Integrated Circuits for Digital Communications

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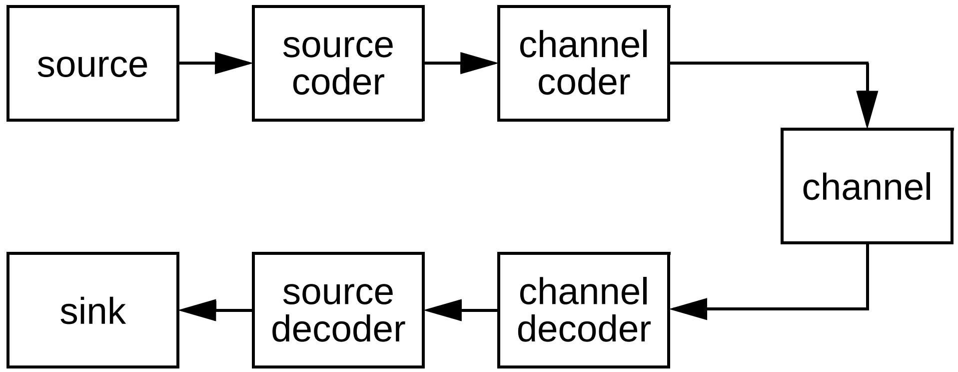
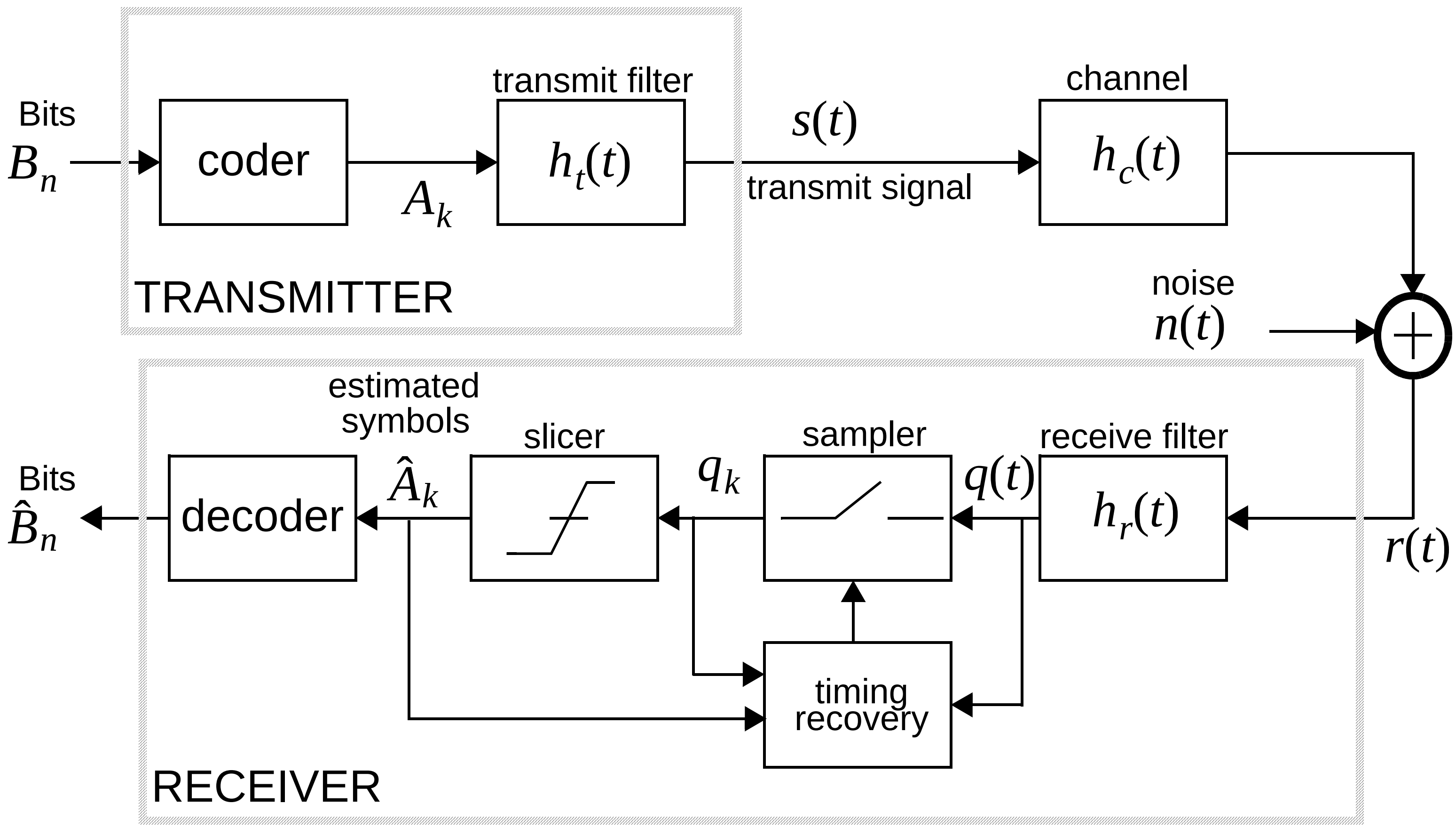
Basic Baseband PAM Concepts

