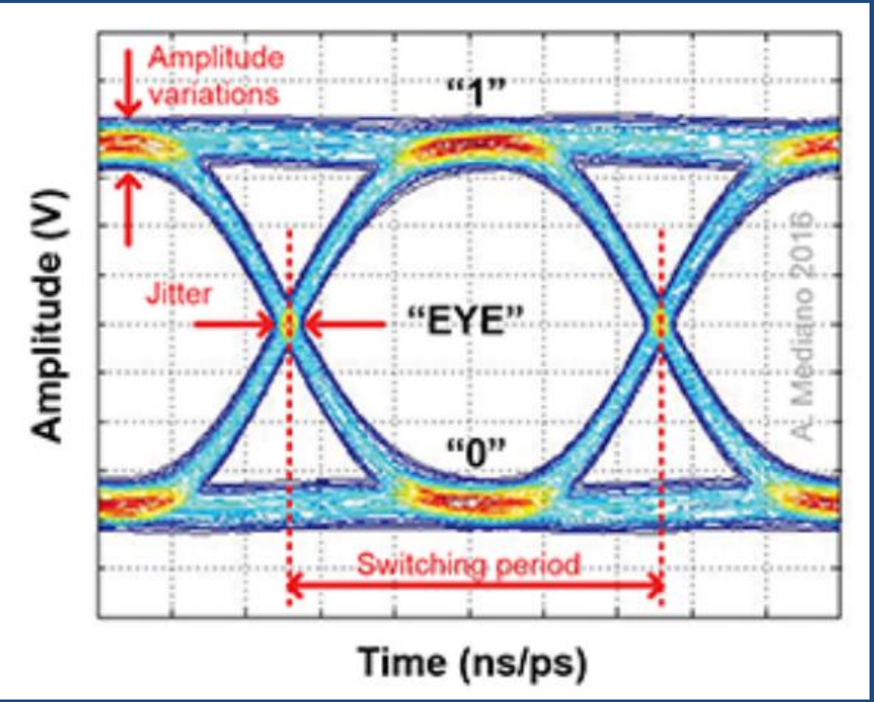
LECTURE 6:
DIGITAL BASEBAND TRANSMISSION



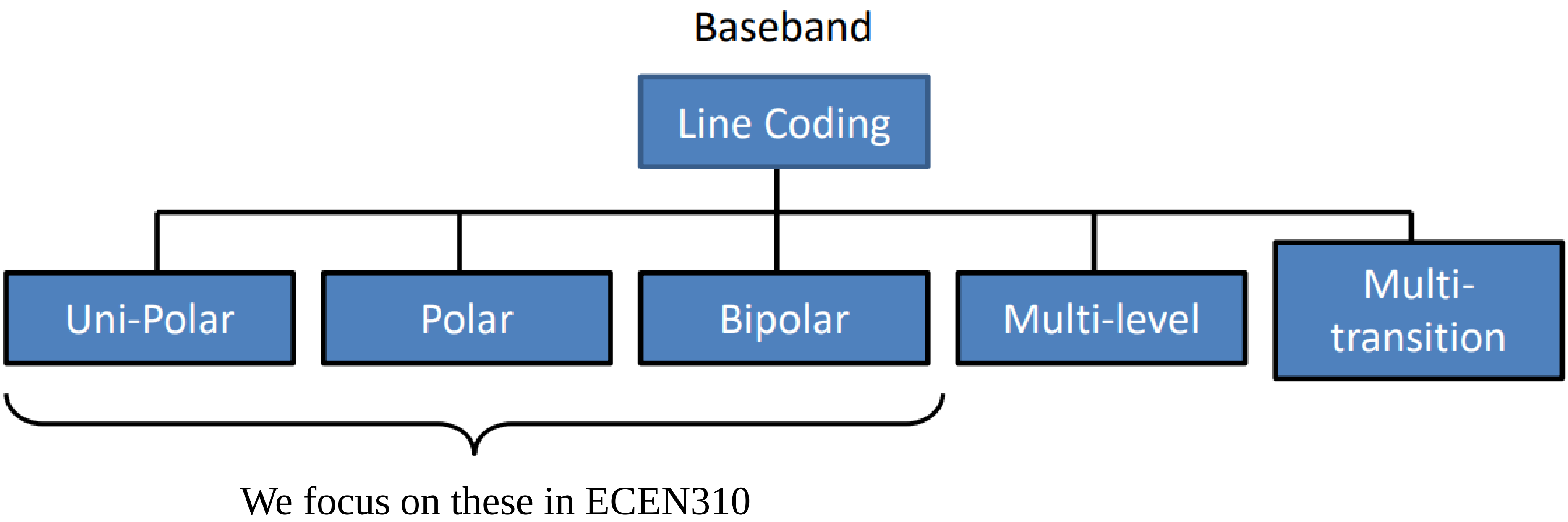
Digital Baseband Transmission

Read Proakis and Salehi, Chapter 10.1 – 10.3

Overview

- Thus far we talked about baseband transmission of digital data by means of pulse modulation (PAM, PCM) - without discussing any degradation along the way
- There are a number of factors contributing to signal degradation, important ones being
  - Intersymbol interference due to imperfections of the channel response
  - Channel noise
- Today we will discuss intersymbol interference (ISI), namely
  1. what is ISI
  2. how to design the pulse shape to mitigate ISI
  3. visualisation of ISI performance by examining eye pattern
- We will talk about *baseband* transmission, but the theory applies to *bandpass* communications as well.

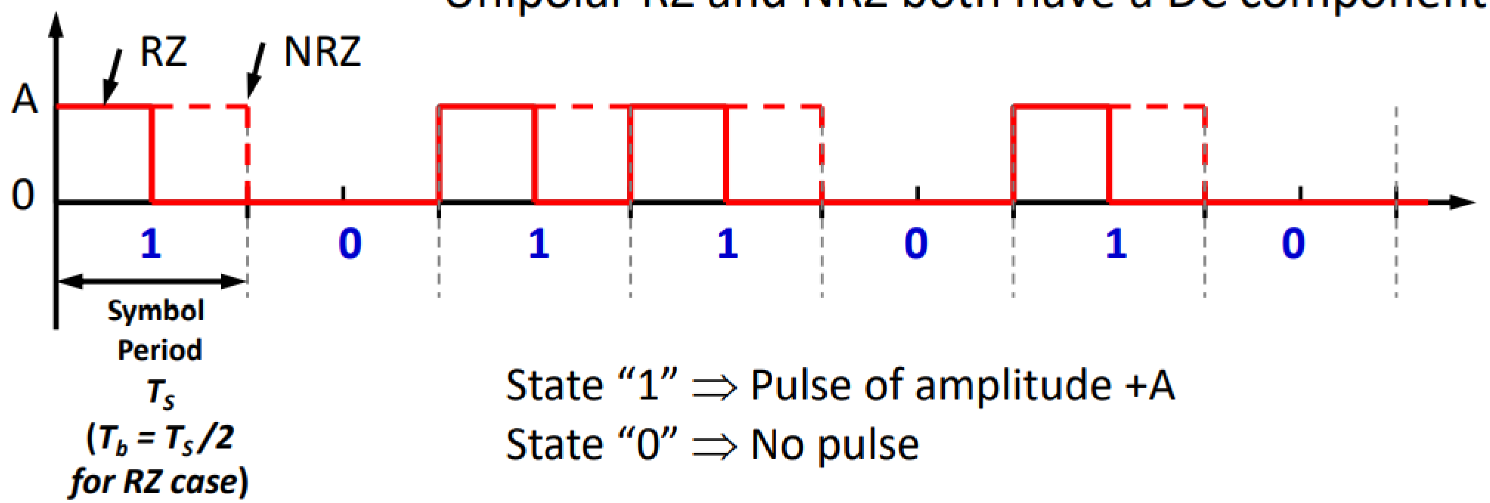
Another look at line coding: baseband



A **line code** is a specific code (with precisely defined parameters) used for transmitting a **digital signal** over a channel. **Line coding** is used in **digital data** transport – the pattern of voltage, current or photons used to represent **digital** data on a transmission link is called **line encoding**.

Binary Unipolar – RZ and NRZ (*aka* "On-Off Keying")

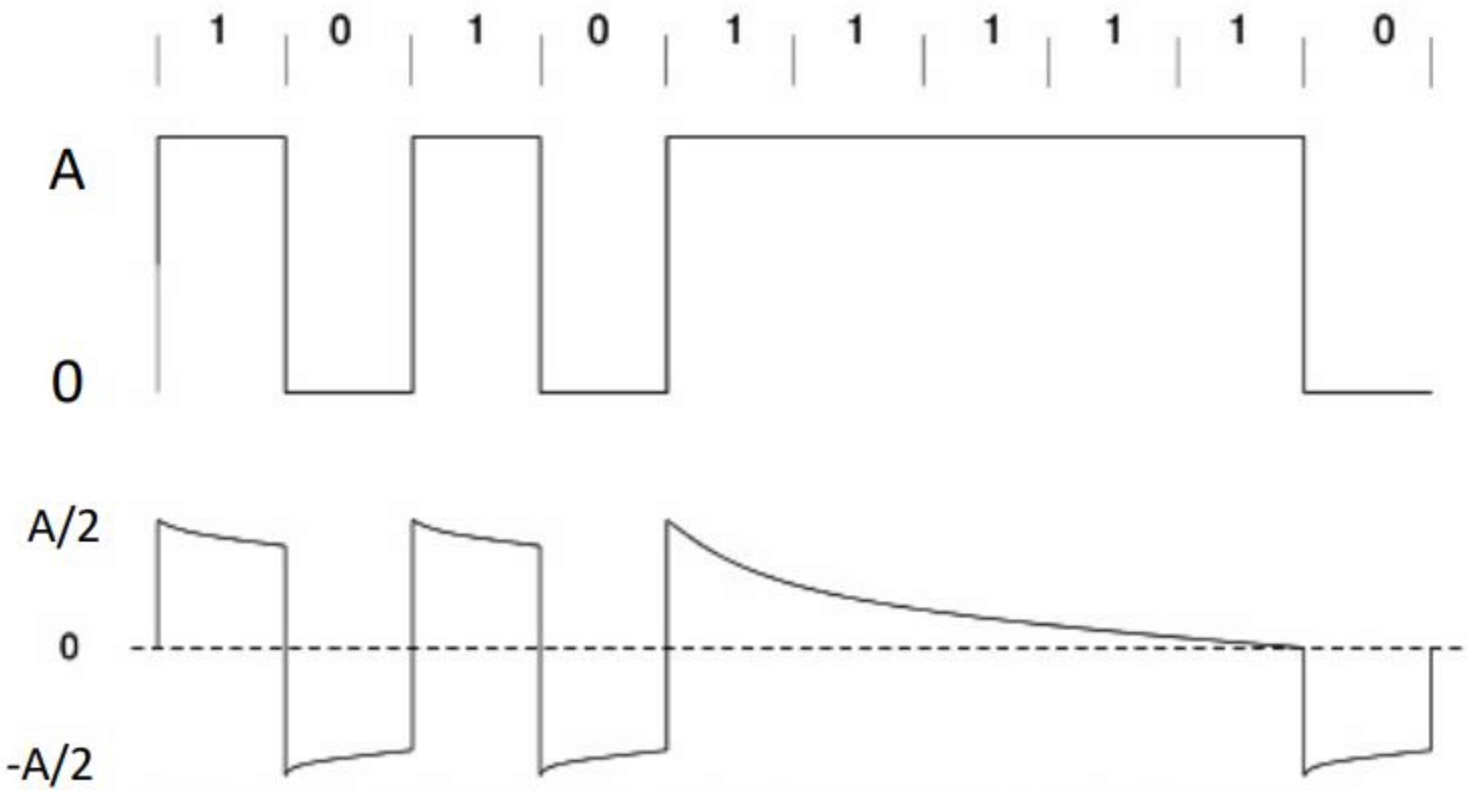
Unipolar RZ and NRZ both have a DC component.



