (P_k A_k^2)$$

$$= P(A_k = 0)0^2 + P(A_k = a)a^2$$

$$= \frac{1}{2} \times 0 + \frac{1}{2} a^2 = \frac{a^2}{2}$$

Autocorrelation means each possibility has the same probability. In the seq

0 a 00 a 0 0a a a 0 a 0 0

$p(0) = p(a) = \frac{1}{2}$

$R_{xx}(\tau) = E\left[x(t) \, x(t+\tau)\right]$

$\Rightarrow \quad E(x^3) = \int x^3 f_x(x)\,dx.$

111ly $\quad E(A_k^2) = \sum_{k=1}^{2} (P_k A_k^2)$

$= P(A_k = 0)\, 0^2 + p(A_k = a) a^2$

$= \frac{1}{2} 0^2 + \frac{1}{2} a^2 = \frac{a^2}{2}.$

111ly $\quad R_A(1) = E(A_k \, A_{k+1})$

we also know $R_A(1) = R_A(-1)$ since it's symmetric.

So $E(A_k A_{k+1})$ Possible combination — 00, 0a, a0, aa.

A combination, $\sum P(A_k A_{k+1}), \; A_k A_{k+1}$

Than this is the joint probability of A_k, A_{k+1}

$p(A_k A_{k+1}) = \frac{1}{4}$ because $P(A_k) * P(A_{k+1})$

$= \frac{1}{2} * \frac{1}{2}$

$\Rightarrow \qquad = \frac{1}{4}.$

$R_A(1) = \sum A_k A_{k+1} \; P(A_k A_{k+1})$

for a0 $\quad = a \cdot 0 \cdot \frac{1}{4} = 0$

0a $\quad = 0 \cdot a \cdot \frac{1}{4} = 0$

00 $\quad = 0 \cdot 0 \cdot \frac{1}{4} = 0$

aa $\quad = a \cdot a \cdot \frac{1}{4} = a^2/4.$

Summing all this = ~~$R_A(1) = a^2$~~
$R_A(1) = a^2/4$

then $R_A(-1) = a^2/4$

$R_A(2)$

Autocorrelation A_k, A_{k+2}

1st 3rd bit — possibility
0 0
a 0
0 a
a a

so similar results are obtained

Thus $R_A(n) = \frac{a^2}{4}$ for $n \neq 0$.

$R_A(0) = \frac{a^2}{2}$

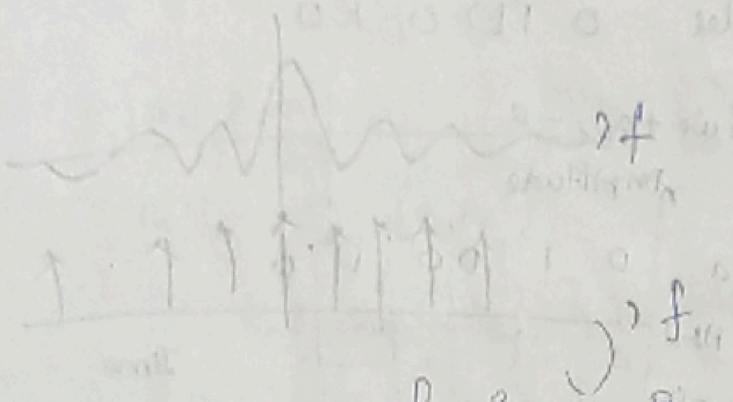
$\Rightarrow |V(f)|^2 = T_b^2 \operatorname{sinc}^2(\pi f T_b)$

$S_{gg}(f) = \frac{1}{T_b} |V(f)|^2 \sum_{n=-\infty}^{\infty} R_A(n) e^{-j2\pi n f T_b}$

$S_{gg}(f) = \frac{1}{T_b} T_b^2 \operatorname{sinc}^2(\pi f T_b) \left[\frac{a^2}{2} + \frac{a^2}{4} \sum_{n=-\infty}^{\infty} e^{-j2\pi f n T_b} \right]$

$= \frac{a^2}{4} T_b \operatorname{sinc}^2 \pi f T_b \{$

Consider two functions of freq



When 0 in sinc, value in ~~the~~ ~~impulse~~ ~~st step~~

value in step, 0 in step

Except at $f(0)$, = value

PSD of ~~an~~ unipolar NRZ waveform

Bandwidth = $\frac{1}{T_b}$ DC component.

$$PSD = \frac{a^2}{4} T_b \, \text{sinc}^2(f T_b) + \frac{a^2}{4} \delta(f)$$

19/08/2020

Unipolar no-return zero

Input data :

$$\sum_{k=-\infty}^{\infty} A_k \, \delta(t - kT_b)$$

NRZ encoded $g(t)$

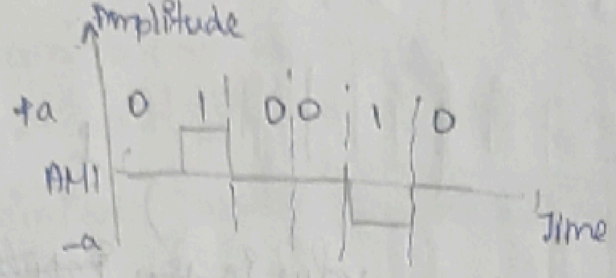
$$= \sum_{k=-\infty}^{\infty} A_k \, v(t - kT_b)$$

17/8/20

Bipolar Non return to zero

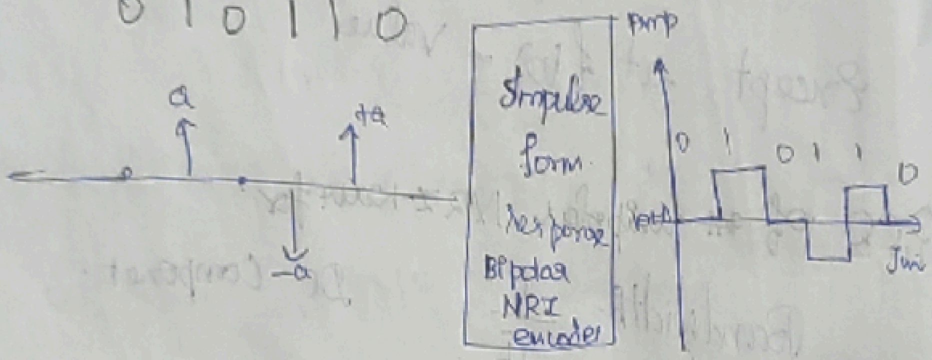
Line coder 0 1 0 0 1 0

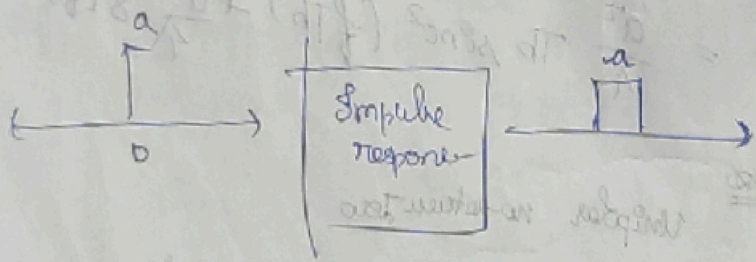
Alternative $+a$, $-a$



Data

0 1 0 1 1 0





Input data

$$x(t) = \sum_{k=-\infty}^{\infty} A_k\, \delta(t-kT_b)$$

NR encoded waveform

$$g(t) = \sum_{n=-\infty}^{\infty} A_k\, V(t-kT_b)$$

Power Spectral Density:-

$$S_{gg}(f) = \frac{1}{T_b} |V(f)|^2 \sum_{n=-\infty}^{\infty} R_A(n)\, e^{-j2\pi f T_b}$$

Autocorrelation equation:

0 0 1 0 ... etc

$R_A(n) = E[A_k A_{k+n}]$

$R_A(0) = \Sigma (A_k^2) = \sum_{k=1}^{2} [P_k A_k^2]$

$= \frac{1}{2} 0 + \frac{1}{4} a^2 + \frac{1}{4} -a^2$

$= \cancel{\dots} = a^2/2$

IIIly:-

$R_A(1) = \Sigma [A_k A_{k+1}]$ $\qquad R_A(1) = R_A(-1)$

Possible combinations:- $\cancel{00}$ $\;$ $-a\,0$ $\;$ $a\,-a$
$0a$ $\;$ $0\,-a$ $\;$ $-a\,a$
$a0$ $\;$ $a\,a$ $\;$ $-a\,-a$.