General Data Communication System



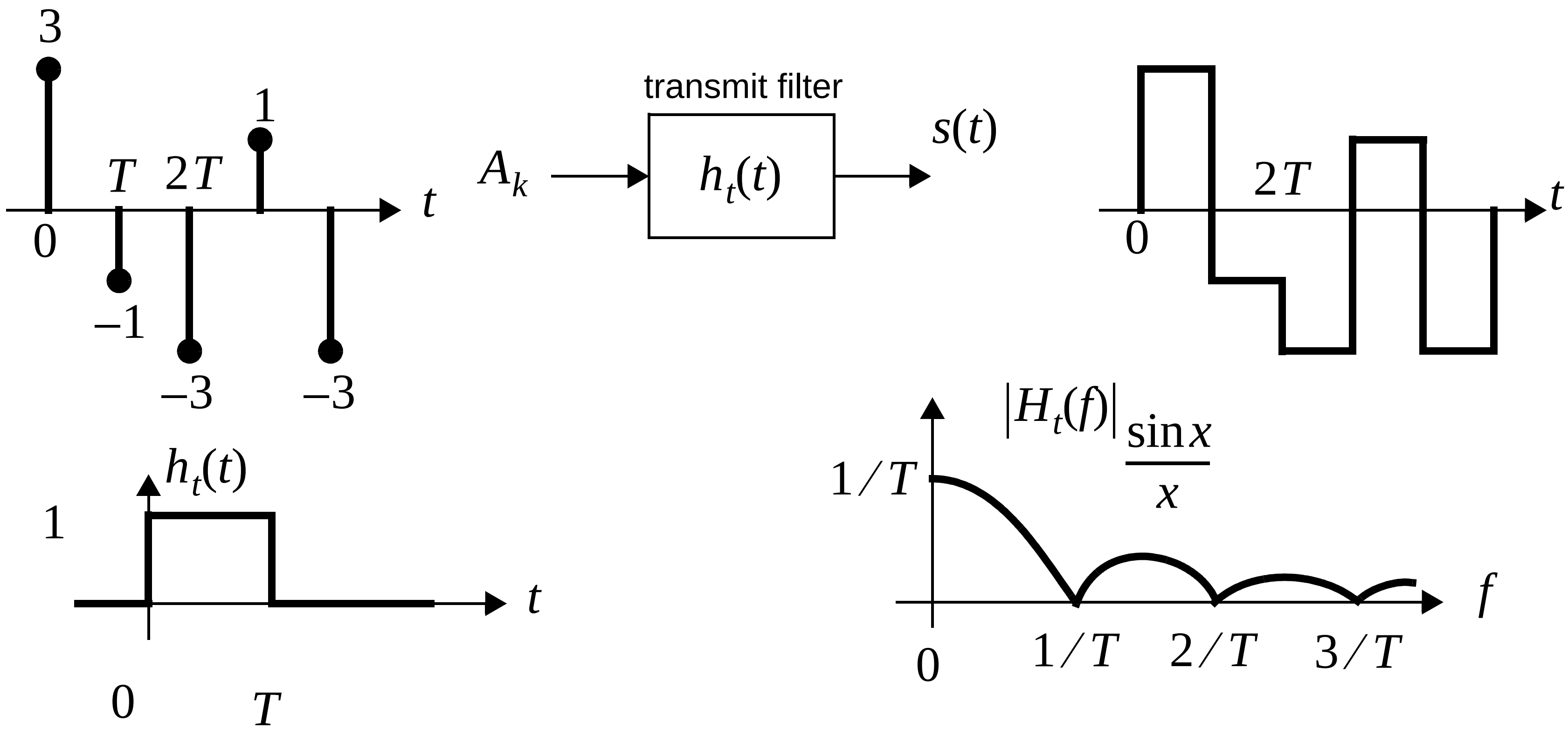
- Source coder removes redundancy from source (i.e. MPEG, ADPCM, text compression, etc.)
- Channel coder introduces redundancy to maximize information rate over channel.
(i.e. error-correcting codes, trellis coding, etc.)
- Our interest is in channel coding/decoding and channel transmission/reception.