State "1" $\Rightarrow$ Pulse of amplitude +A
State "0" $\Rightarrow$ No pulse

ADVANTAGES	DISADVANTAGES
1. Simplicity 2. Doesn't require a lot of bandwidth	1. Presence of DC level 2. Contains low-frequency components (leads to drooping) 3. No clocking component to synchronize to at receiver 4. Long string of zeros causes loss of synchronization

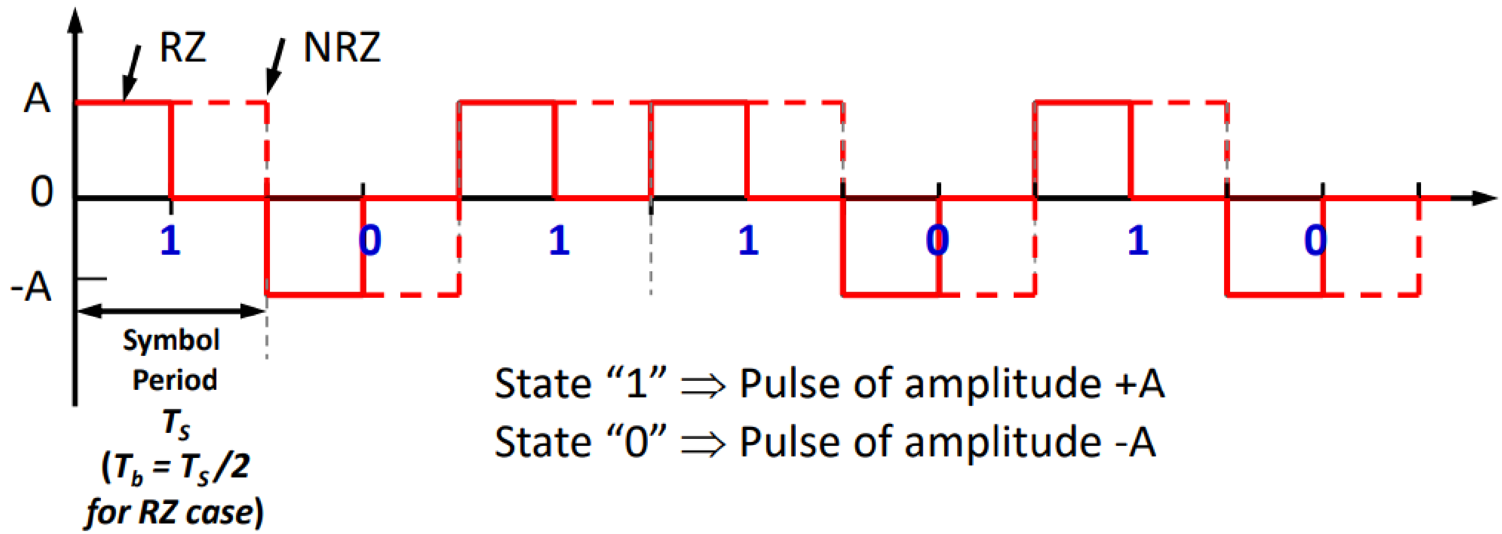
Unipolar NRZ Signal Droop Problem



Signal droop distortion is due to AC coupling

Binary Polar – RZ and NRZ

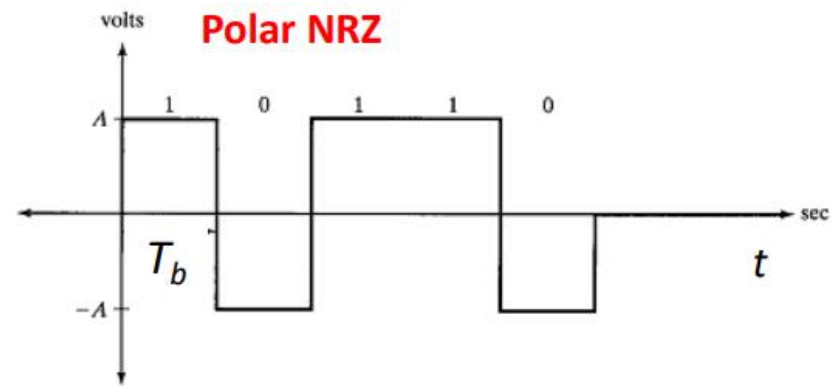
Polar RZ takes twice as much bandwidth as polar NRZ.



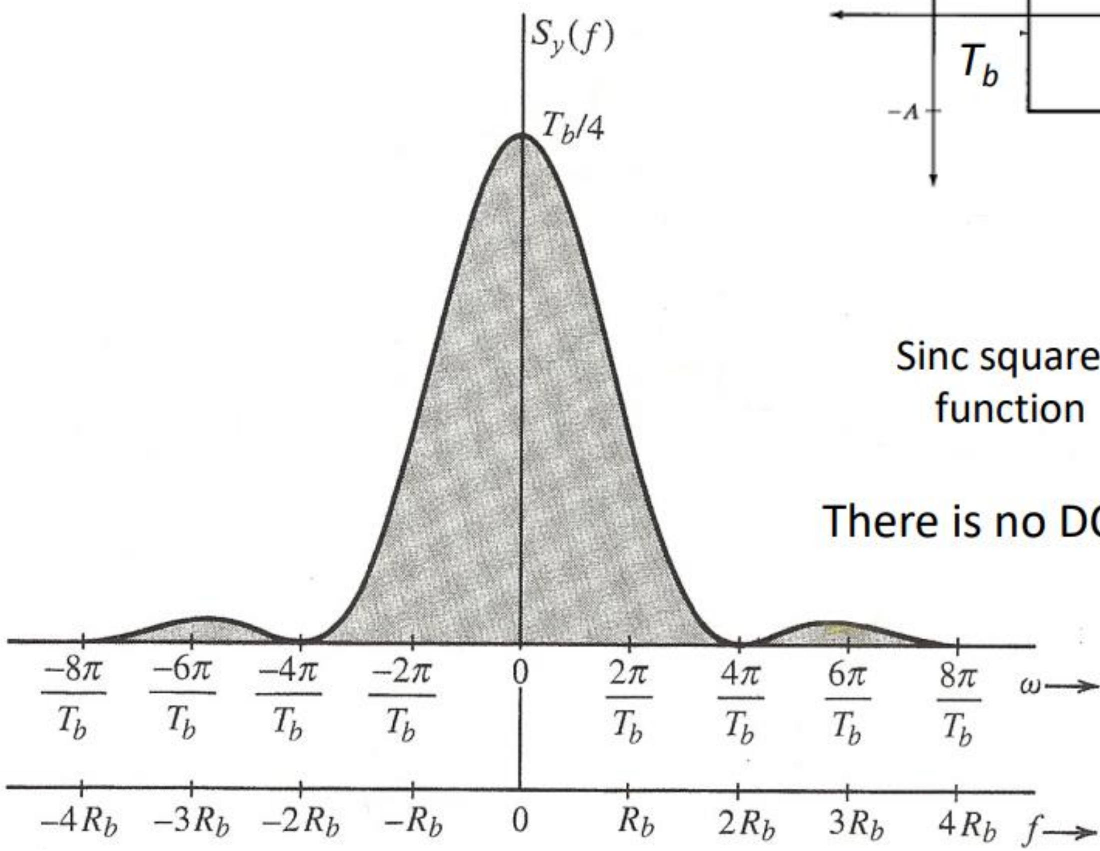
State "1" $\Rightarrow$ Pulse of amplitude +A
State "0" $\Rightarrow$ Pulse of amplitude -A

ADVANTAGES	DISADVANTAGES
1. Simplicity 2. No DC component	1. Can contain low-frequency components (leads to signal drooping) 2. No clocking component to synchronize to at receiver 3. No error correction capability

Power Spectral Density Example: Binary Polar NRZ Signal



$R_b = 1/T_b$

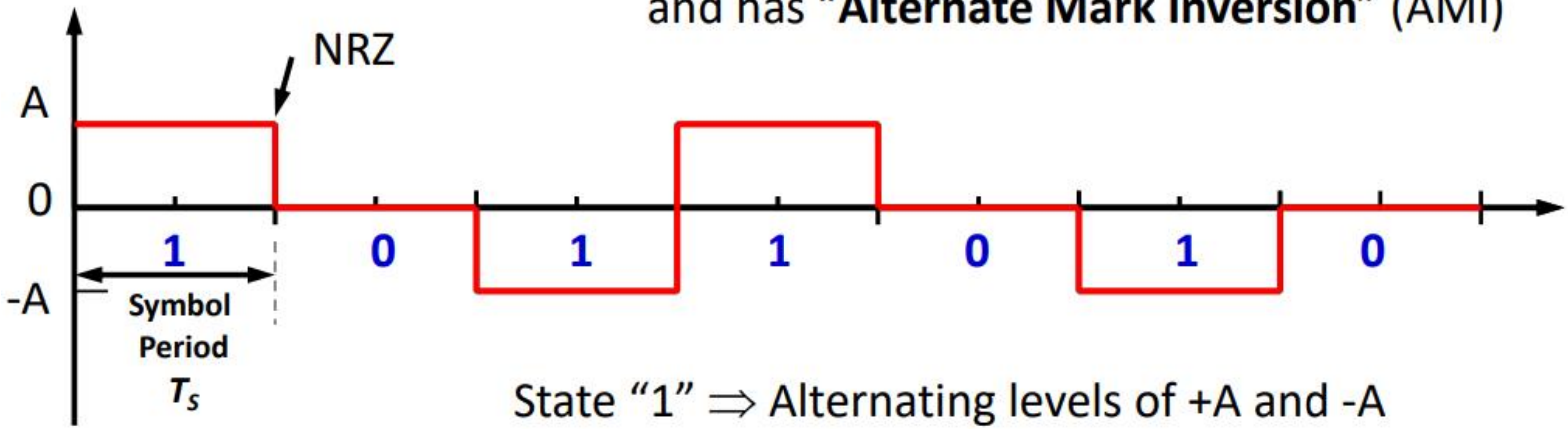


Sinc squared function

There is no DC component.

Bipolar NRZ

Uses three levels of signal level (+A, 0, -A) and has "**Alternate Mark Inversion**" (AMI)



State "1" ⇒ Alternating levels of +A and -A

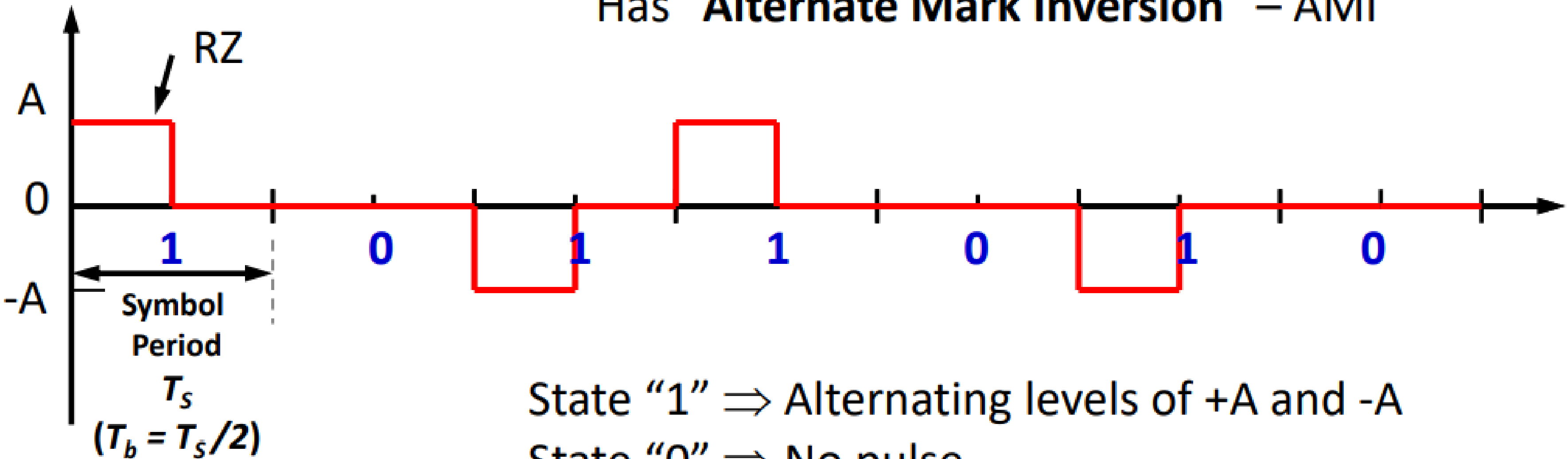
State "0" ⇒ No pulse

ADVANTAGES	DISADVANTAGES
1. No DC component 2. Less bandwidth than for unipolar & polar NRZ 3. No signal droop problem	1. No clocking component to synchronize to at receiver 2. Error correction has limited capability (detect only single error)

Bipolar RZ

Uses three levels of signal level (+A, 0, -A)

Has "**Alternate Mark Inversion**" – AMI



State "1" ⇒ Alternating levels of +A and -A

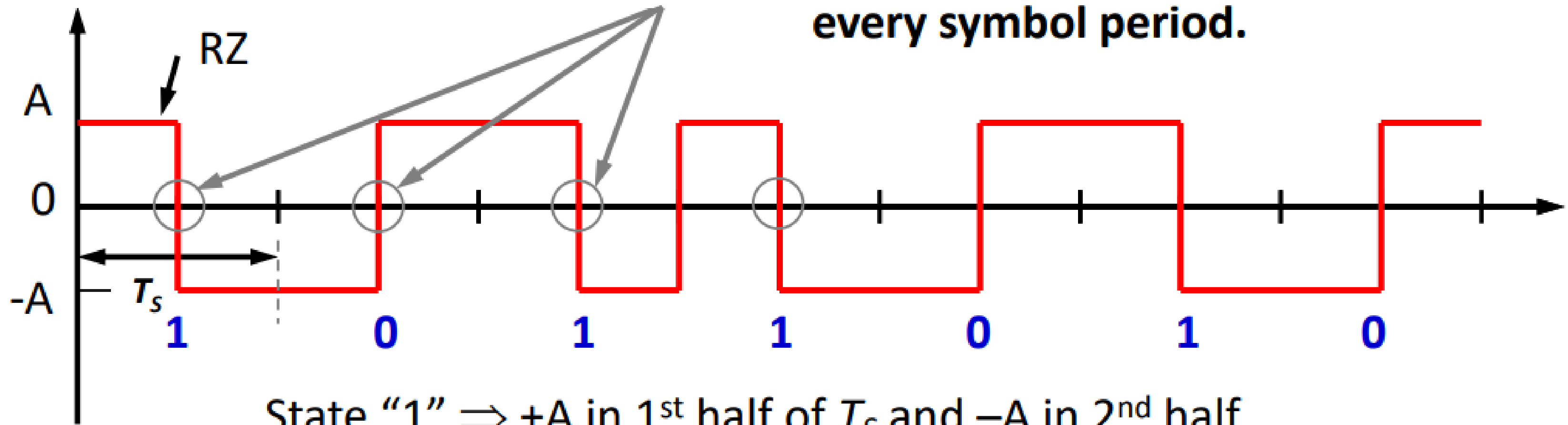
State "0" ⇒ No pulse

ADVANTAGES	DISADVANTAGES
1. No DC component 2. No signal droop problem	1. No clocking component to synchronize to at receiver 2. Limited error correction capability

Manchester (Bi-Phase or Split-Phase) Encoding

The duration of a symbol is divided into two halves.