$p(A_k A_{k+1})$

$R_A(1) = \Sigma A\, A_{k+1}\; p(A_k \cdot A_{k+1})$

$\Rightarrow 0\ 0 \cdot 1/4 = 0$

$0a \cdot 1/8 = 0$

$0.-a \cdot 1/8 = 0$

$a0 \cdot 1/8 = 0$

$-a0 \cdot 1/8 = 0$

$\cancel{a\ a}$

$a\ -a \times \frac{1}{\cancel{16}8} = -a^2/\cancel{16}8$

$-a\ a \times \frac{1}{\cancel{16}8} = -a^2/\cancel{16}8$

$\cancel{-a\ -a}$

$R_A(1) = -a^2/\cancel{4}$ $\qquad R_A(-1) = a^2/\cancel{4}$

$R_A(2) = E(A_k, A_{k+2}) = -a^2/\cancel{4}$

$|V(f)|^2 = T_b^2 \sin^2 c(\pi f T_b)$

$S_{gg}(f) = \frac{1}{T_b} |V(f)|^2 \sum_{k=-\infty}^{\infty} R_A(n)\, e^{-jn\pi f T_b}$

$$= \frac{1}{T_b} T_b^2 \, sinc^2(\pi f T_b) \left[\frac{a^2}{2} + \frac{-a^2}{48} \sum_{n=-\infty}^{\infty} e^{-j2\pi f n T_b} \right]$$

$$\cancel{= \frac{a^2}{2} T_b \, sinc^2(\pi f T_b) - \frac{a^2}{84} \delta(f)}$$

So, probability $\Rightarrow \frac{1}{4} \cdot \frac{1}{2} = \frac{1}{8}$, because it's a

conditional probability $P(A_k) \cdot P\left(\frac{A_{k+1}}{A_k}\right)$

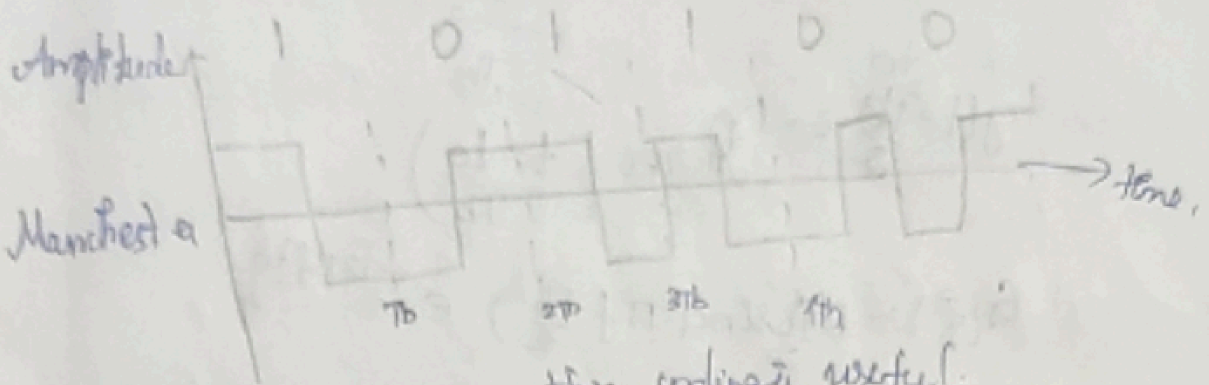
$$\frac{1}{4} \cdot \frac{1}{2}$$

~~20/9/20~~ $R_a(n) = 0$.

$S_{gg}(f) =$

20/08/2020

Manchester Coding:-

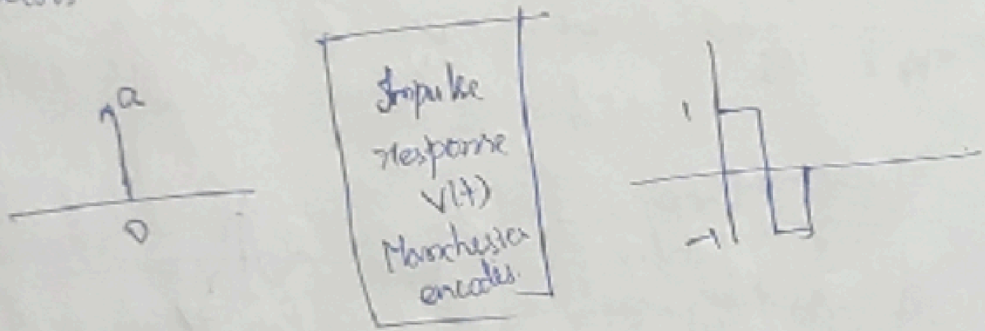


As transitions occurs, this coding is useful. But it is important to know starting point

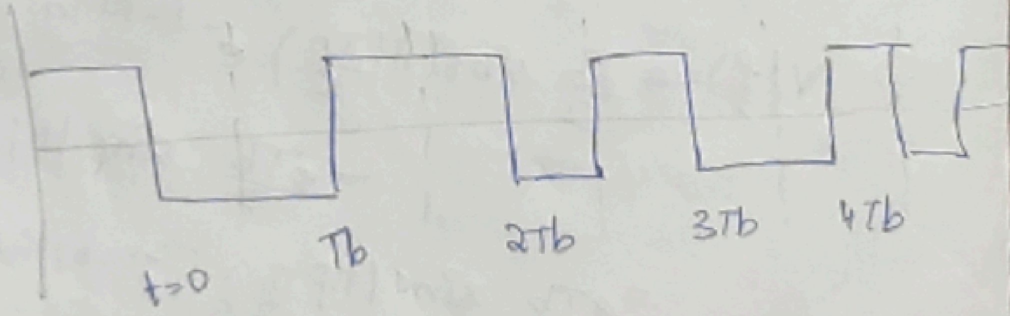
I determine by sampling, +, − ⇒ 1
−, + ⇒ 0

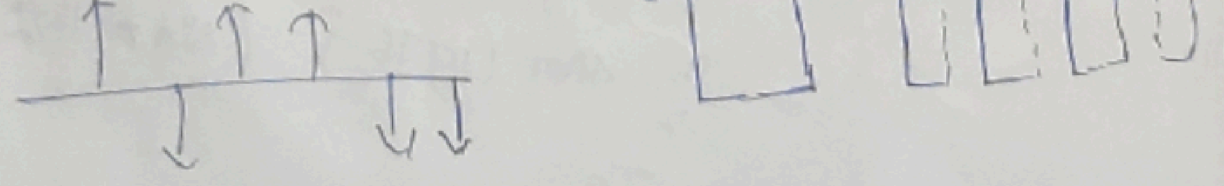
But if there is a delay, then tapping signal is hard.

But in Manchester coding, + = − samples that enables to detect errors.