Basic Baseband System



- In 2B1Q, coder maps pairs of bits to one of four levels — $A_k = \{-3, -1, 1, 3\}$

Rectangular Transmit Filter



- The spectrum of A_k is flat if random.
- The spectrum of $s(t)$ is same shape as $H_t(f)$

Nyquist Pulses

- $h(t)$ is the impulse response for transmit filter, channel and receive filter ($\otimes$ denotes convolution)

$$h(t) = h_t(t) \otimes h_c(t) \otimes h_r(t) \tag{1}$$

$$q(t) = \sum_{m=-\infty}^{\infty} A_m h(t - mT) + n(t) \otimes h_r(t) \tag{2}$$

- The received signal, $q(t)$, is sampled at kT.

$$q_k = \sum_{m=-\infty}^{\infty} A_m h(kT - mT) + u(kT) \quad , u(t) \equiv n(t) \otimes h_r(t) \tag{3}$$

- For zero intersymbol interference (i.e. $q_k = A_k + u_k$)

$$h(kT) = \delta_k \qquad (\delta_k = 0, 1, 0, 0, 0, \ldots) \tag{4}$$

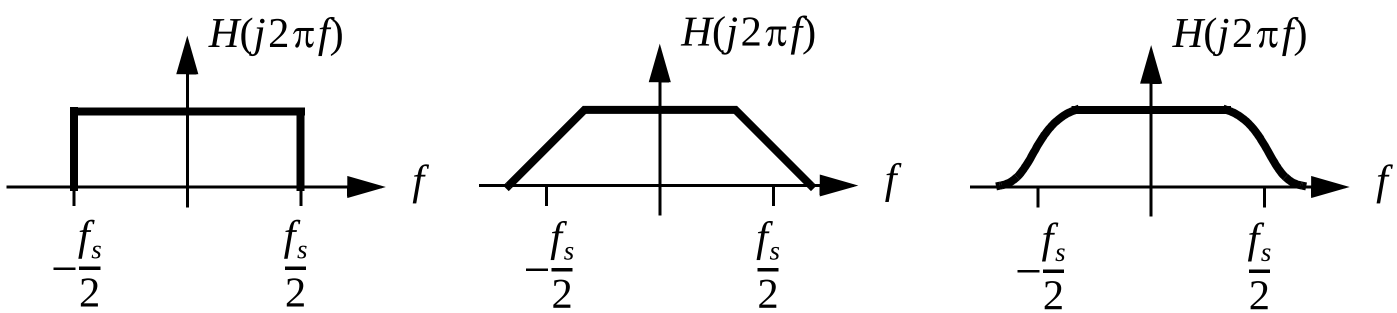
Nyquist Pulses

- For zero ISI, the same criteria in the frequency domain is: ($f_s = 1/T$)

$$\frac{1}{T}\sum_{m=-\infty}^{\infty} H(j2\pi f + jm2\pi f_s) = 1 \qquad (5)$$

- Known as Nyquist Criterion

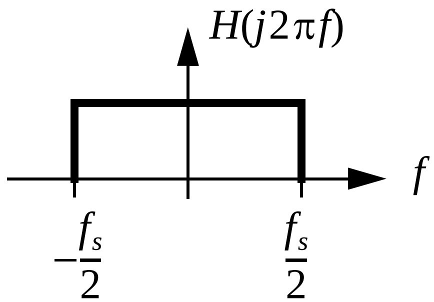
Example Nyquist Pulses (in freq domain)