There is a transition at the center of every symbol period.

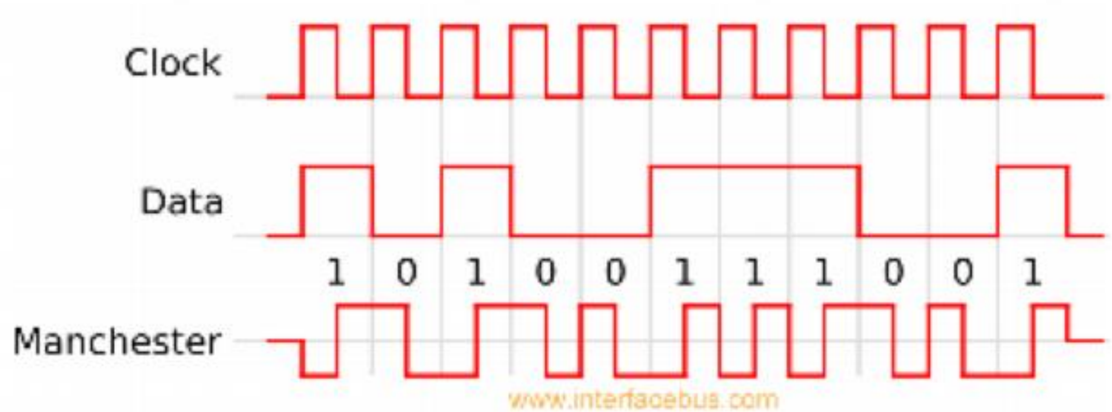


State "1" $\Rightarrow$ +A in 1st half of T_S and –A in 2nd half

State "0" $\Rightarrow$ -A in 1st half of T_S and +A in 2nd half

ADVANTAGES	DISADVANTAGES
1. No DC component 2. No signal droop problem 3. Easy to synchronize to the waveform	1. Greater bandwidth required for this waveform 2. No error correction capability

Manchester (Bi-Phase or Split-Phase) Encoding



- Manchester encoding is a form of binary phase-shift keying (BPSK).
- It is designed to encode both the clock and the data in a bit stream.
- This is also referred to "self-synchronizing data steam."
- Manchester encoding has the disadvantage of requiring higher frequencies.
- It is used in **Ethernet** (**IEEE 802.3** standard) with lower data rates.
- Also used for consumer IR protocols and in some RFID systems.

What Are the Primary Considerations When Comparing Line Codes?

1. We want the transmission bandwidths to be as small as possible.

1. Power efficiency – Keep power as low as possible.

2. Error detection and correction capability – error correcting codes are a special topic

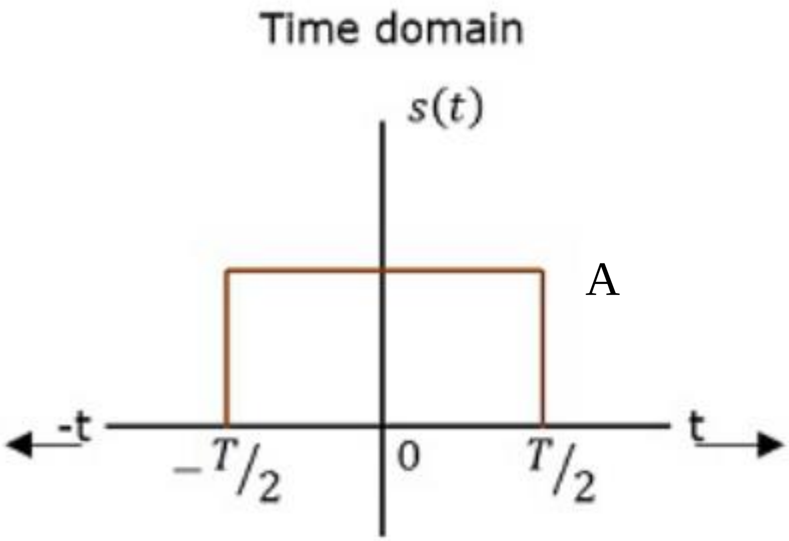
4. Favorable Power Spectral Density – We want zero power at DC ($f = 0$) to avoid baseline drift.

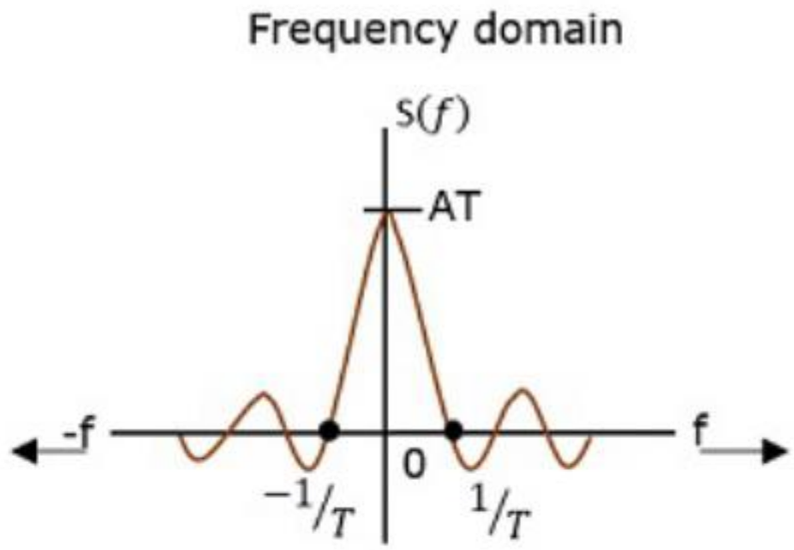
5. Adequate timing content – Often we must extract the timing or clock information from the signal.

6. Transparency – This means for every possible sequence of data the coded signal is received faithfully.

7. Signal is easily regenerated by repeaters.

Recall: rectangular <> sinc

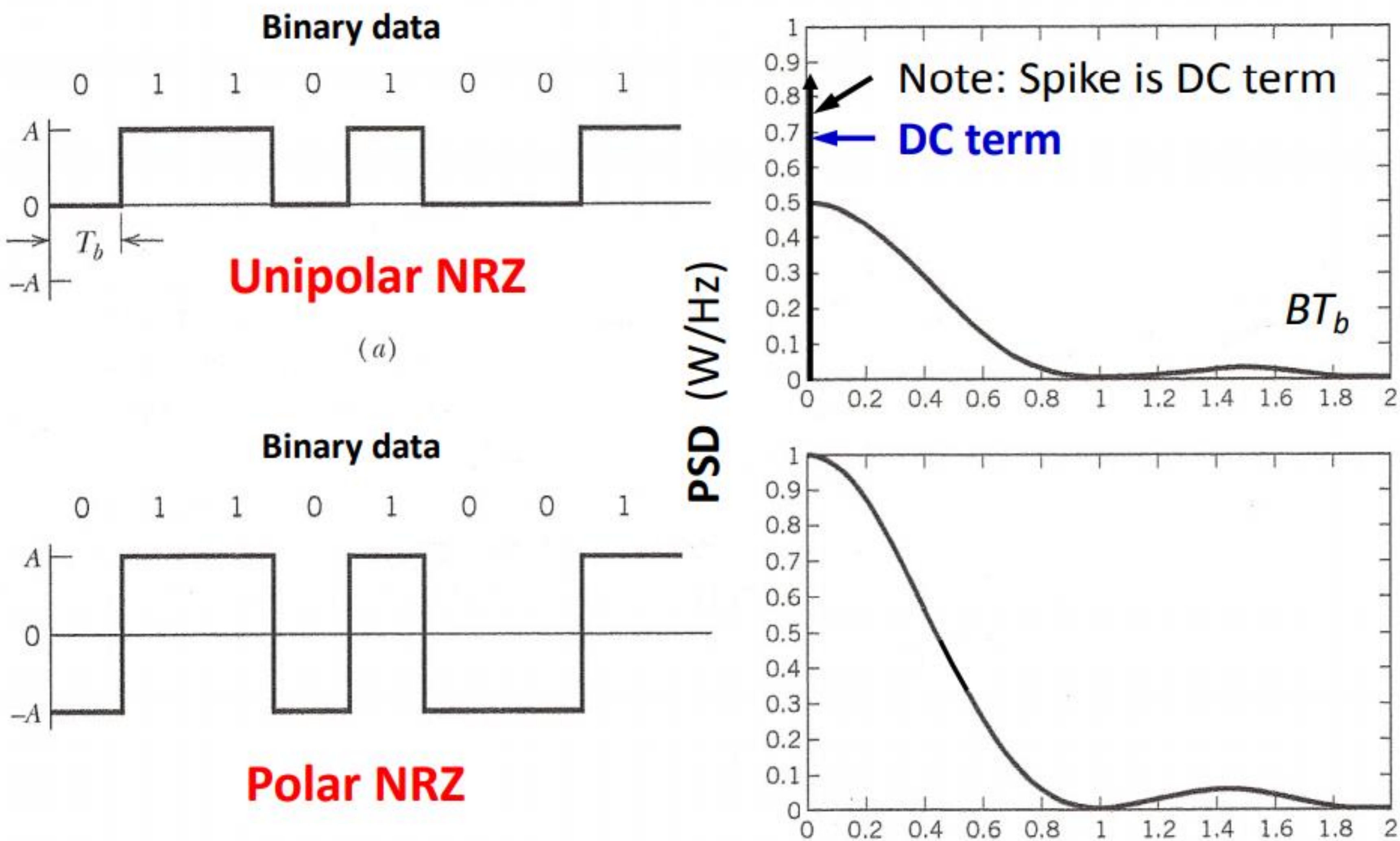




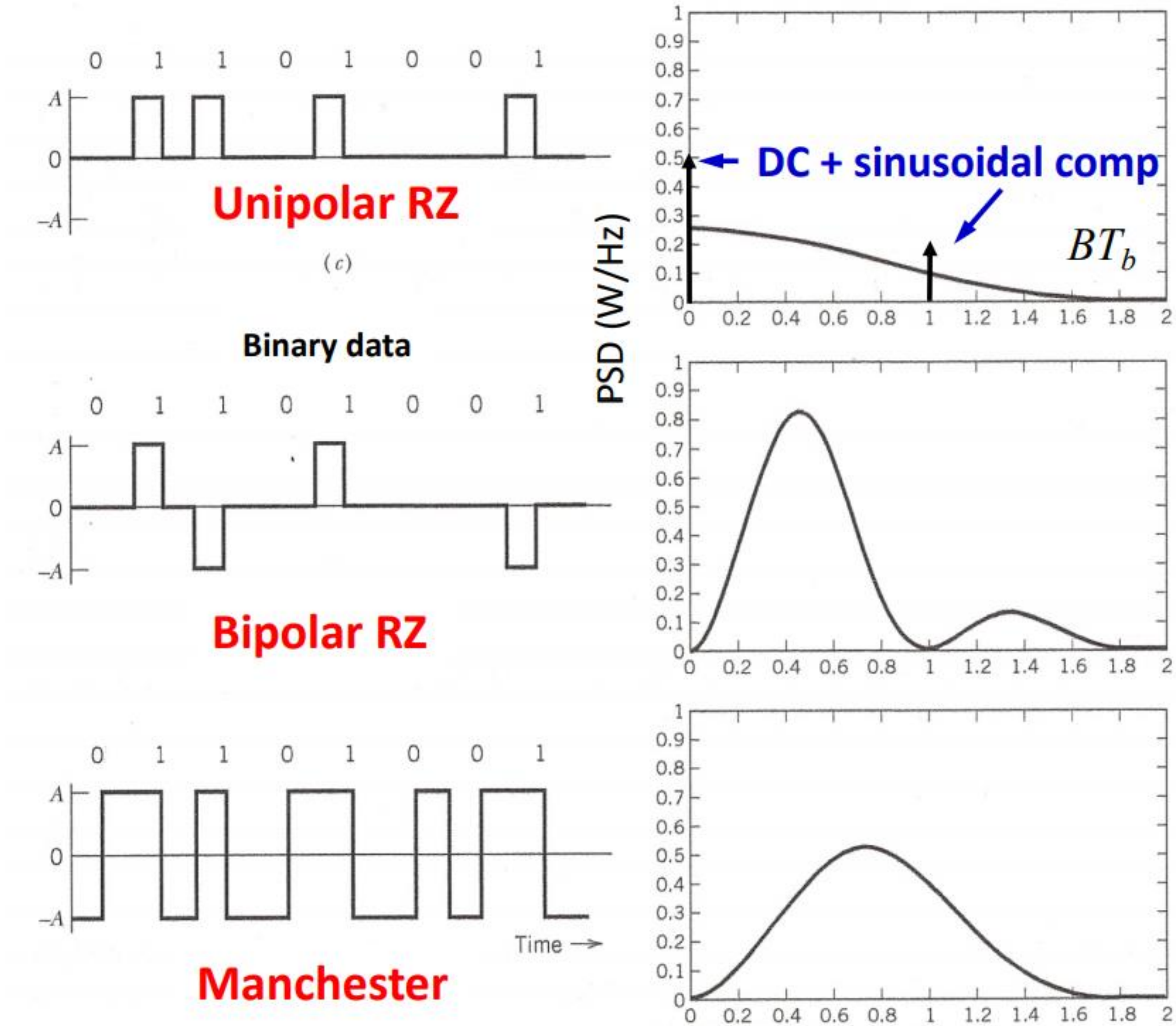
$$s(t) = A \quad |t| < T/2$$
$$= 0 \quad |t| > T/2$$

$$s(f) = AT\frac{\sin(\pi T f)}{\pi T f}$$

Power Spectral Density of NRZ and Polar NRZ Waveforms



Power Spectral Density of Unipolar RZ, Bipolar RZ & Manchester Waveforms



Qualitative Comparison of Line Coding Schemes

Line Code	Simple	BW	DC component	Error Correct ?	Clock Sync
Unipolar NRZ	Yes	Low	Yes	No	No
Unipolar RZ	Yes	$2\times$ BW	Yes	No	Yes
Polar NRZ	Yes	Moderate	No	No	No
Polar RZ	Yes	$2\times$ BW	No	No	No
Bipolar NRZ (AMI)	Moderate	Smaller BW	No	Single error capability	No
Bipolar RZ	Moderate	Smaller BW	No	Single error capability	No
Manchester	No	Greater BW	No	No	Yes