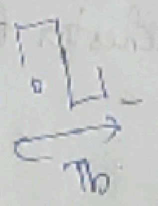


10 1100





$V(f) = ?$

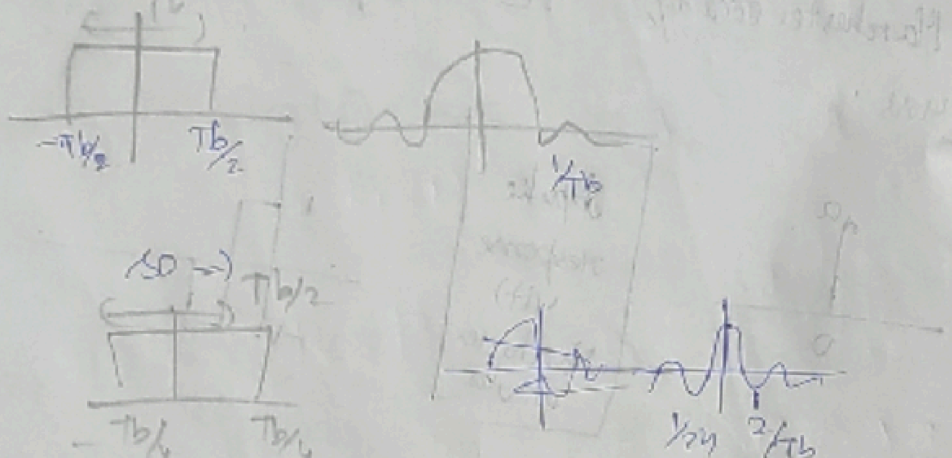


$$g\left(t - \frac{T_b}{2}\right) = \text{rect}\left(\frac{t - t_0}{T_b}\right)$$

$$G(f) = \frac{T_b}{2}\,\text{sinc}\left(\pi f \frac{T_b}{2}\right) e^{-j2\pi f \frac{T_b}{4}}$$

Let $\sqcap \Rightarrow p(t)$ $\Rightarrow \sqcap\!\sqcup \Rightarrow p\left(t + \frac{T_b}{4}\right) - p\left(t - \frac{T_b}{4}\right)$

$$v(t) = p\left[t + \frac{T_b}{4}\right] - p\left[t - \frac{T_b}{4}\right]$$



$$\Rightarrow V(f) = \frac{T_b}{2}\,\text{sinc}\left(\pi f \frac{T_b}{2}\right) e^{-j2\pi f t_0}$$

$$- \frac{T_b}{2}\,\text{sinc}\left(\pi f \frac{T_b}{2}\right) e^{j2\pi f \frac{T_b}{2}}$$

$$= \frac{T_b}{2}\,\text{sinc}\left(\pi f \frac{T_b}{2}\right) \left[\frac{e^{-j2\pi f T_b/2} - e^{j2\pi f T_b/2}}{2}\right]$$

$$= \frac{T_b}{2}\,\text{sinc}\left(\pi f \frac{T_b}{2}\right) \left(\sin \pi f T_b/2\right)$$

$$= T_b$$

$R_A(0) = ?$

$R_A(n) = E[A_k A_{k+n}]$

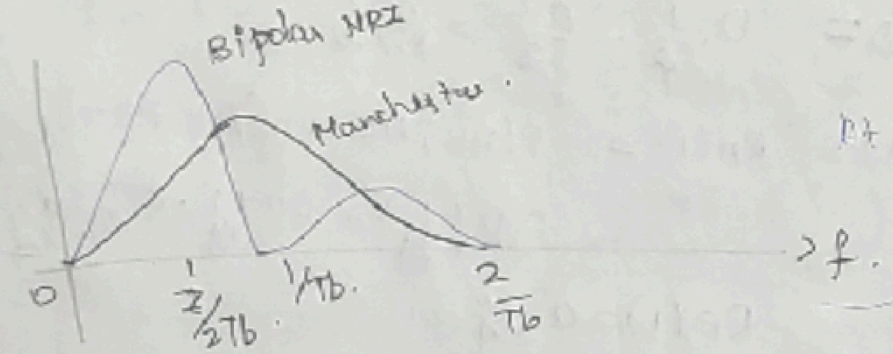
$R_A(n) = E[A_k^2] = \sum_{k=1}^{2} [P_k A_k^2]$

$R_A(0) = \frac{1}{2} \cdot a^2 + \frac{1}{2} \cdot a^2$

$R_A(0) = a^2$

$R_a(n) = 0$

$S_{gg}(f) = a^2 T_b \, \mathrm{sinc}^2 f(T_b) \sin^2(\pi f T_b)$ in Bipolar NRZ

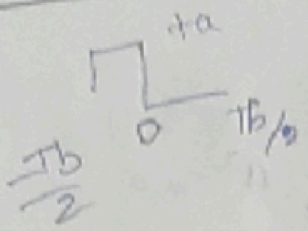


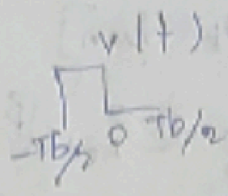
At first zero $\pi f \frac{T_b}{2} = \pi/2$

Manchester

$S_{xx}(f) = T_b a^2 \, \mathrm{sinc}^2\left(\frac{\pi f T_b}{2}\right) \sin^2\left(\frac{\pi f T_b}{2}\right)$

Unipolar RZ





$$V(f) = \frac{T_b}{2} \operatorname{sinc}\left(f \frac{T_b}{2}\right) e^{j2\pi f \frac{T_b}{4}}$$

$$|V(f)|^2 = \frac{T_b^2}{4} \operatorname{sinc}^2 f\left(\frac{T_b}{2}\right)$$

$$R_A(0) = 0 \cdot \frac{1}{2} \cdot \frac{a^2}{2} = a^2/2$$

$$R_A(n) = R_A(1) = P(A_k, A_{k+1})(A_k \cdot A_{k+1})$$

$$= 0 \cdot 0\left(\frac{1}{4}\right) + 0 \cdot a\left(\frac{1}{4}\right) + a \cdot 0\left(\frac{1}{4}\right) \cdot \frac{a^2}{4}$$

$$\boxed{R_A(1) = a^2/4.}$$

$$S_{YY}(f) = \frac{1}{T_b}|V(f)|^2 \sum R_A(n)\, e^{-j2\pi n f T_b}$$

$$= \frac{1}{T_b}\left[\frac{T_b^2}{4} \operatorname{sinc}^2 f \frac{T_b}{2} \cdot \left[\frac{a^2}{2} + \frac{a^2}{4} \delta(f) T_b^2\right]\right.$$

$$= T_b \operatorname{sinc}^2 \underline{a^2}$$

Polar RZ

$-T_b/_{4}$ 0 $T_b/2$

$\Pi \rightarrow \frac{T_b}{2} \, sinc\left[f \frac{T_b}{2}\right]$

$$v(t) \xrightarrow{F} V(f) = \left(e^{-j2\pi f \frac{T_b}{4}}\right) \frac{T_b}{2} \, sinc\left[f \frac{T_b}{2}\right]$$

$$V(f)^2 = \frac{T_b^2}{4} \, sinc^2 \, f \, T_b/2 .$$

$R_A(0) = a^2$

$R_A(n) = 0 \qquad n \neq 0$

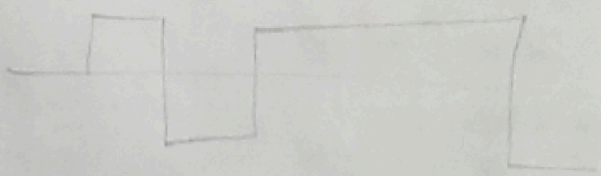
$$S_{gg}(f) = \frac{1}{T_b} \, |V(f)^2| \, \Sigma R_A(k) \, e^{-j2\pi f \frac{T_b}{2} n}$$

$$= \frac{1}{T_b} \, \frac{T_b^2}{4} \, sinc^2\left(f \frac{T_b}{2}\right) \cdot a^2$$

$$\boxed{S_{yy} = \frac{a^2 T_b}{4} \, sinc^2\left(f \frac{T_b}{2}\right)}$$

Polar NRZ

0	Tb	2Tb.	3Tb	4Tb.
1	0	1	1	0
↑a+	↓-a	↑+a	↑+a	↓-a.



$V(f) = T_b \, sinc \, f(T_b)$

$R_A(0) = \frac{1}{2} a^2 + \frac{1}{2} a^2 = a^2$

$R_A(1) \Rightarrow$

A_k	A_{k+1}	$A_k A_{k+1}$	$P(A_k A_{k+1})$	Product
$+a$	$+a$	a^2	$1/4$	$\frac{a^2}{4}$
$+a$	$-a$	$-a^2$	$1/4$	$a^2/4$
$-a$	$+a$	$-a^2$	$1/4$	$-a^2/4$
$-a$	$-a$	a^2	$1/4$	$-a^2/4$