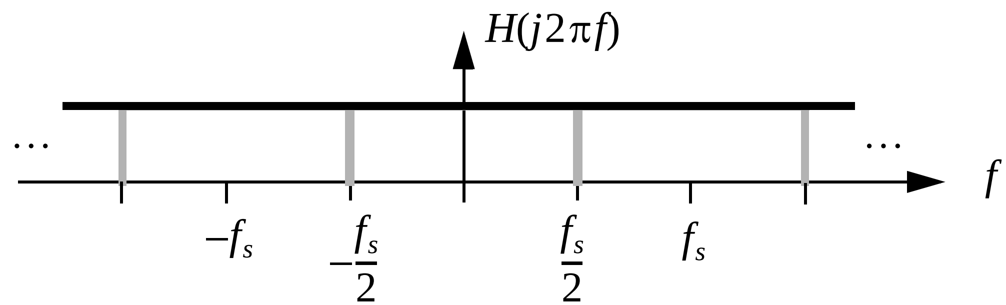
Sinc pulse

Raised-cosine pulse

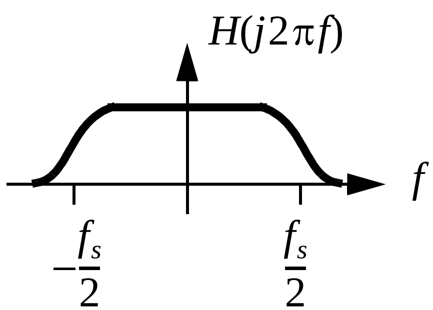
Nyquist Pulses



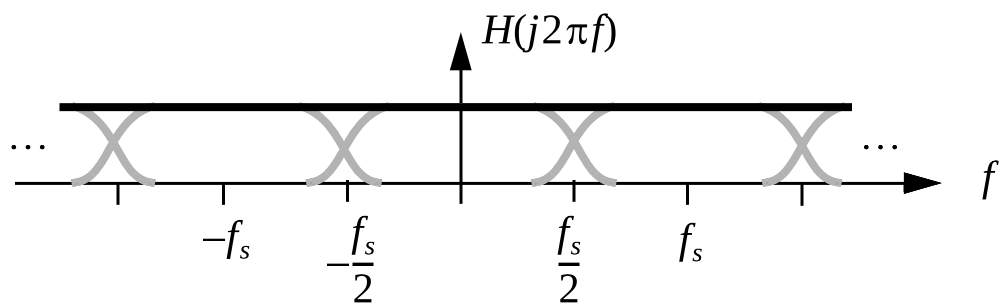
Sinc pulse



Sinc pulse

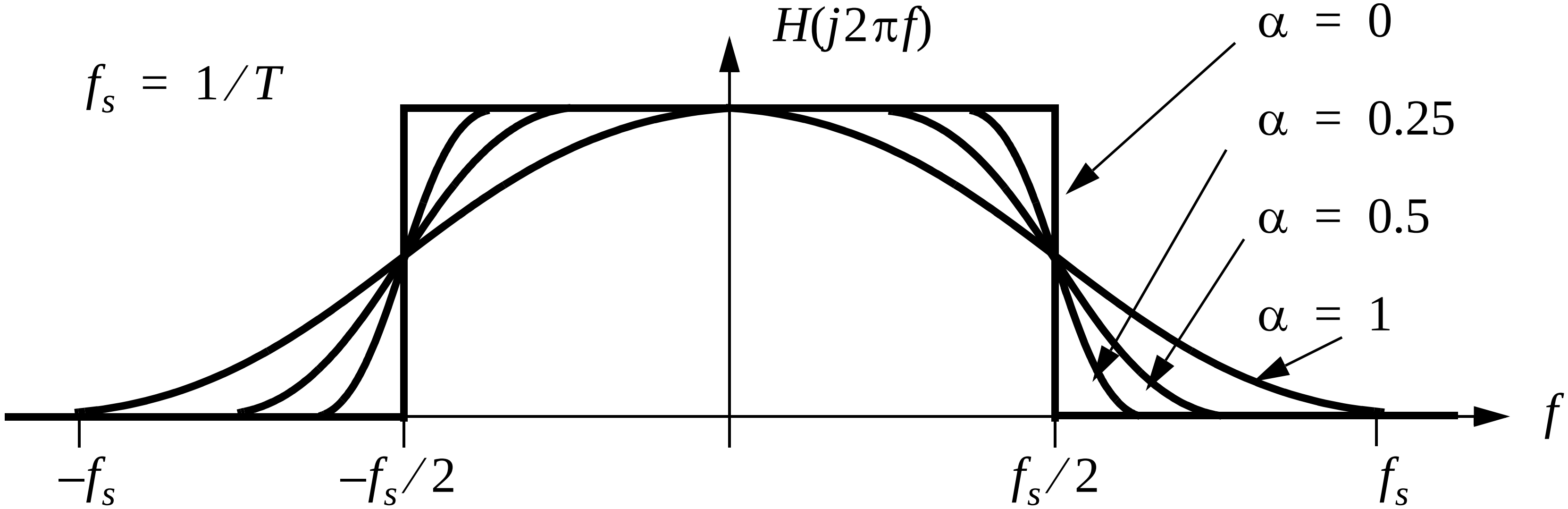


Raised-cosine pulse



Raised-cosine pulse

Raised-Cosine Pulse

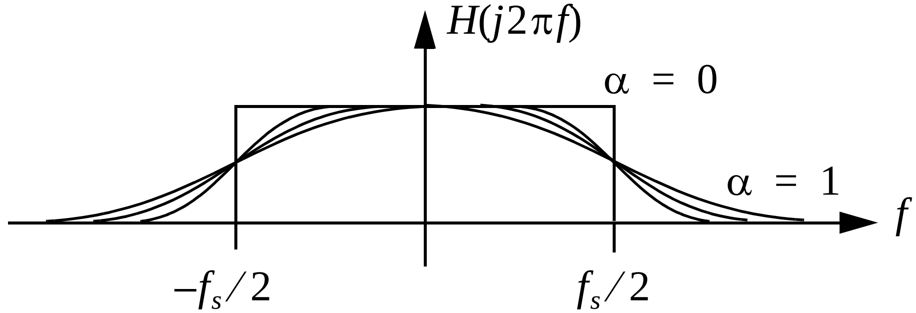


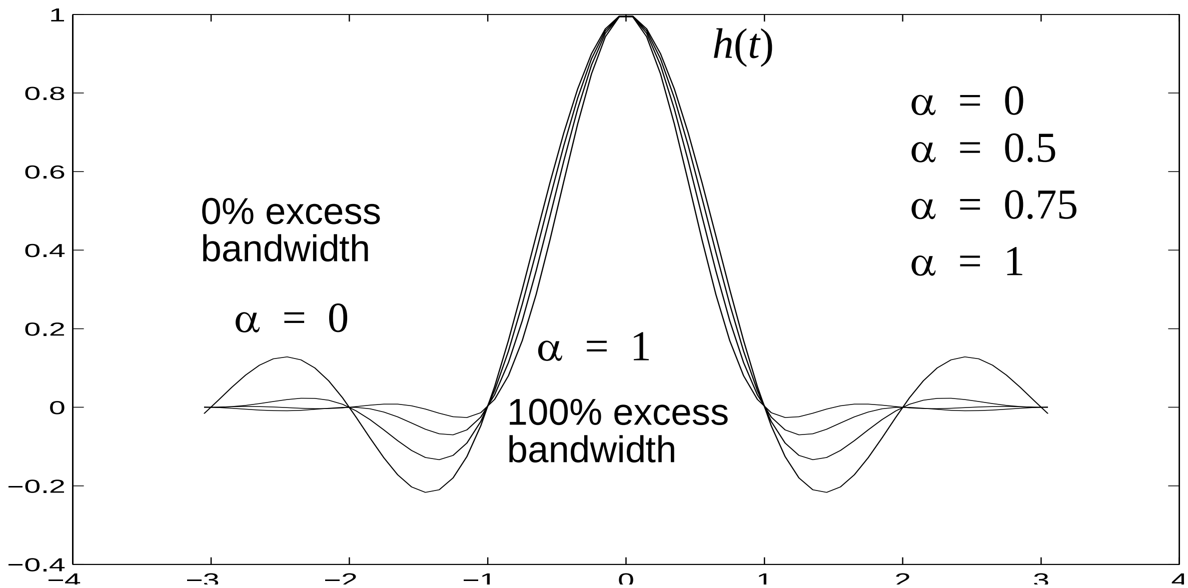
$f_s = 1/T$