Comparing Line Coding Schemes

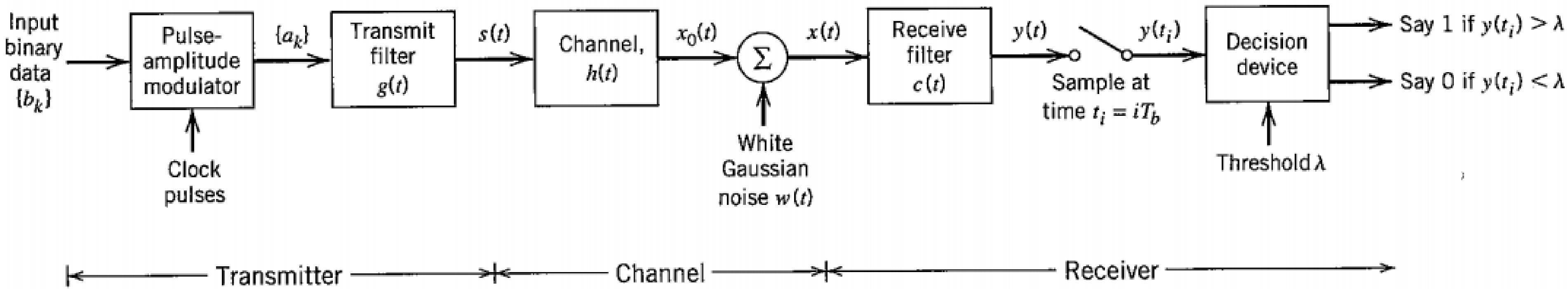
1. Unipolar NRZ and unipolar RZ only need a single-sided power supply to implement them. But polar NRZ, polar RZ, AMI and Manchester require dual power supplies.

2. AMI receivers must detect three levels. All others need only detect two levels. (AMI = Alternate Mark Inversion)

3. Polar NRZ, polar RZ and Manchester deliver a pulse every symbol period. This is not true of unipolar NRZ, unipolar RZ, and AMI – a long sequence of **0**s can be mistaken as a transmission failure.

4. AMI has a built-in error detecting capability because any **0** is interpreted as a **1**, or a **1** is interpreted as a **0**; it violates the rules for AMI coding scheme.

Binary PAM

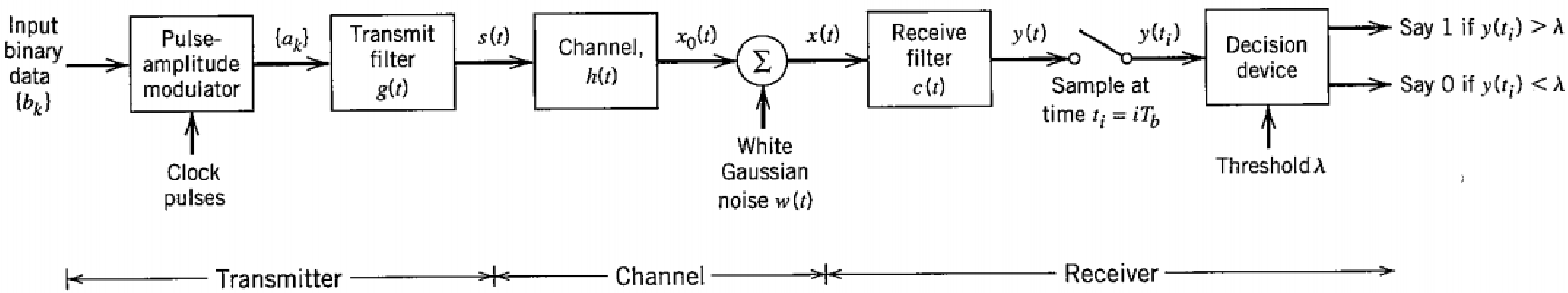


- Consider a generic *binary* PAM, where the amplitude of the pulses is varied in *discrete* manner according to the digital input data $\{b_k\}$
- The incoming binary sequence $\{b_k\}$ consists of 1's and 0's arriving every T_b seconds; these are mapped onto a symbols $\{a_k\}$; let us assume the mapping of

$$a_k = \begin{cases} +1 & \text{if } b_k = 1 \\ -1 & \text{if } b_k = 0 \end{cases}$$

- These symbols 'amplitude modulate' a *transmit filter* $g(t) \rightleftharpoons G(f)$
- Note that these two steps are analogous to line encoding (which assumed rectangular pulses), but with a pulse shape $g(t)$

Binary PAM



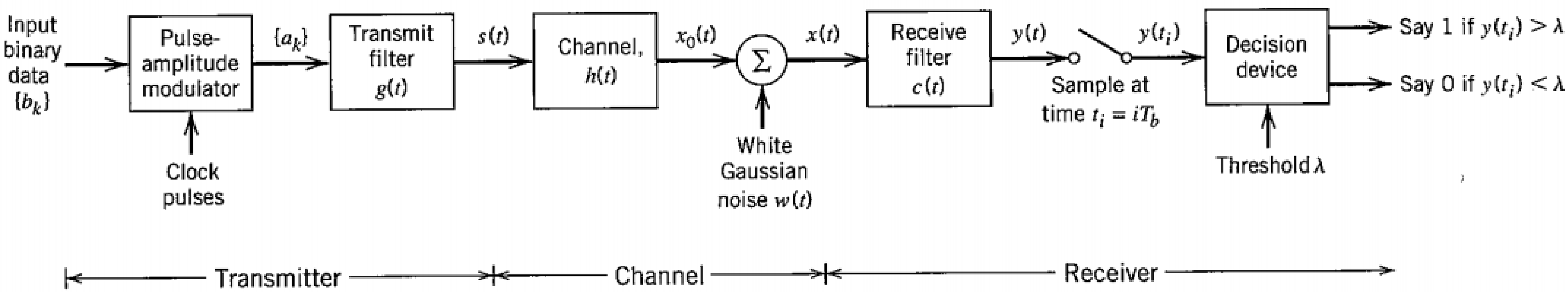
- The output of the transmit filter is a sequence of pulses

$$s(t) = \sum_{k=-\infty}^{\infty} a_k g(t - kT_b)$$

- The signal $s(t)$ is then transmitted over a *linear* communication channel, described by $h(t) \rightleftharpoons H(f)$
- Ignoring noise (for now), the channel output is the convolution

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

Binary PAM

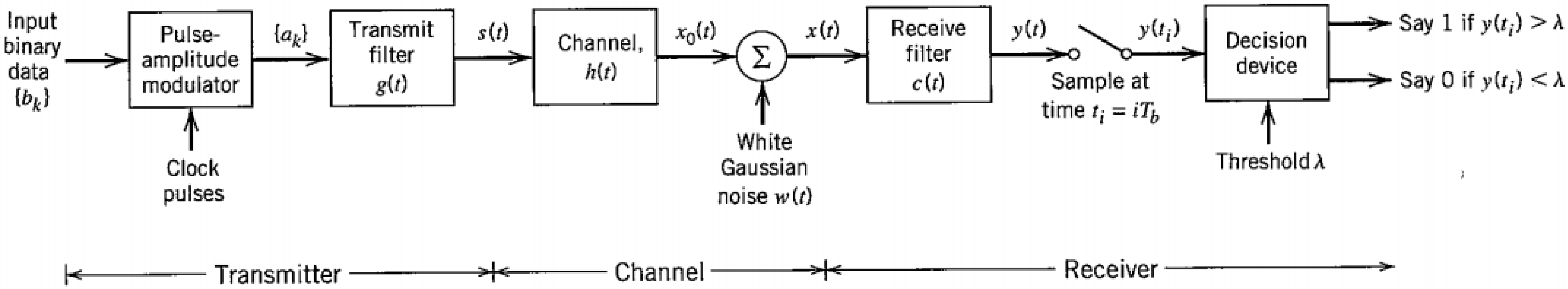


- At the receiver, channel output $x(t)$ is passed through a *receive filter*, given by $c(t) \rightleftharpoons C(f)$, resulting in

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

- The filter output $y(t)$ is sampled synchronously - the timing instants being determined by a *timing synchronisation unit*
- The samples $y(t_k)$ are interpreted by means of *decision device* - it determines the best estimate $\hat{a}_k$ of the data symbol a_k, which is them mapped back to a bit value (1 or 0).
- Typically, this is done by comparing the sample value to a threshold: if the threshold is exceeded, the symbol is assumed to be a 1, otherwise it is interpreted as a 0

Binary PAM



- NOTE 1: The effects of the channel are typically compensated for at the receiver by an *equaliser* (a transversal filter, just prior to the Decision Device), such that the equalised channel response is $\delta(t)$ (no distortion);
- We will discuss equalisation later on in the course, but for now we will assume that the channel has been equalised;
- NOTE 2: In his book, Lathi does not consider receive filtering in Chapter 7; Instead the discussion is simply in terms of a an overall pulse shape, which he denotes $p(t)$.
- Assuming an equalised channel, the overall pulse in the above system description is $p(t) = g(t) * c(t)$
- **In what follows, we will talk about he design of the overall pulse shape $p(t)$, assuming the channel has been equalised.**

Intersymbol Interference

- Assuming the sampling is done synchronously, the samples of the receive filter output at time $t = iT_b$ is

$$y(iT_b) = \sum_{k=-\infty}^{\infty} a_k p((i-k)T_b), \quad i = 0, \pm 1, \pm 2, \ldots$$

- We can simplify notation by letting

$$y_i = y(iT_b), \quad p_i = p(iT_b)$$

  and thus the input to the decision device is a *discrete convolution*

$$y_i = \sum_{k=-\infty}^{\infty} a_k p_{i-k}$$

- For now, assume the amplitude at the peak of the pulse is $p(0) = 1$

Intersymbol Interference

- When we isolate the term $i = k$, we have

$$y_i = a_i + \underbrace{\sum_{k=-\infty, k\neq i}^{\infty} a_k p_{i-k},}_{ISI} \quad i = 0, \pm 1, \pm 2, \ldots$$

- a_i is the transmitted symbol (one we are interested in), while the second term is a residual phenomenon referred to as *intersymbol interference* (ISI)
- In the absence of ISI, we have

$$y_i = a_i \quad \text{for all } i$$

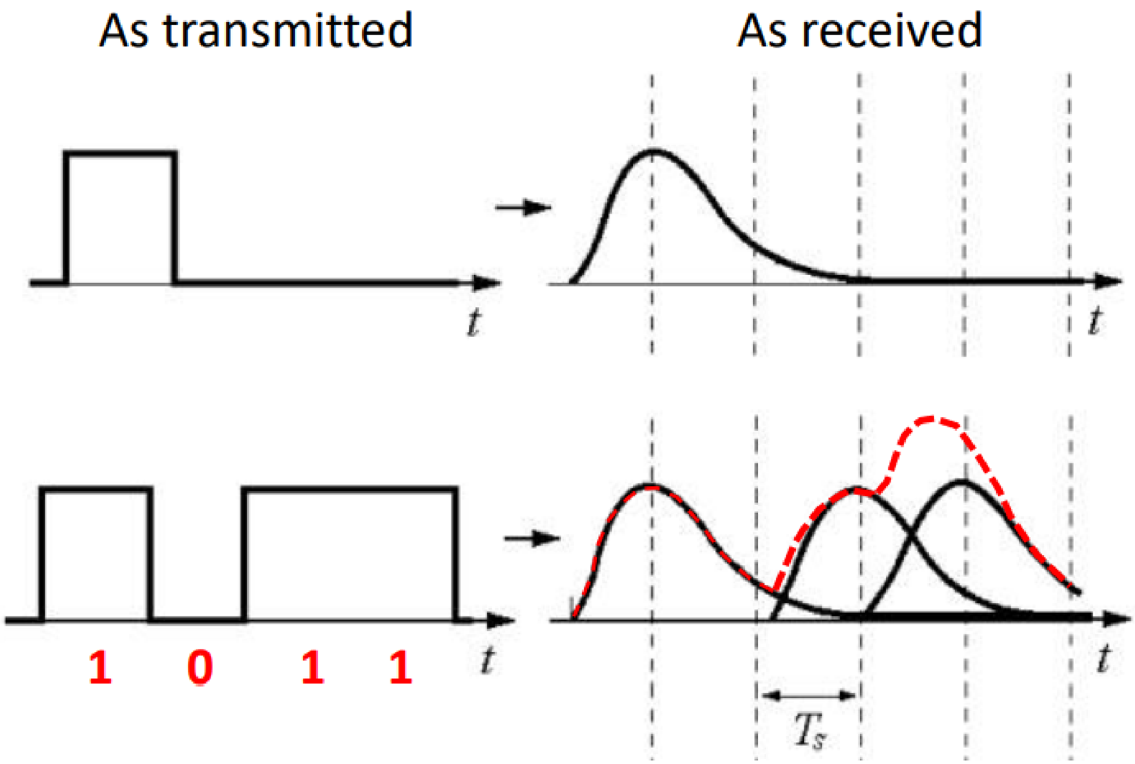
- Thus the pulse-shaping problem is:

  Determine the transmit-pulse $p(t) \leftrightharpoons P(f)$ so that

  1. **Intersymbol interference is reduced to zero**
  2. **Transmission bandwidth is conserved**

ISI

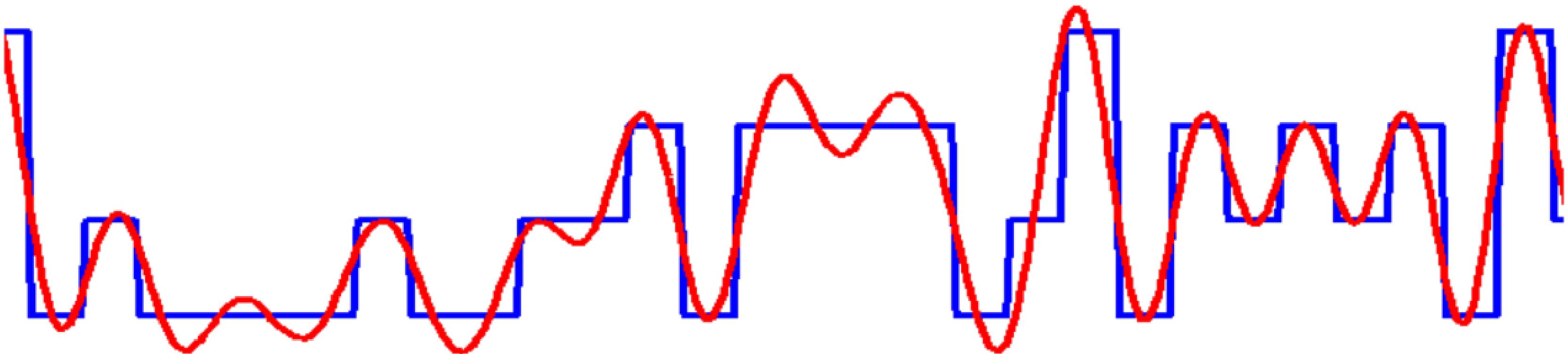
ISI is unwanted interference from adjacent (usually previous) symbols

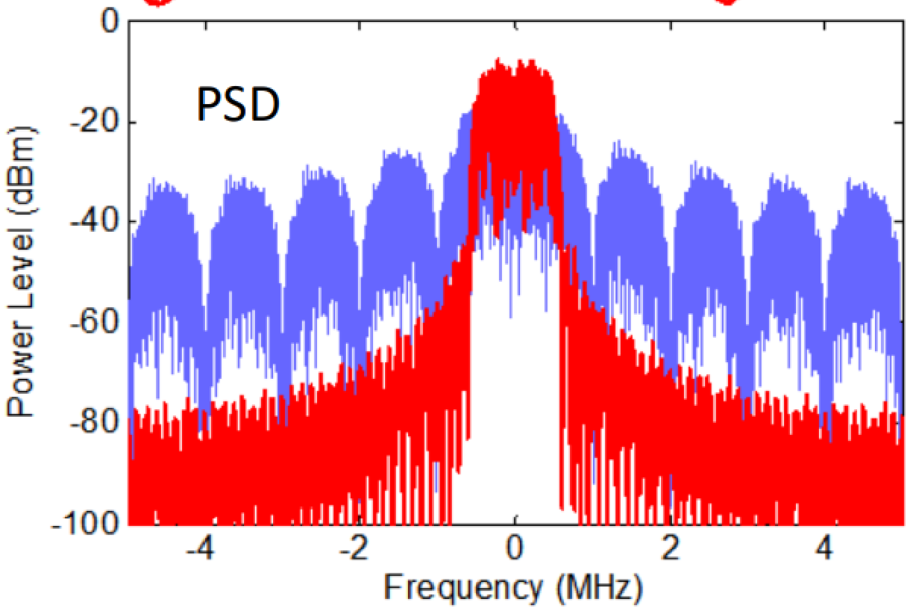


ISI is caused dispersion and other non-ideal transmission in channels

Benefit of pulse shaping in picture

Pulse Shaping

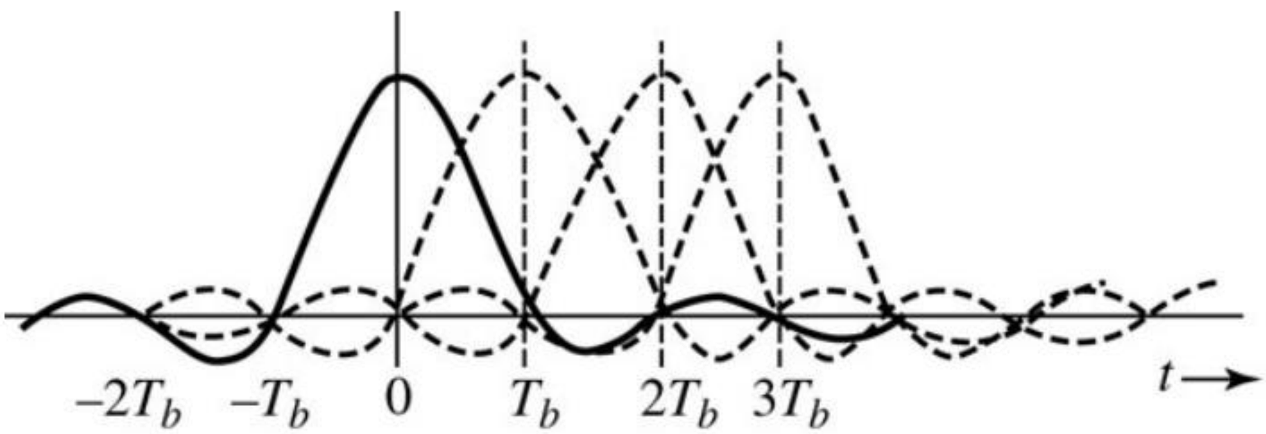




Nyquist Channel

- The optimum solution to the pulse shaping problem is one that results in zero ISI and also minimum bandwidth.
- In order to satisfy zero-ISI, we have that the overall pulse shape $p(t)$ must satisfy

$$p(t) = \begin{cases} 1 & \text{for } t = 0 \\ 0 & \text{for all } t = \pm nT_b \quad \left(T_b = \frac{1}{R_b}\right) \end{cases} \tag{1}$$



Nyquist Channel

- Recall that for a channel with bandwidth B_0, the highest achievable transmission rate is $R_b = \frac{1}{T_b} = 2B_0$, or $B_0 = \frac{1}{2T_b}$
- From sampling theory, we have that the pulse shape can be reconstructed from its samples by

$$p(t) = \sum_{n=-\infty}^{\infty} p(nT_b)\,\mathrm{sinc}(\pi R_b(t - nT_b)) \tag{2}$$

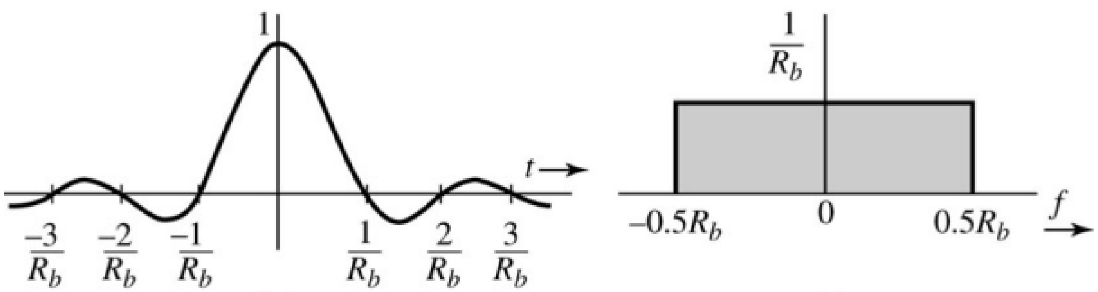
Substituting (1) into (2) gives the optimum pulse

$$p_{\text{opt}}(t) = \mathrm{sinc}(\pi R_b t) = \mathrm{sinc}(2\pi B_0 t) = \frac{\sin(2\pi B_0 t)}{2\pi B_0 t}$$

- Thus the optimum pulse shape is a sinc function
- the pulse spectrum is a *brick-wall* function

$$P_{\text{opt}}(f) = \begin{cases} \frac{1}{2B_0} & \text{for } -B_0(= -\frac{R_b}{2}) < f < B_0(= \frac{R_b}{2}) \\ 0 & \text{otherwise} \end{cases}$$

Nyquist Channel



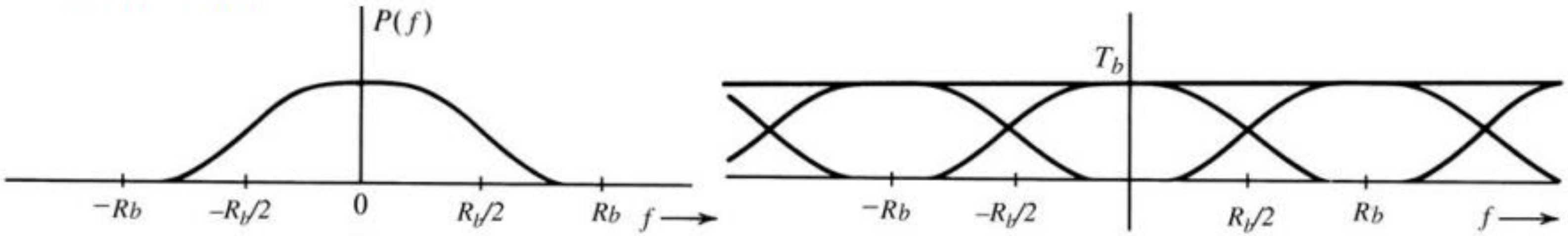
- Key observations:
  1. $P_{\text{opt}}(f)$ defines B_0 as the *minimum transmission bandwidth for zero ISI*;
  2. B_0 is referred to as the *Nyquist bandwidth*
  3. The overall system $P_{\text{opt}}(f)$ is referred to as the *Nyquist channel*
  4. The pulse $p_{\text{opt}}(t)$ has a peak value at the origin, and zeros at integer multiples of $T_b = 1/2B_0$ - thus adjacent pulses do not interfere at the sampling instants

Nyquist Criterion for zero ISI

- Condition (1) can also be stated in the frequency domain as

$$\sum_{m=-\infty}^{\infty} P(f - mR_b) = T_b \tag{3}$$

ie, infinite summation of pulse spectrum replicas spaced by $1/T_b$ is constant.



Theorem (Nyquist Zero ISI Criterion)

Given a pulse shape $p_o(t)$ for transmitting data over an imperfect channel using discrete pulse-amplitude modulation at the data rate $1/T_b$, the pulse shape $p_o(t)$ eliminates ISI if and only if its spectrum satisfies (3)

- note: the simplest way of satisfying (3) is using a rectangular pulse, as is the case with $P_{\text{opt}}(f)$

Nyquist Criterion for zero ISI

- The frequency domain formulation of the Nyquist Criterion for zero ISI can be easily derived from the time domain using basic results of sampling theory
- Let the pulse be $p(t) \leftrightharpoons P(f)$, with its bandwidth in the range $(R_b/2, R_b)$.
- Sampling the pulse every T_b is equivalent to multiplying it by an impulse train $\delta_{T_b}(t)$, giving
$$p_s(t) = p(t)\delta_{T_b}(t) = \delta(t) \tag{4}$$
where the second equality follows from the (1)
- But from Nyquist sampling theory we have that the spectrum of $p_s(t)$ is the spectrum $P(f)$ repeated periodically at intervals $1/T_b$, that is
$$P_s(f) = \frac{1}{T_b} \sum_{m=-\infty}^{\infty} P(f - mR_b) \tag{5}$$
- Taking the FT (4) ($\mathcal{F}(\delta(t) = 1)$ and substituting (5) results in (3)

Nyquist Channel

- There are two practical problems with the optimum pulse shape
  1. the requirement that the spectrum $P_o(f)$ is flat on $-B_0 < f < B_0$, and zero elsewhere; this is physically unrealizable
  2. in the time domain, $p(t)$ decreases as $1/|t|$ for large $|t|$; this is a slow rate of decay, which results in strict requirements on the sampling times
- Regarding 2, one can show that for a timing error Δt

$$y(\Delta t) = \sqrt{E} a_0 \operatorname{sinc}(2B_0 \Delta t) + \sqrt{E} \left(\frac{\sin(2\pi B_0 \Delta t)}{\pi} \right) \sum_{k=-\infty, k \neq 0}^{\infty} \frac{(-1)^k a_k}{2\pi B_0 \Delta t - k}$$