$R_A(1) = 0 \quad \Rightarrow R_A(n) = 0$

$$S_{YY}(f) = \frac{1}{T_b} |V(f)|^2 \sum_{n=-\infty}^{\infty} R_A(n) e^{-j2\pi f n T_b}$$

$$= \frac{1}{T_b} T_b \, \mathrm{sinc}^2(f T_b) \, a^2$$

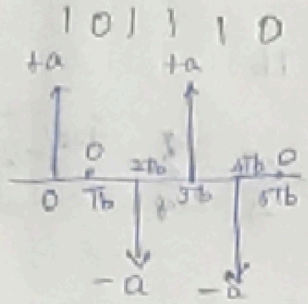
$$\boxed{S_{YY}(f) = a^2 T_b \, \mathrm{sinc}^2(f T_b)}$$

22/08/2020

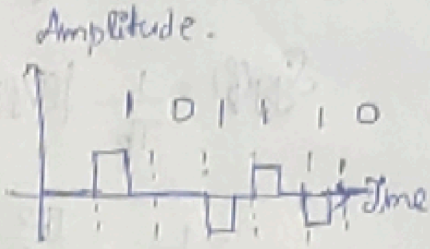
S. Sai Jayasree
2018504575

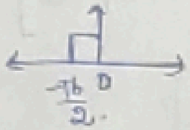
Power Spectral Density of Bipolar Return to Zero:

Consider Bipolar line code



Impulse response Encoder for BIPRZ





$\Rightarrow V(f) = \frac{Tb}{2}\ \text{sinc}\left(f\ \frac{Tb}{2}\right)$

Autocorrelation sequence $R_A(n)$

$R_A(n) = E(A_k\, A_{k+n})$

$R_A(0) = E(A_k A_k) = E(A_k^2) = \sum (P_k A_k^2)$

$= \frac{1}{2}(0) + \frac{1}{4}(a^2) + \frac{1}{4}(a^2) = a^2/2$

$\boxed{R_A(0) = \frac{a^2}{4}}$

$R_A(1) = E(A_k\, A_{k+1})$

A_k	A_{k+1}	$A_k A_{k+1}$	$P A_k . A_{k+1}$	$(A_k A_{k+1}) . P(A_k . A_{k+1})$
0	0	0	$\frac{1}{2}.\frac{1}{2} = 1/4$	0
0	+a	0	1/8	0
0	-a	0	1/8	0
+a	0	0	1/8	0
+a	+a	a^2	~~1/4~~ 0	0
+a	-a	$-a^2$	1/8	$-a^2/8$
-a	a	~~0~~	1/8	0
-a	+a	$-a^2$	1/8	$-a^2/8$
-a	-a	$+a^2$	0	0

$R_n(1) = \sum A_k A_{k+1} P(A_k \cdot A_{k+1}) = -a^2/4$

$R_n(-1) = R_n(1)$

$R_n(2) = 0$ since for polar $R_n(2) = 0$

$$S_{gg}(f) = \frac{1}{T_b} |V(f)|^2 \sum R_n(n) e^{-j2\pi f T_b}$$

$$= \frac{1}{T_b} \frac{T_b^2}{4} \operatorname{sinc}^2\left(f \frac{T_b}{2}\right) \left[\frac{a^2}{2}(1) - \frac{a^2}{4} \sum_{-1}^{+1} e^{-j2\pi f n T_b}\right]$$

$$= \frac{T_b}{4} \operatorname{sinc}^2\left[f\frac{T_b}{2}\right] \left[\frac{a^2}{2} - \frac{a^2}{4} \times 2\cos(2\pi f T_b)\right]$$

$$\boxed{S_{gg}(f) = \frac{a^2}{4} T_b \operatorname{sinc}^2\left(f\frac{T_b}{2}\right) \sin^2\left(\pi f \frac{T_b}{2}\right)}$$

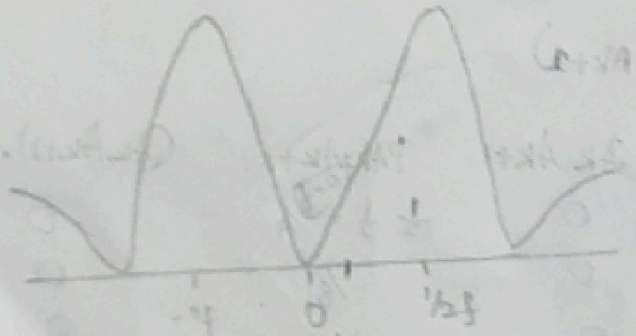
⇒ Here there is no DC components.

⇒ There is no clock component to extract at the end of receiver. Therefore not suitable for clock extraction.

⇒ Bipolar RZ has higher bandwidth has other polar RZs

$T_b = 1$

$a = a$



⇒ Since no constants DC component = 0

⇒ Bandwidth of the waveform = $1/T_b$ compared to

⇒ manchester, this is less

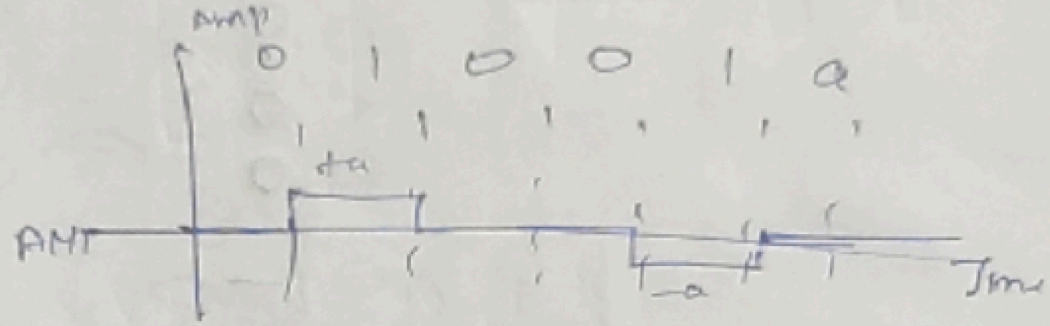
⇒ This isn't useful / suitable for clock extraction

Power requirements:

$$\text{Power} = \frac{1}{2T_b} \sum |x(t)|^2$$

26/08/2020

Bipolar [AMI] NRZ:



NRZ ⇒ $\cancel{\frac{1}{T_b}}$ T_b | f↑ B.W↓

RZ ⇒ $\frac{T_b}{2}$

0 0 0 1 1 00 00000000000 1 1 00 000

B4NS 00 0 1 1 [B00V] [B00V] [B00V] 1 1 [B0 0V] 0

B2NS 000 1 1 (B0V) (B 0V) (B0V) (B0V) 1 1 (B0 V) 00

B6NS 00.0 1 1 (0VB0 VB) (0V B0VB) 1 1 00 000

B8NS 0 00 1 1 (000V B0 V B) 0000 1 1 00000 0

1 000 0 00 1 1 0 1 0 00 00000 00 0000 1 1 1

B4ZS 1 (000 V) 00 1 1 0 1 0 (B0 0V) (B0 0 V) (B00V) 1 1 1

B3ZS 1 (0 0V) (0 0 V) 1 1 0 1 0 1 (B0 V) (B 0V) (B0 V) (B0 V) 1 1 1

B6ZS 1 (0 BV 0 V B) 1 1 0 1 0 1 (0BV0VB) (0BV 0VB) 1 1 1

B8ZS 1 0 00 0 00 1 1 01 01 (0000BV0 BV0000) 1 1 1

0010

B4ZS 0010

B3ZS 0010

B6ZS 0010

B8ZS 0010

27/8/2020

We assume only one path between transmitter and Reciever. Single path effect only

Impulse Response $h(t)$ Frequence response $H(f)$

Transmitter $\xrightarrow{\text{Channel}}$ Reciever

- AWGN [Additive White Gaussian Noise] Channel
  - ⇒ Noise is white has all freq
  - ⇒ Noise is Gaussian
  - ⇒ Added at front end of reciever.
- Channel Attenuation
- Bandwidth limitation

Channel Limitations :-

- Attenuation ⇒ only power change shape remains same.