$$H(j2\pi f) = \begin{cases} T; & 0 \le |f| \le (1-\alpha)\left(\frac{f_s}{2}\right) \\ \frac{T}{2}\left[1 + \cos\left[\frac{\pi}{2\alpha}\left(\frac{|2f|}{f_s} - (1-\alpha)\right)\right]\right] & (1-\alpha)\left(\frac{f_s}{2}\right) \le |f| \le (1+\alpha)\left(\frac{f_s}{2}\right) \\ 0; & |f| > (1+\alpha)\left(\frac{f_s}{2}\right) \end{cases}$$

- α determines ***excess bandwidth***

Raised-Cosine Pulses





- More excess bandwidth — impulse decays faster.

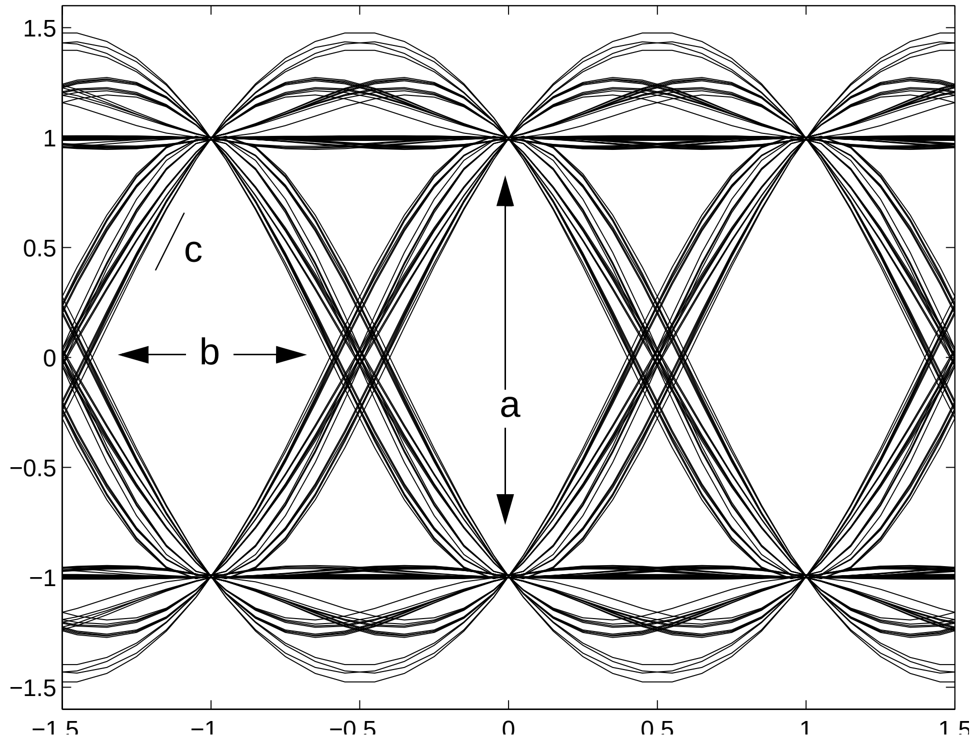
Raised-Cosine Pulse

- α determines amount of excess bandwidth past $f_s/2$
- Example: $\alpha = 0.25$ implies that bandwidth is 25 percent higher than $f_s/2$ while $\alpha = 1$ implies bandwidth extends up to f_s.
- Larger excess bandwidth — easier receiver
- Less excess bandwidth — more efficient channel use

Example

- Max symbol-rate if a 50% excess bandwidth is used and bandwidth is limited to 10kHz
- $1.5 \times (f_s/2) = 10\ \text{kHz}$ implies $f_s = 13.333 \times 10^3$ symbols/s