- the second term is the ISI due to the *timing error* Δt

Nyquist Channel

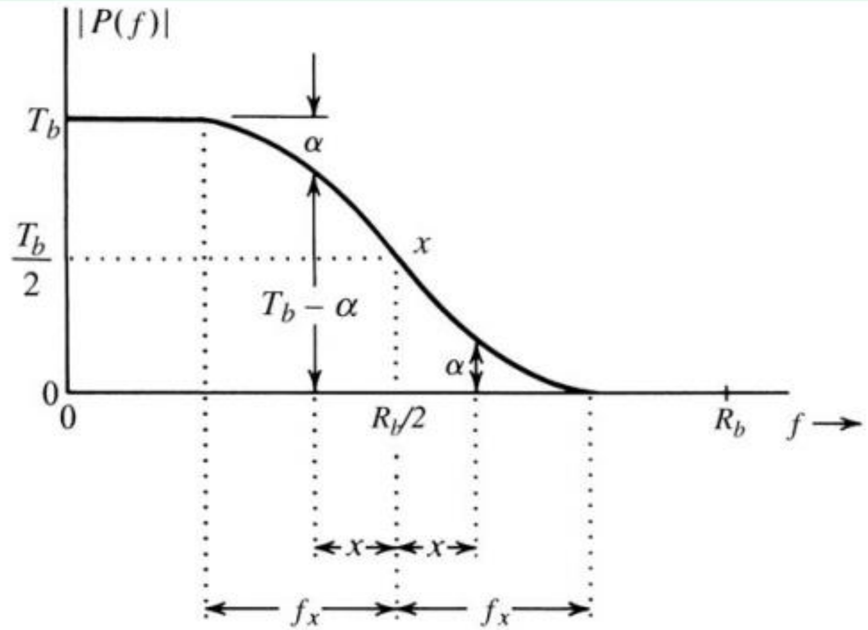
- Nyquist has shown that there exists a pulse which satisfies the zero-ISI criterion and also decays faster than $1/t$. He has shown that such a pulse requires bandwidth of $kR_b/2$ with $1 < k < 2$.
- Consider the spectrum of overlapping pulses in the range $0 < f < R_b$
- The summation over this range reduces to $P(f) + P(f - R_b) = T_b$
- Using the fact that the pulses are real valued, and applying the conjugate symmetry property of FT, one can show

$$P\left(\frac{R_b}{2}\right) + P^*\left(\frac{R_b}{2} - x\right) = T_b \quad |x| < 0.5R_b$$

- Furthermore, we can chose $P(f)$ to be real and positive, and thus

$$\left|P_o\left(\frac{R_b}{2} + x\right)\right| + \left|P_o\left(\frac{R_b}{2} - x\right)\right| = T_b \quad |x| < 0.5R_b$$

Nyquist Channel



- The bandwidth is $f_x + 0.5R_b$, ie f_x greater than the minimum for rate R_b
- We define the ratio r, referred to as the *rollof factor*

$$r = \frac{\text{excess bandwidth}}{\text{minimum bandwidth}} = 2f_x R_b = \frac{f_x}{0.5R_b}$$

- Since $f_x < R_b/2$, $0 \leq r \leq 1$ and the actual pulse bandwidth exceeds the minimum required (B_0)

$$B_T = \frac{R_b}{2} + \frac{rR_b}{2} = \frac{(1+r)R_b}{2} = (1+r)B_0$$

Raised Cosine Pulse

- As per the above, we are after a pulse consisting of
  1. *Flat portion* on some frequency interval $0 \leq f \leq \frac{R_b}{2} - f_x$ for some f_x
  2. *A Roll-off portion*, on the interval $\frac{R_b}{2} - f_x \leq f \leq \frac{R_b}{2} + f_x$
- The answer is a **raised cosine** pulse

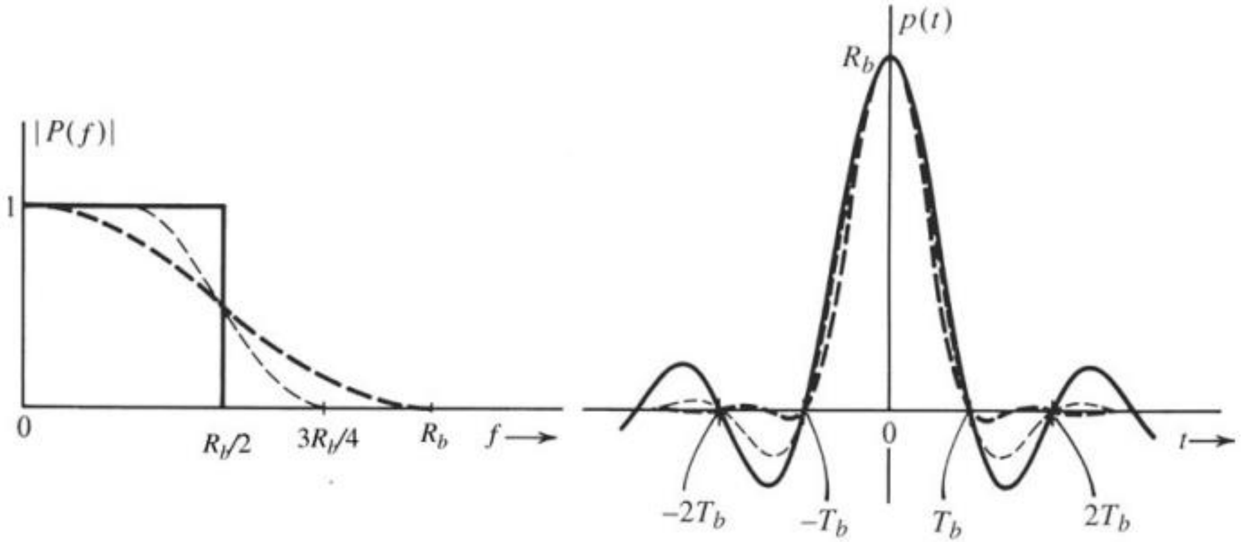
$$P(f) = \begin{cases} 1 & |f| < \frac{R_b}{2} - f_x \\ \frac{1}{2}\left\{1 - \sin\pi\left[\frac{f - R_b/2}{2f_x}\right]\right\} & \left|f - \frac{R_b}{2}\right| < f_x \\ 0 & |f| > \frac{R_b}{2} + f_x \end{cases}$$

where the parameter f_x is related to the chosen roll-off r by

$$r = \frac{2f_x}{R_b}$$

- for $r = 0$ we obtain the ideal (brick wall) Nyquist channel

Raised Cosine Pulse



- The time domain raised cosine pulse can be shown to be (for any $0 \leq r \leq 1$)

$$p(t) = R_b \operatorname{sinc}(\pi R_b t) \left(\frac{\cos \pi r R_b t}{1 - 4r^2 R_b^2 t^2} \right)$$

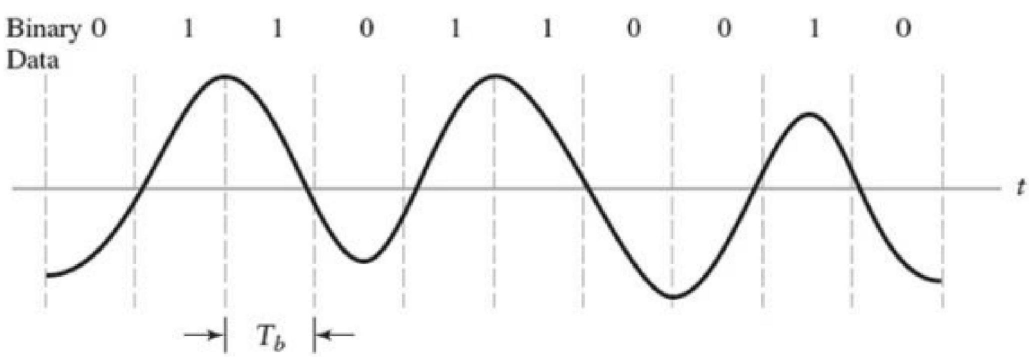
- the first factor $\operatorname{sinc}(\pi R_b t)$ defines the Nyquist channel, and ensures zero-crossings at the desired sampling time (also: decays as $1/t$)
- the second factor decreases as $1/|t^2|$ for large $|t|$; this significantly reduces the tails, making it less sensitive to sampling time errors.
- note that the amount of ISI due to Δt decreases as r increases

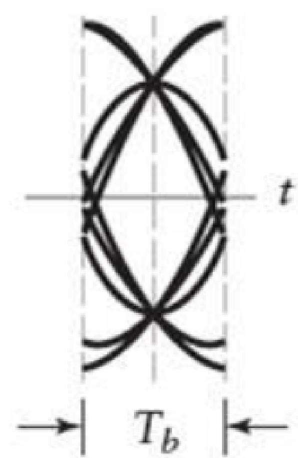
Transmission Bandwidth Requirement

- The parameter r dictates the shape of the roll-off region and the amount of bandwidth utilized.
- when $r = 0$, excess bandwidth is zero and we have the minimum possible bandwidth for transmission
- when $r = 1$ the excess bandwidth is $R_b/2 = B_0$, and thus we are transmitting using twice the minimum bandwidth
- note: case of $r = 1$ allows for easy timing synchronization

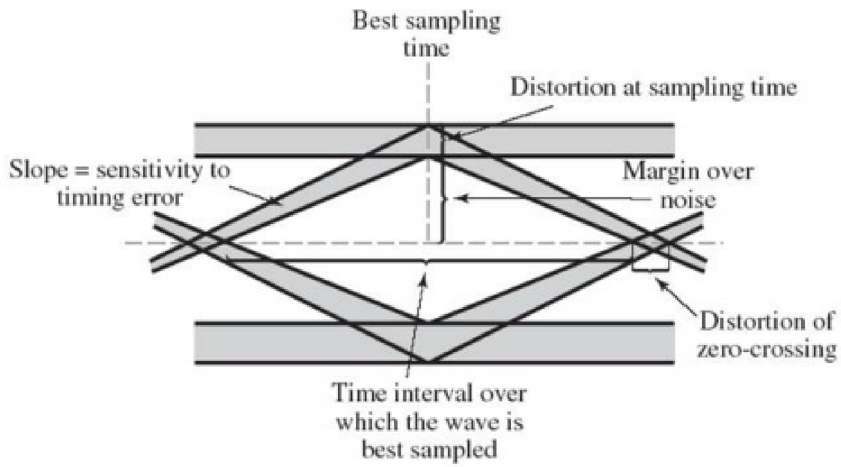
Eye Diagram

- Having discussed the mitigation strategy for ISI, we now talk about a tool for evaluation
- An *eye pattern* or *eye diagram* is obtained by the synchronized superposition of successive symbols intervals of the distorted waveform at the output of the receive filter



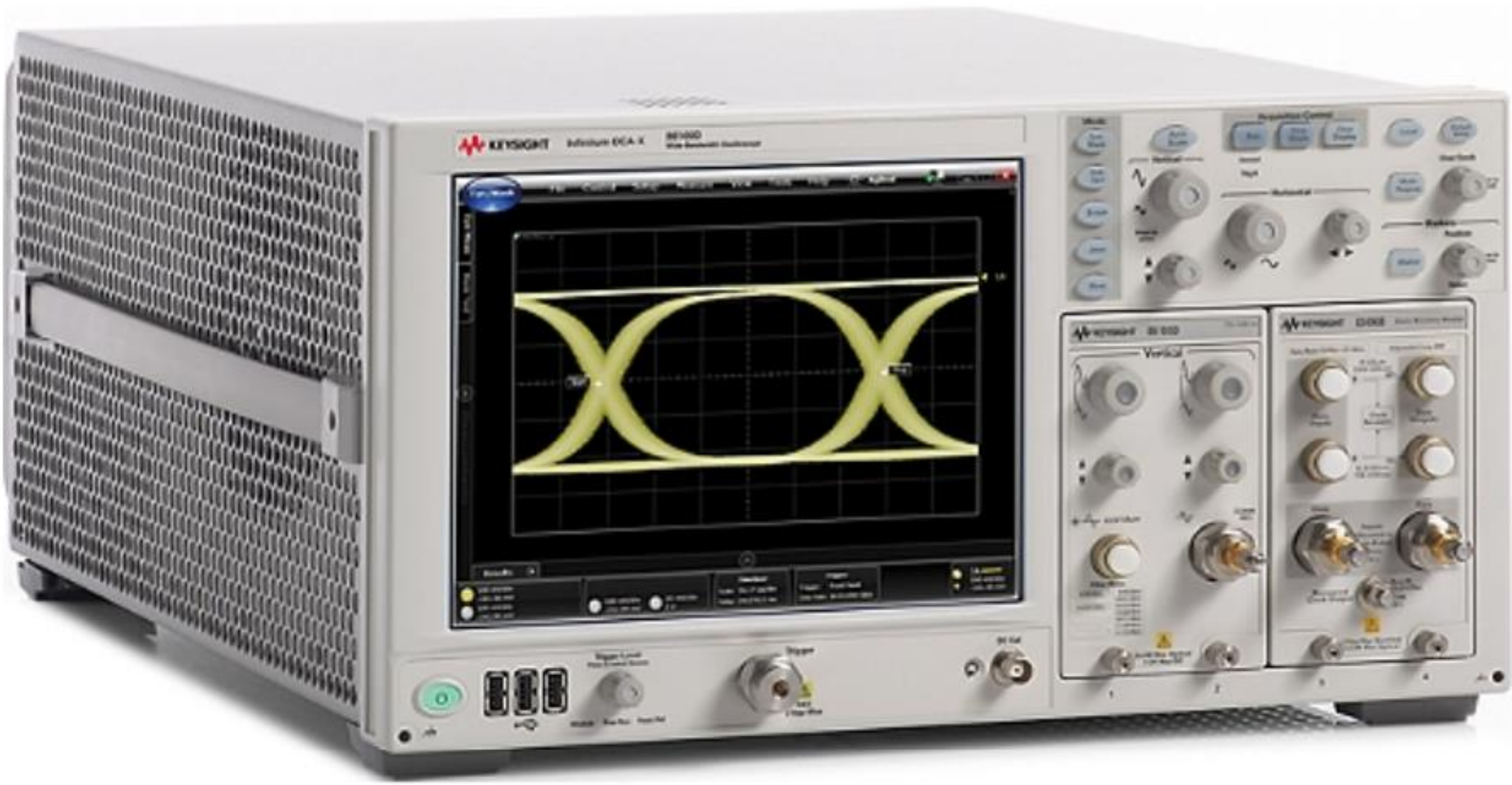


Eye Diagram



- An eye diagram can be used to evaluate the following characteristics of a binary data transmission
  1. *optimum sampling time* - the width of the eye opening determines the interval for sampling resulting in no decision errors (assuming no noise); optimum sampling time is the point where the eye opening is the widest
  2. *zero-crossing jitter* - zero crossings are used for timing synchronisation; irregularities in these crossings result in non-optimum sampling times
  3. *timing sensitivity* - the rate at which the eye patter is closed with varied sampling time