Bandwidth Limitation:- shape may change

(i) Signal distortion:

Recieved Pulse shape is different [Broader than] the transmitted that of pulse

This is called Intersymbol Interference.

Soln:- Channel Equalization

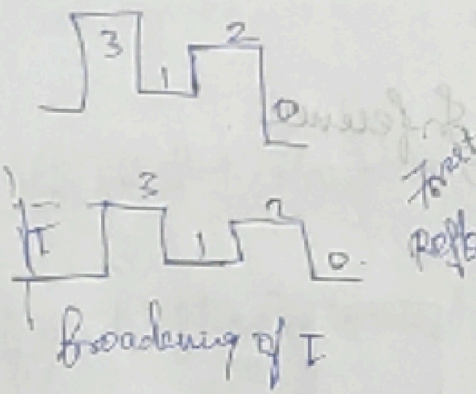
Wireless Communication:-

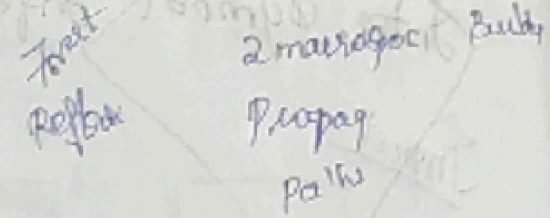
⇒ The broadening will be due to Multipath effect

ISI due to Multipath effect in Wireless Channel:-

Signal recieved at shorter path

Signal recieved over longer path



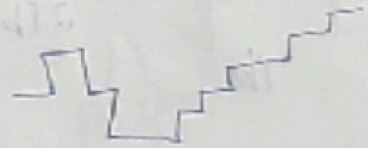


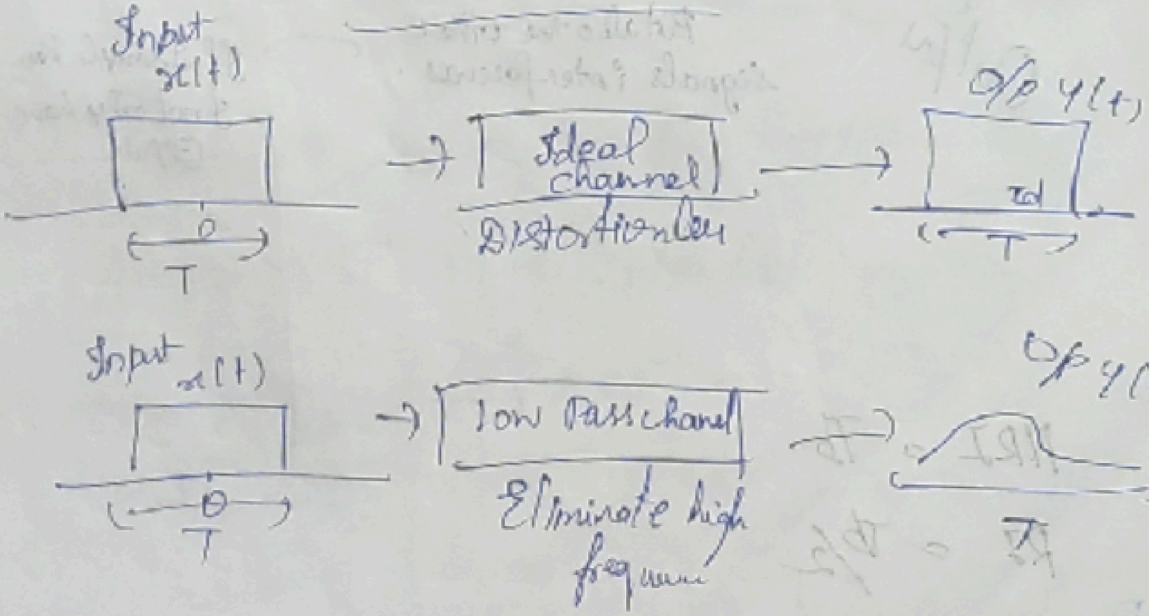
Reciever signal

Superposition of two

Channel Width Bandwidth Limitation

Pulse Broadening





The Broadening is because the

compression in time domain } $\longrightarrow$ Expansion/Broadening in freq domain

$\Downarrow$

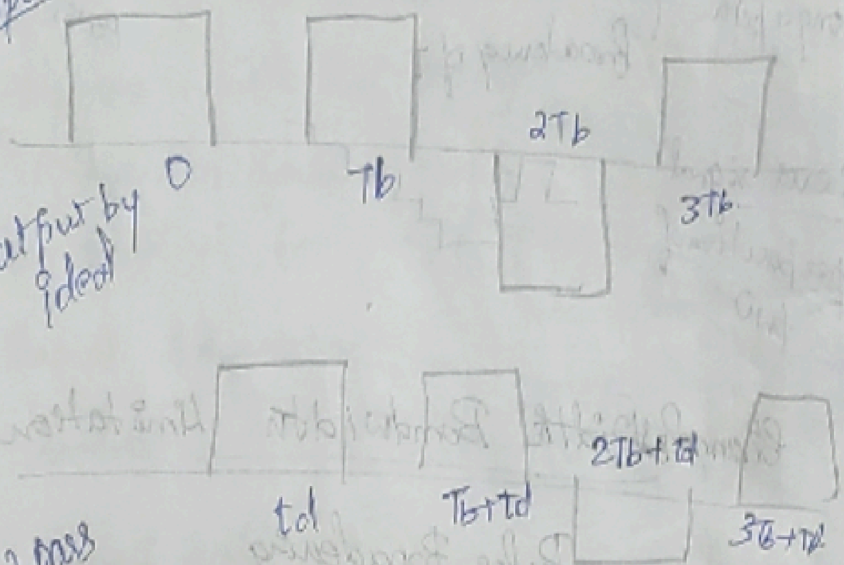
Due to distortion.

Inter Symbol Inference

Input

Output by ideal

Low pass Output



But also the other signals interference ← If I sample then I not only have ③ pulse

NRZ = Tb

RZ = Tb/2

Wired channel always have a tone restriction

Wireless will have filters which will broadens the pulse.

So No matter what, Broadening is inevitable.