Measuring Eye Diagram



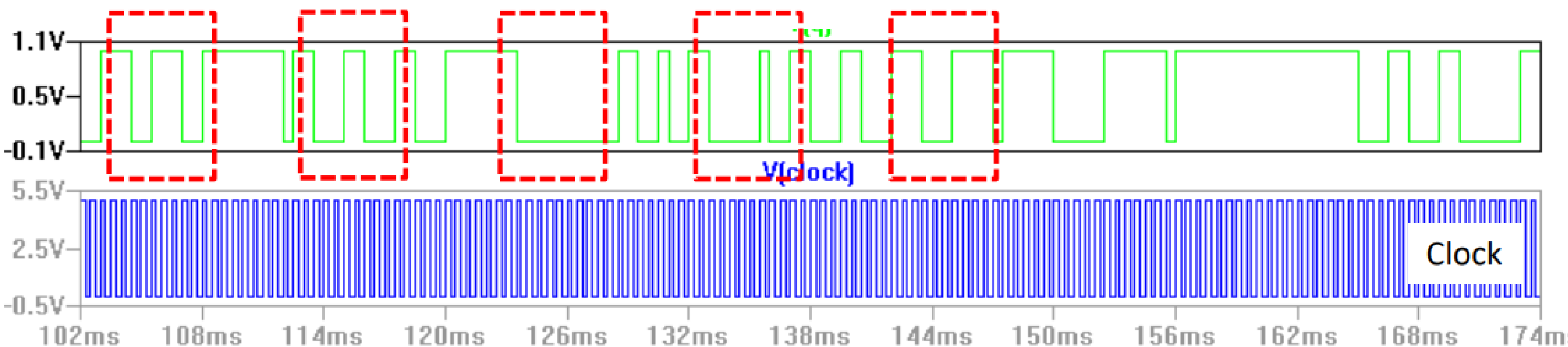
86100D Infiniium DCA-X Wide-Bandwidth Oscilloscope Mainframe

Oscilloscope with Memory for Displaying Multiple Traces

Display window of oscilloscope – trigger sets the beginning of each trace or window

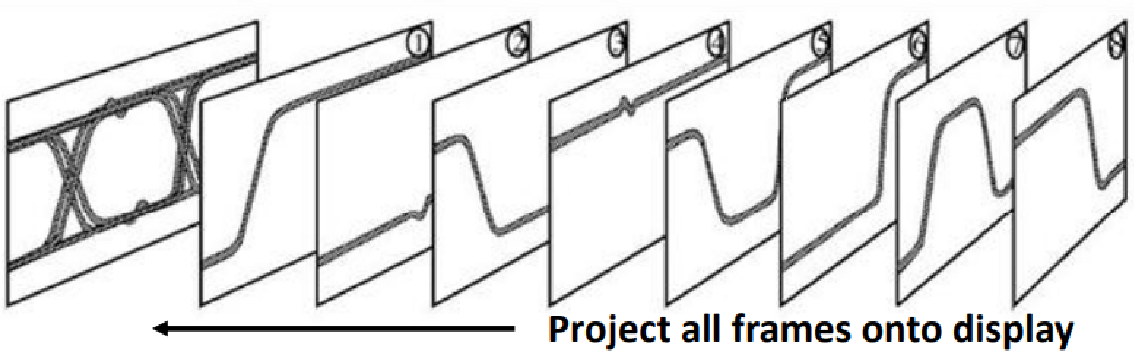
PRBS (pseudo-random bit stream) is generated for testing BER.

Data stream of pulses



(a)

Projected Eye Diagram



(b)

Preliminary: Bit Rate *vs*. Symbol Rate (or Baud Rate)

The speed of data is commonly expressed in bits/second or bytes/second. The ***data rate*** R_b is related to the ***bit period*** T_b (duration of a bit).

$$R_b = 1/T_b$$

The bit rate is commonly referred to as the ***channel capacity***.

Communication systems use ***symbols*** to convey information.
A symbol may be one bit per symbol (called ***binary***), or a group of bits, or a collection of defined voltage levels (multiple level symbols), etc.

The ***symbol rate*** R_S is related to the symbol's period (or duration) T_S by

$$R_S = 1/T_S$$

The symbol rate is also called ***Baud rate***. Bit rate R_b can be written as

$$R_b = R_S \times \log_2(\eta) = R_S \times n$$

where $\eta = 2^n$ = number of levels (for n bits per symbol).

Signal-to-Noise Ratio *vs.* Energy/Bit-to-Noise Ratio

In analog and digital communications, **signal-to-noise ratio**, usually written ***S/N*** or ***SNR***, is a measure of **signal** strength relative to background **noise** strength. The **ratio** is usually expressed in decibels (dB) and equals $10 \cdot \log_{10}[S/N]$.

Another metric that is often more useful in digital systems is the **energy per bit-to-noise power ratio**, denoted by E_b/N_0.

Define: R_b = bit rate (in bits per second)
S = total signal power (watts)
E_b = energy per bit (in joules/bit)
N = total noise power (over entire bandwidth B in Hz)
N_0 = noise spectral density ($N = N_0 \cdot B$ where B = bandwidth)

Then,

$$\frac{S}{R_b} = E_b \quad \text{and} \quad \frac{E_b}{N} = \frac{S}{R_b \cdot N} \quad \text{and} \quad SNR = \frac{R_b E_b}{N_0 B}$$

Increasing the bit rate R_b increases the *SNR*. However, in general it also increases the noise in the denominator, which lowers the *SNR*.

Relationship Between E_b/N_o and S/N

$$E_b = \left(\frac{\text{Signal power}}{\text{Bit rate}}\right) = \frac{S}{R_b} = \left(\frac{E/t}{B/t}\right)$$

$$N_O = \left(\frac{\text{Noise power}}{\text{Bandwidth}}\right) = \frac{N}{B}$$

$$\frac{E_b}{N_O} = \frac{\left(\frac{S}{R_b}\right)}{\left(\frac{N}{B}\right)} = \left(\frac{S}{R_b}\right)\times\left(\frac{B}{N}\right) = \underbrace{\left(\frac{S}{N}\right)}_{\text{Signal-to-noise ratio}}\times\underbrace{\left(\frac{B}{R_b}\right)}_{\text{Processing gain}}$$

Signal-to-Noise Ratio *vs.* Energy/Bit-to-Noise Ratio (continued)

In digital communications systems the E_b/N_0 ratio can be thought of as a "normalized signal-to-noise ratio."

We can roughly equate signal power to energy per bit by

$E_b = P_{signal} \cdot T_S$, where T_S is the symbol period,

and the noise power per hertz, denoted by N_0, is the total noise power N divided by bandwidth B.

E_b/N_0 is commonly used to as the primary variable in establishing the bit rate parameters for all modulation schemes.

M-ary PAM

- Recall the symbol mapper we considered for converting the bit values into amplitudes of pulses

$$a_k = \begin{cases} +1 & \text{if } b_k = 1 \\ -1 & \text{if } b_k = 0 \end{cases}$$

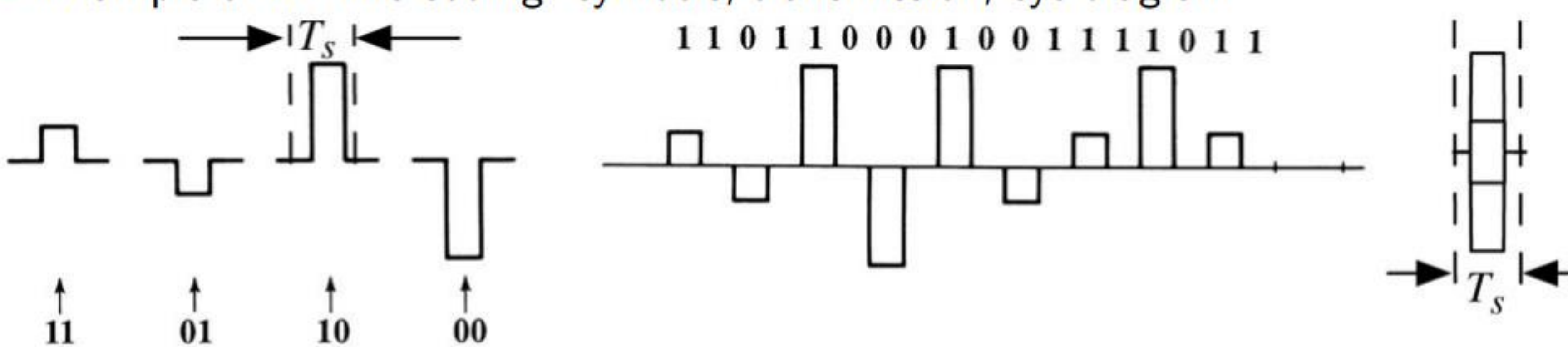
- Thus, regardless of the line coding used, each pulse carries exactly one bit of information; the data (or *bit)* rate is $R_b = 1/T_b$
- If we wanted to increase the bit rate by M we can reduce T_b by a factor of M
- This, however, would increase the bandwidth of the pulses by the corresponding amount, which is a steep price to pay
- A better approach is to let each pulse represent more than one data bits...

M-ary PAM

- Consider a symbol mapper which maps *pairs* of bit into one of $M = 4$ symbols

$$a_k = \begin{cases} +3 & \text{message bits } 10 \\ +1 & \text{message bits } 11 \\ -1 & \text{message bits } 01 \\ -3 & \text{message bits } 00 \end{cases}$$

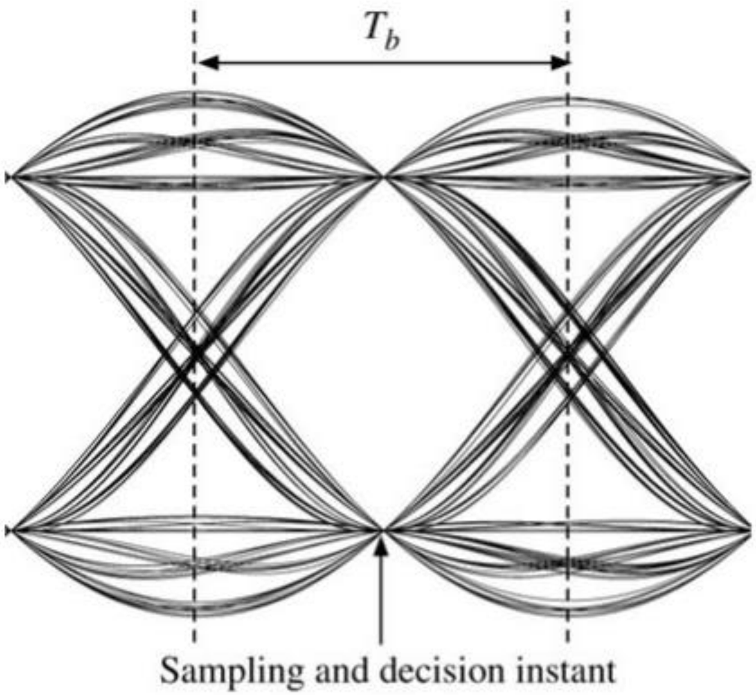
- Each symbol is transmitted by a pulse of corresponding amplitude, over a time interval $(0, T_b)$
- Example of RZ line coding: symbols, transmission, eye diagram