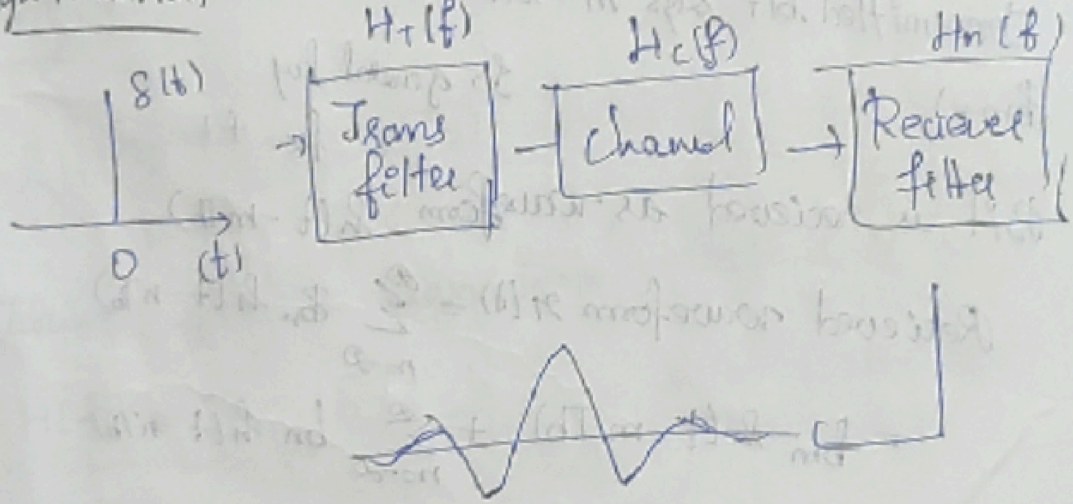
Bit rate ↓ Problem ↓

Solution:- Channel Estimation & Channel Equalization

⇓

Obtaining h(f) of a channel.

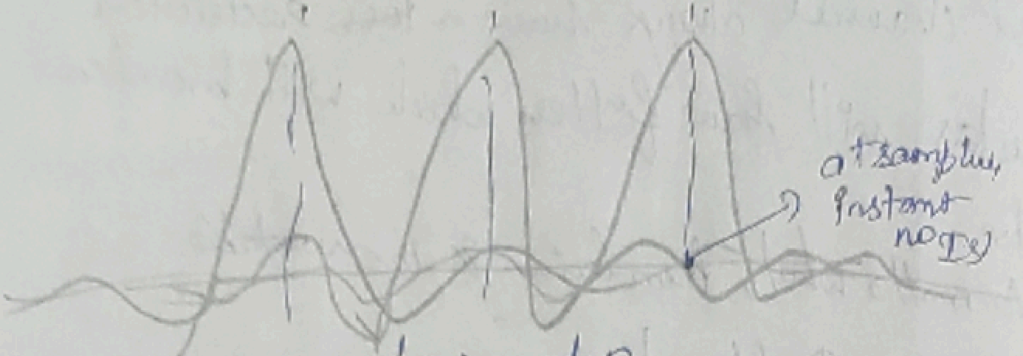
Equalization



Composite system Transfer function = $H(f)$

$$H(f) = H_T(f) \cdot H_C(f) \cdot H_R(f)$$

$$H(f) \xleftrightarrow{FT} h(t)$$



so no overlapping at 0.

Equalization: Reciever filter compensating the effect of transmit filter and channel but gives rectangular pulse.

Derivation :-

Transmitted bit says, m^{th} bit 1 encoded as +1

$b_m = 1$ $\quad$ In general say $b_k = \pm 1$

Bit is recieved as waveform $h(t - mT_b)$

Recieved waveform $r(t) = \sum_{n=-\infty}^{\infty} b_n \, h(t - nT_b)$

$$= b_m \, h(t - mT_b) + \sum_{\substack{n=-\infty \\ n \neq m}}^{\infty} b_n \, h(t - nT_b)$$

To take decision for m^{th} bit, sampling it at $t = mT_b$

$$r(mT_b) = \underbrace{b_m \, h(mT_b - mT_b)}_{\text{Due to } m^{th} \text{ bit}} + \underbrace{\sum_{\substack{n=-\infty \\ n \neq m}}^{\infty} b_n \, h(mT_b - nT_b)}_{\text{from other bit}}$$

$$r(mT_b) = b_m \, h(0) + \sum_{\substack{k=-\infty \\ k \neq 0}}^{\infty} b_k \, h(kT_b)$$

To avoid ISI,

$y(mT_b) = b_m h(0)$

Equivalently

$$\boxed{\begin{array}{ll} h(kT_b) = 0 & k \neq 0 \\ h(kT_b) \neq 0 & k = 0 \end{array}}$$

since at 0 we are going to decide the fate

This is possible only for impulse

Consider $h_\delta(t)$ = discrete time version of $h(t)$

$$h_\delta(t) = h(t) \sum_{k=-\infty}^{\infty} \delta(t - kT_b) = \sum_{k=-\infty}^{\infty} h(kT_b)\, \delta(t - kT_b)$$

For Zero ISI :-

$$\sum_{k=-\infty}^{\infty} h(kT_b)\, \delta(t - kT_b) = h(0)\,\delta(t)$$

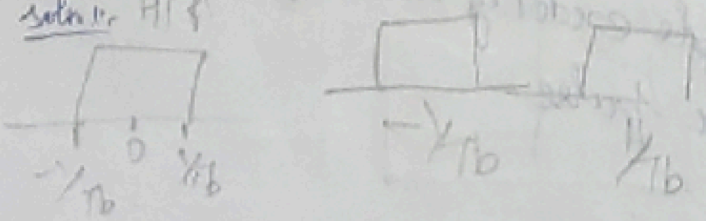
$$h(t) \sum_{k=-\infty}^{\infty} \delta(t - kT_b) = h(0)\,\delta(t)$$

Taking fourier Transform on both sides.

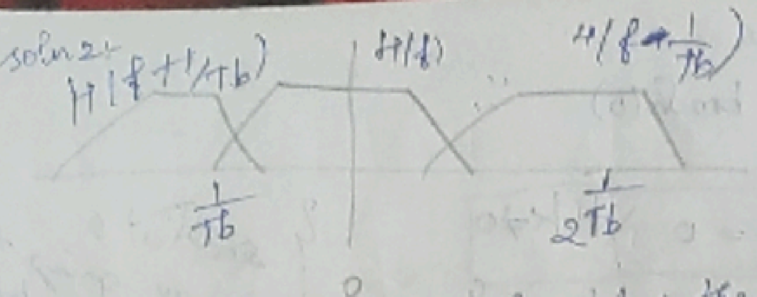
$$H(f) * \sum_{k=-\infty}^{\infty} \delta\left(f - \frac{k}{T_b}\right) = h(0) = \text{constant}.$$

$$\boxed{\sum_{k=-\infty}^{\infty} H\left(f - \frac{k}{T_b}\right) = \text{constant}}$$

soln:- $H(f)$



soln 2:



But theoretically rectangular pulse is the best one because of lower Bandwidth.

Symbol Rate $f(s) = \frac{1}{T_s} = 2W$

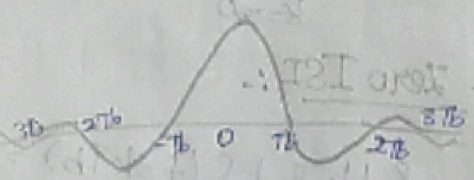
⇓

symbol = 2

Binary system ⇒ Bit rate

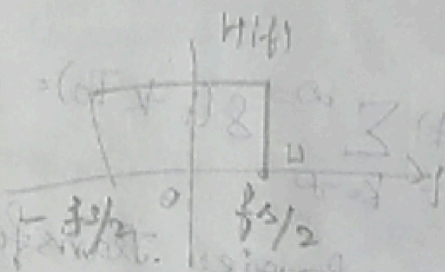
Symbol = 4

M-ary.