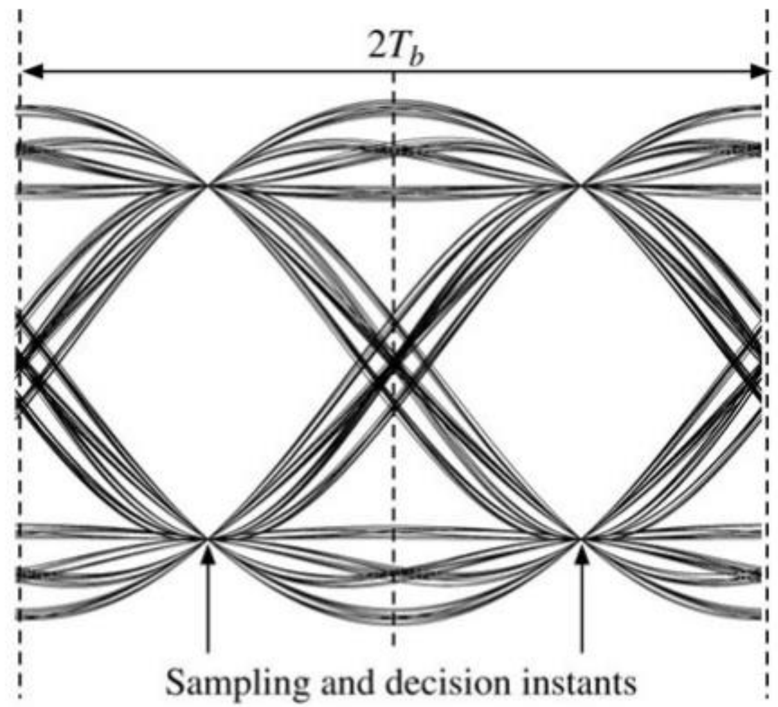


M-ary PAM

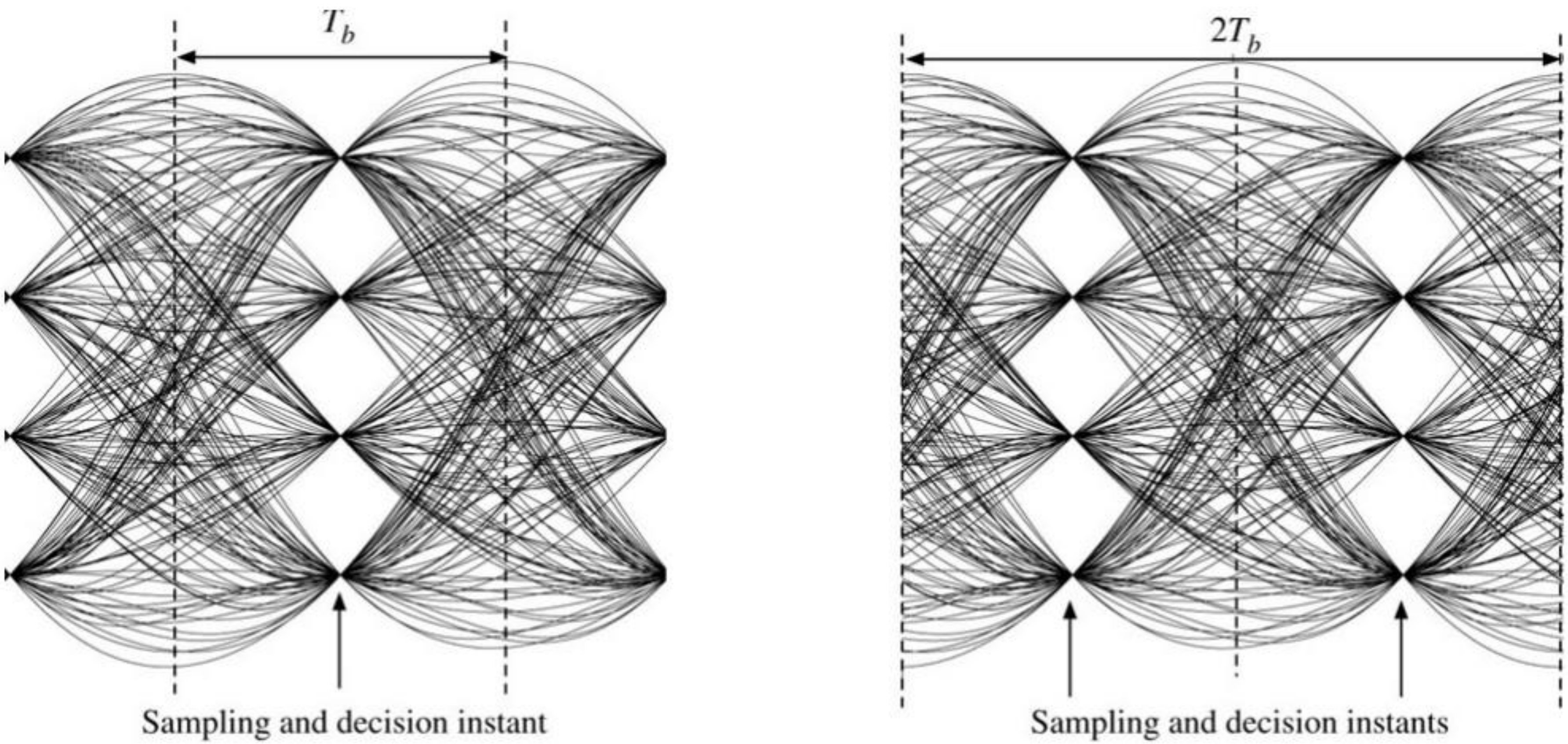
- What we have is a data (bit) rate increase by a factor of $\log_2 M = 2$, ie $R_b = \log_2 M \frac{1}{T_b}$
- Note that the symbol rate remains at $R_s = \frac{1}{T_b}$
- The symbol rate is often referred to as the *baud rate*
- The technique of transmitting one of M symbols is referred to as *M-ary transmission*
- This particular example, of transmitting M pulse amplitudes is *M-ary PAM* - the same term used for the analogue PAM modulation if sampled signal
- Other $M-$ary techniques will be discussed later, in the context of *bandpass* transmission

Binary PAM Eye Diagram





M-ary PAM Eye Diagram



M-ary PAM

- Note that the pulse shaping criterion applies to M-ary PAM modulation
- Disadvantages of M-ary transmission is the power penalty.
    - In order to maintain the error performance, the minimum separation between pulse amplitudes must be similar to that for binary signaling
    - Pulse amplitude increases with M, and thus required power with M^2 (for a data rate increase of $\log_2 M$

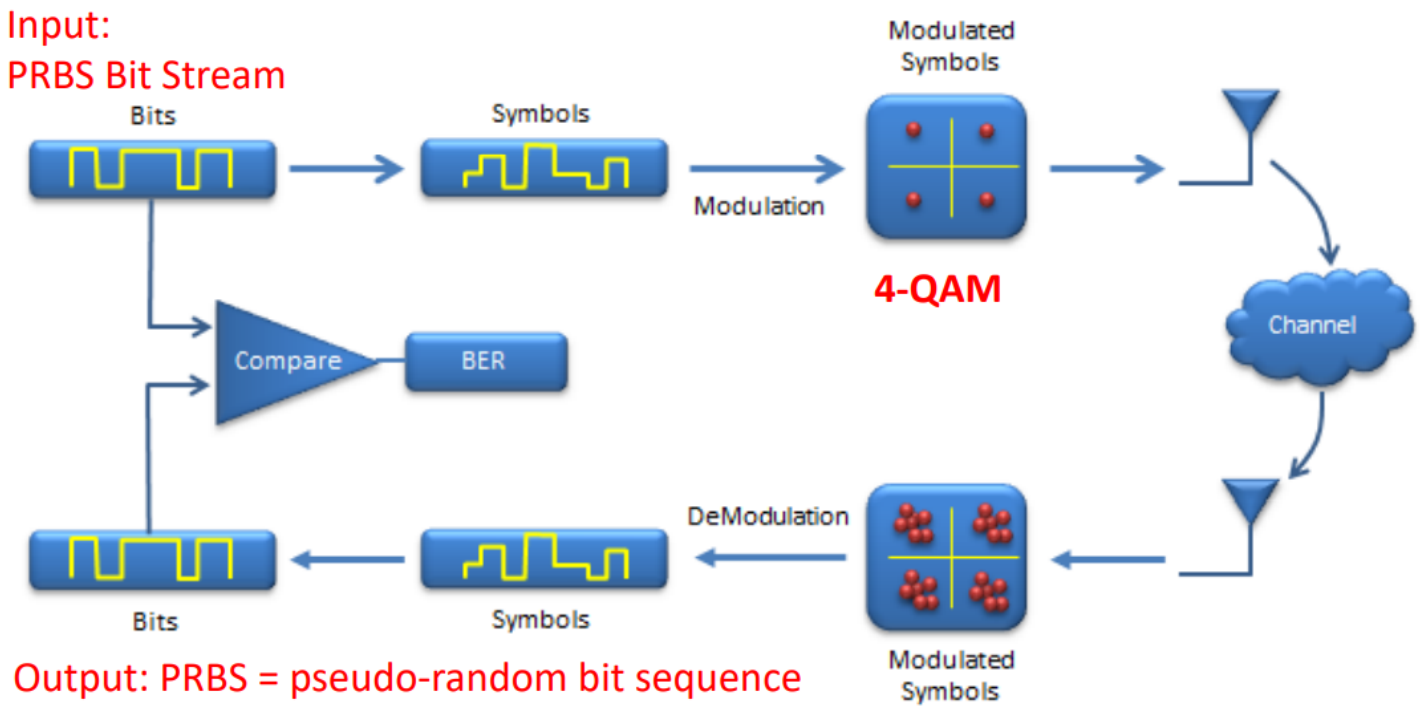
A look at BER

In digital transmission, the number of **bit errors** is the number of received bits of a data stream over a communication channel that have been altered due to noise, interference, distortion or bit synchronization errors.

The **bit error rate** (**BER**) is the number of bit errors per unit time. The **bit error ratio** (also **BER**) is the number of bit errors divided by the total number of transferred bits during a studied time interval.

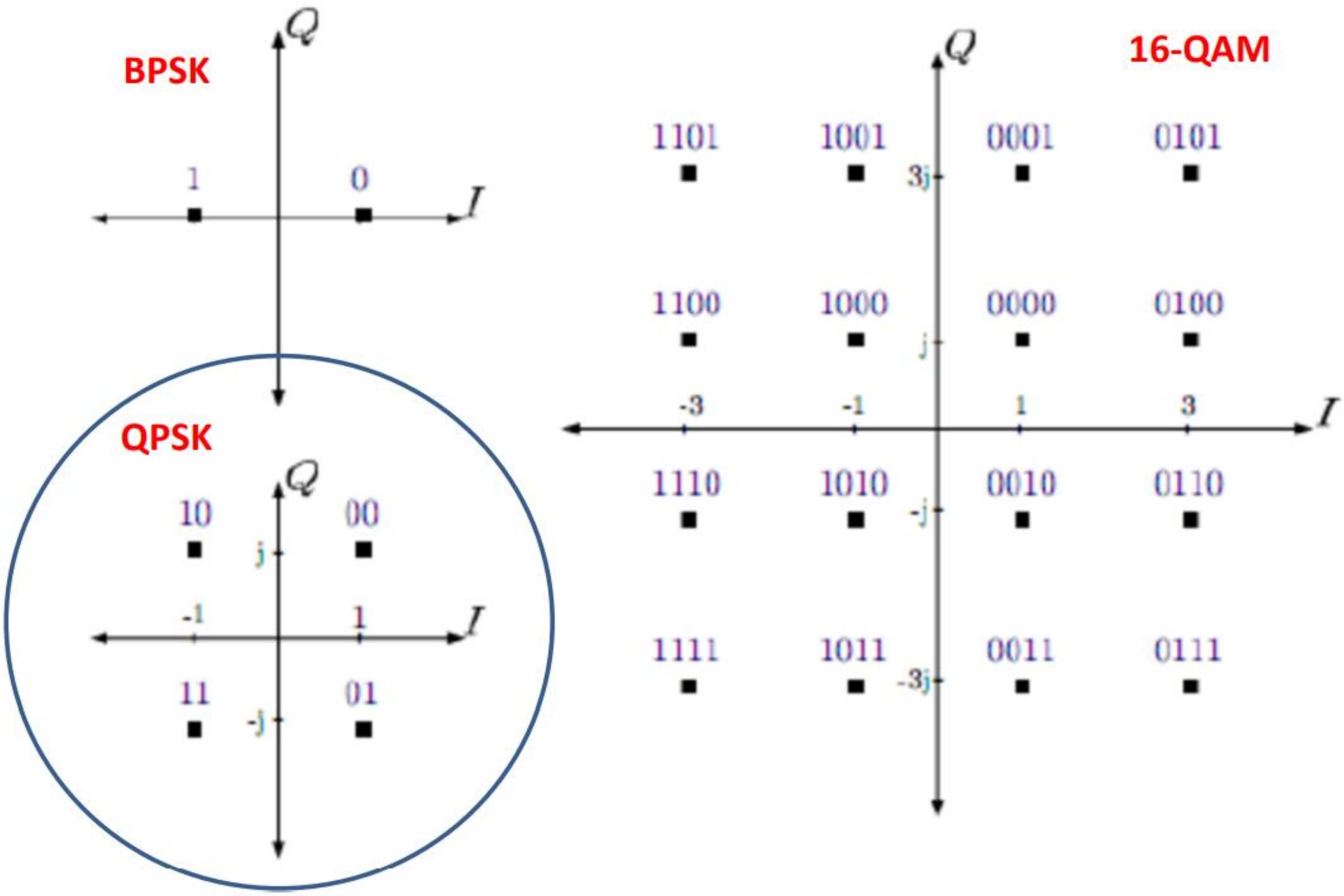
BER calculation

$$BER = \frac{\text{Number of bit errors}}{\text{Total number of bits transmitted}}$$



How many bits in the PRBS bit pattern stream?
The lowest BER_{min} to be measured should have a PRBS length equals $3 \times (1/BER_{min})$. Example: For $BER_{min} = 10^{-9}$, then length > 3×10^9 bits.

Signal constellation



Summary

- Thus far we have derived the pulse shape criteria to eliminate ISI- Nyquist Zero ISI Criterion
- That is the overall pulse shape $p(t)$ has to satisfy
  - Time domain formulation: zero crossings at iT_b, ie $p(t - iT_b) = 0$, $i \neq 0$
  - Freq domain formulation: $\sum_{m=-\infty}^{\infty} P(f - mR_b) = T_b$
- We saw that the easiest way to satisfy the above criteria is an ideal brick-wall pulse shape (sinc in the time domain)
  - Such a pulse has the minimum bandwidth for a given data rate: $R_b = 2B_0$
  - It is however impractical and has poor timing synchronization properties
- We looked at the Raised Cosine shape
  - satisfies Zero ISI condition
  - uses extra bandwidth (controlled by a rolloff parameter r) to achieve better synchronization characteristics