Absolute Bandwidth

$W = \frac{f_s}{2}$ [Minimum bandwidth]



$$\boxed{\text{If Bitrate} = R_b = \frac{1}{T_b} \qquad BW = \frac{R_b}{2} = W}$$

Absolute Bandwidth $= \frac{f_s}{2}$

Symbol Rate $= f_s$

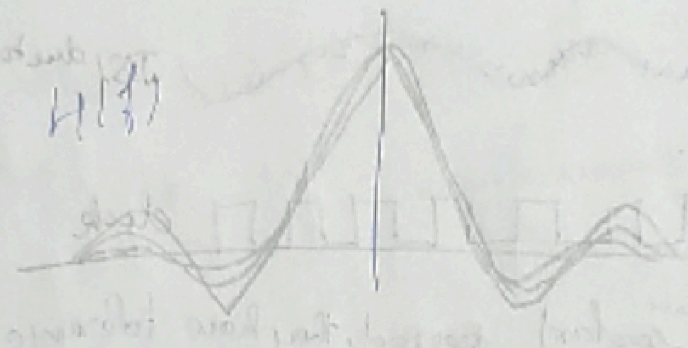
Nyquist signalling rate

Limitation:-

- We cant sample exactly at ISI = 0 in the series of sinc pulse.

I have to compromise. so that we can have some tolerances.

We can use raised cosine pulses



$$h(t) = \frac{\sin \frac{\pi t}{T}}{\pi t / T} \; \cancel{\cos \pi a t} \; \frac{\cos \frac{\pi a t}{T}}{1 - 4a^2t^2/T^2}$$

$a = 0$ ideal solution - $h(t)$ is sinc function

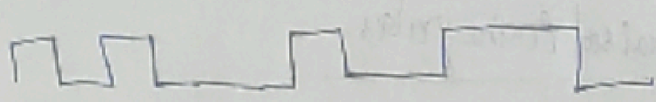
$$\boxed{W = \frac{1}{2} T_b}$$

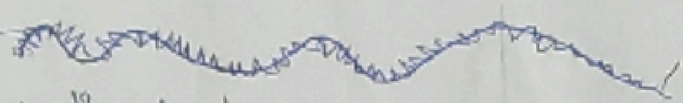
02/09/2020 :

Performance & Bitrate

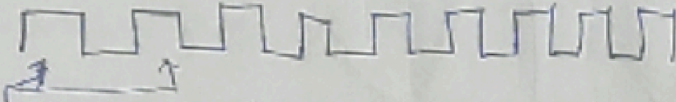
Probability of error ⇒

Eye pattern:-





amp also reduced.

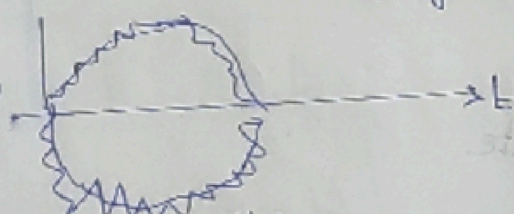


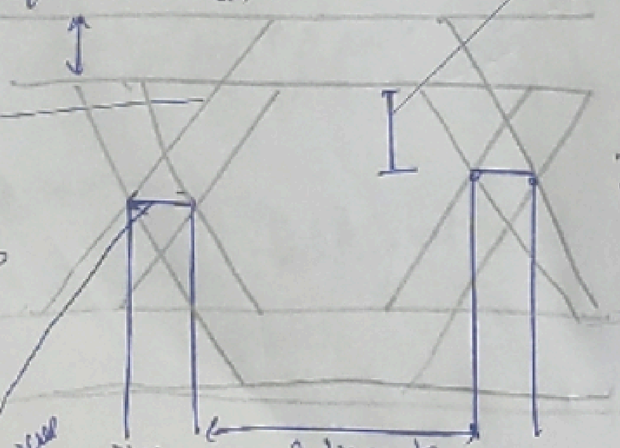
No clock pulse have constant period, they have tolerance

If we trigger the CRO at this point, the first part of wave is traced, second point trigger second part traced

If any change takes place before $1/24$ s, we can't notice it

=> 10Mbps





Sampling is done at falling edge

• W.r.t. the pattern, we can predict the nature. This is qualitative

To exact values, take BER

BER = $P_e = 10^{-6}$ which means for 10^6 bits only one bit is wrong

$= \frac{1}{10^6}$

Power generation

$\int_{-P}^{P} S_{gg}(x) f$ = Power =

$\int \text{sinc}^2(x)\, df = \int |x(t)|^2 dt$

$\Rightarrow$ Average power $= \frac{1}{2T}\left[\int_{-T}^{T} |x(t)|^2 dt\right]$

Energy

03/9/2020 Duobinary Encoding :-

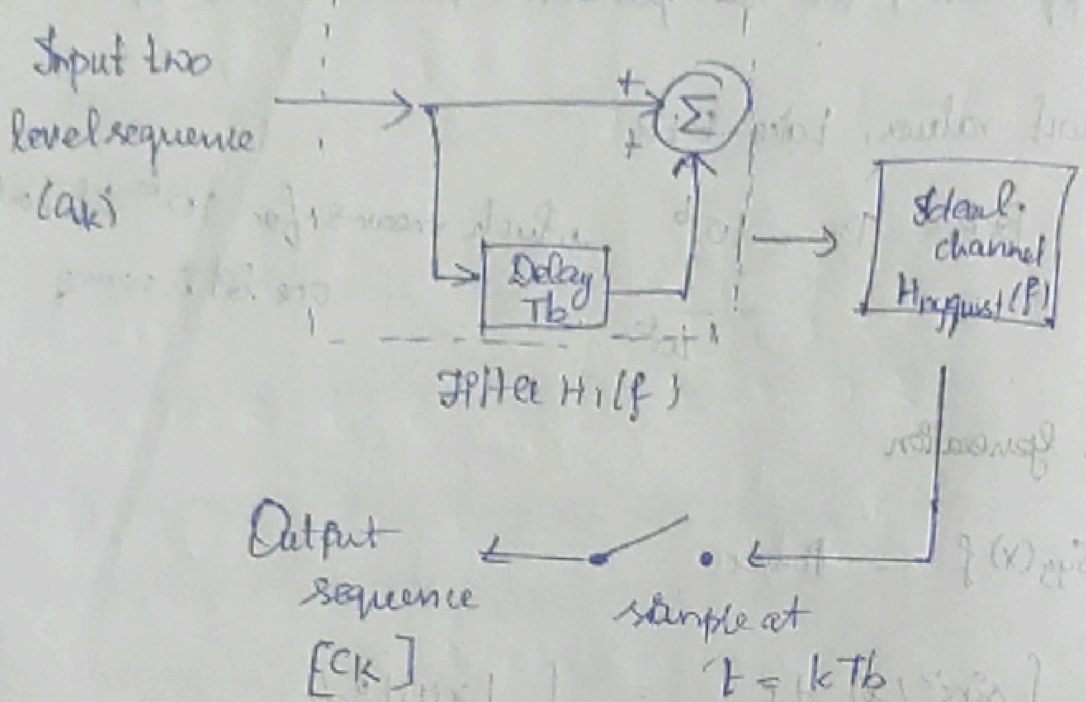
Correlative Encoding

$\Rightarrow$ Controlled ISI → Required minimum Bandwidth

$\longrightarrow \boxed{H_T(f)} \longrightarrow \boxed{H_C(f)} \longrightarrow \boxed{H_R(f)} \longrightarrow$

Recieve filter

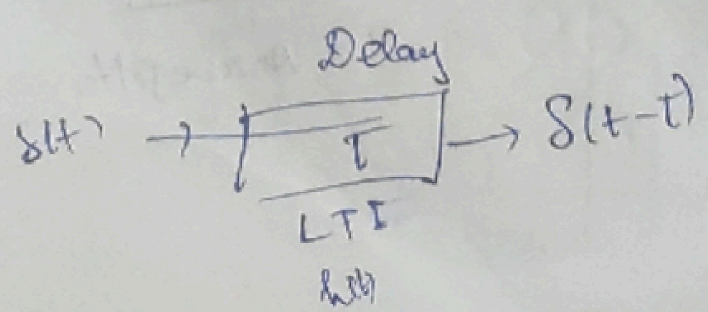
Duobinary signalling scheme



The current bit exactly overlaps / superpositions with the previous bits.

So if $\Sigma \to 2 \to$ then current bit   previous bit

		current bit	previous bit
2	$\to$	1	1
-2	$\to$	-1	
0	$\to$	$0/x$	$0/-x$
1	$\to$	$0/1$	$1/0$
-1	$\to$	$-1/0$	$0/-1$



$H(f) = H_c(f)\left[1 + \exp(-j2\pi f T_b)\right]$

Nyquist ⇒ Ideal filt for bandwidth restriction

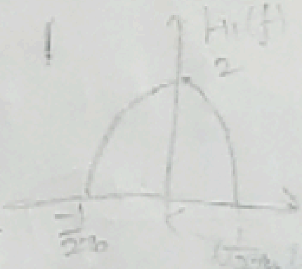
$= H_c(f)\, e^{-j\pi f T_b}\left[e^{j\pi f T_b} + e^{-j\pi f T_b}\right]$

$= H_c(f)\, e^{-j2\pi f T_b}\left[2\cos(\pi f T_b)\right]$

$$\boxed{H(f) = 2H_c(f)\cos(\pi f T_b)\, e^{-j2\pi f T_b}}$$

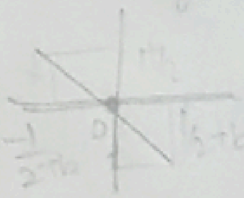
$$H_c(f) = \begin{cases} 1 & |f| \le R_b/2 \\ 0 & \end{cases}$$

$$H(f) = \begin{cases} 2\cos(\pi f T_b)\exp(-j2\pi f T_b) & |f| \le \frac{R_b}{2} \\ 0 & \text{elsewhere} \end{cases}$$

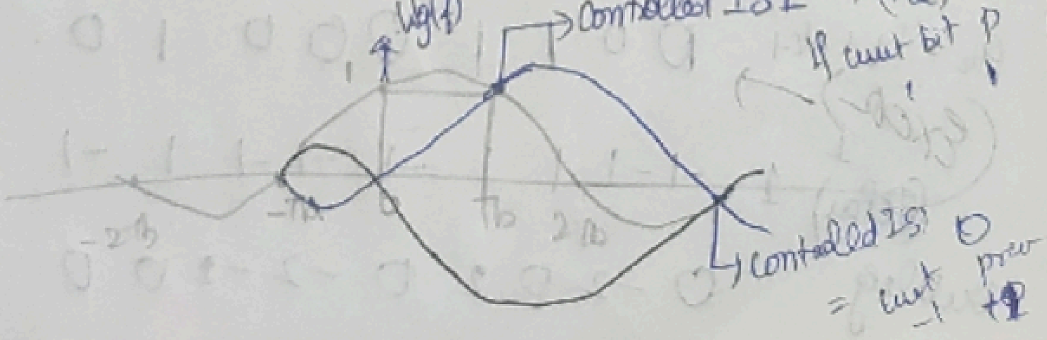


$h(t) = \dfrac{\sin\left(\frac{\pi t}{T_b}\right)}{\pi t/T_b} + \dfrac{\sin\left[\frac{\pi(t - T_b)}{T_b}\right]}{\pi(t - T_b)/T_b}$

$= \dfrac{\sin(\pi t/T_b)}{\pi t/T_b} - \dfrac{\sin(\pi t/T_b)}{\pi(t - T_b)/T_b}$



$= \dfrac{T_b^2 \sin(\pi t/T_b)}{\pi t(T_b - t)}$



Duobinary encoding.

Data seq: 1 0 1 0 1 0 0 1 1

b_k: 1 -1 1 -1 1 -1 -1 1 1

(Polar representation)

Initial +1

$C_k = b_k + b_{k-1}$: 2 0 0 0 0 0 -2 0 2

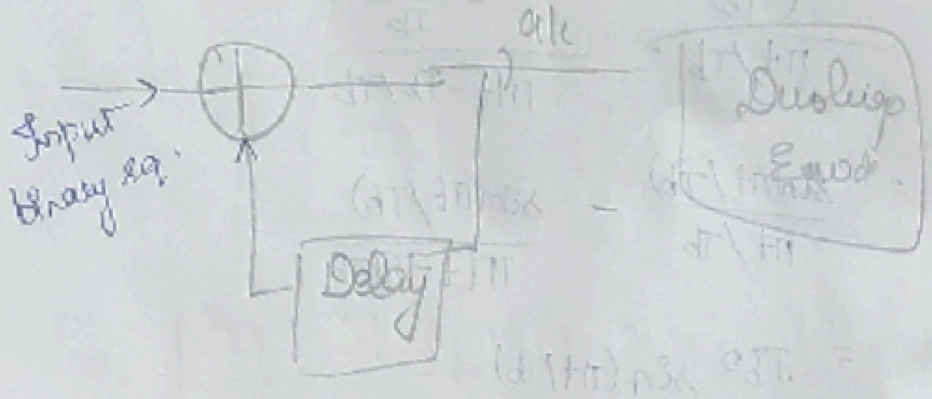
If $C_k > +1$ Data 1

$C_k < -1$ Data 0

$-1 < C_k < 1$ Current bit inverted

1 -1 1 -1 1 -1 -1 1 1

1 0 1 0 1 0 0 1 1



EXOR
01, 10 } 1
00, 11 } 0.

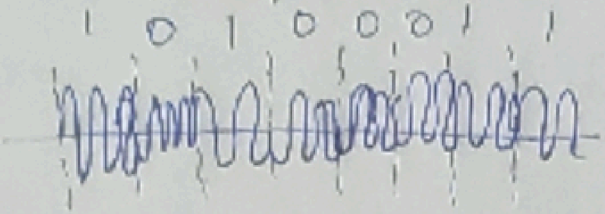
Sequence 1 0 1 0 1 0 0 1 1

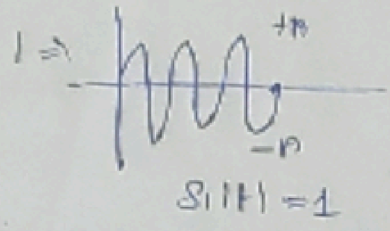
$a_k = b_k + a_{k-1}$ (EXOR): 1 0 0 1 1 0 0 0 1 0.

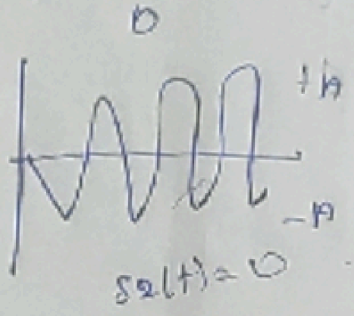
(Polar) 1 -1 -1 1 1 -1 -1 -1 1 -1

Duo seq 0 -2 0 2 0 -2 -2 0 0

09/09/2020 Band pass Signalling:







- $\{ s_i(t) \}$ represents the set of waveforms to be represented in the signal space. $\{ s_1(t), s_2(t) - - - s_M(t) \}$.

The set of orthonormal (real) basis functions

$\{ \phi_1(t), \phi_2(t), \cdots \phi_N(t) \}$ $N \leq M$

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

T = duration of symbol

Energy of the basis function = 1

Let $S_i(t) = \sum_{j=1}^{N} s_{ij} \, \phi_j(t)$

. Each of the waveform, must be a linear combination of the basis orth functions

$0 \leq t \leq T$

. If S received in

$i = 1, 2, 3 \ldots M$

$S_{ij} = \int_0^T S_i(t) \phi_j(t) \, dt$

$\Rightarrow$ $i = 1, 2, \ldots M$
$j = 1, 2, \ldots N$