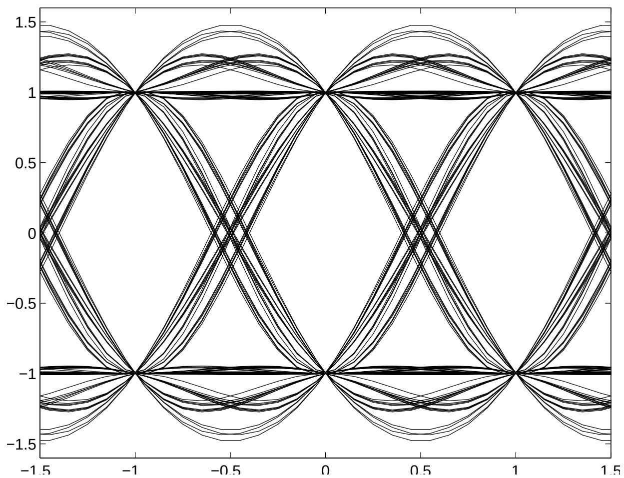
Eye Diagram



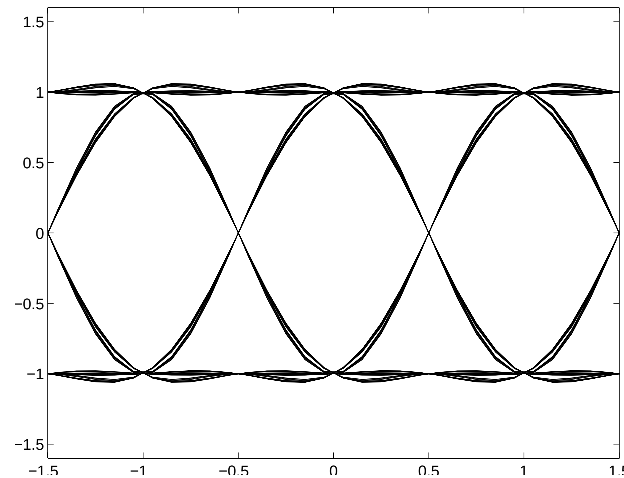
- "a" indicates immunity to noise
- "b" indicates immunity to errors in timing phase
- slope "c" indicates sensitivity to jitter in timing phase

Eye Diagram

- Zero crossing — NOT a good performance indicator
- 100% bandwidth has little zero crossing jitter
- 50% BW has alot of zero crossing jitter but it is using less bandwidth



$\alpha = 0.5$ 50% excess bandwidth



$\alpha = 1$ 100% excess bandwidth

- Less excess BW — more intolerant to timing phase

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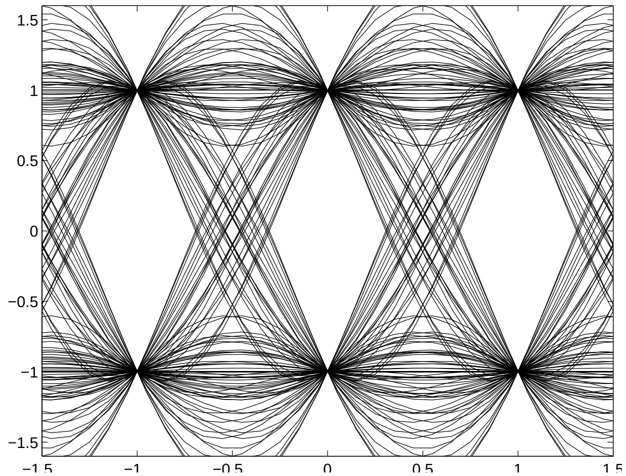
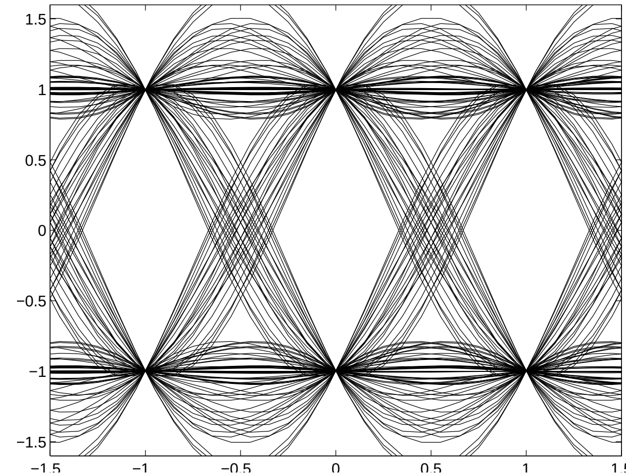
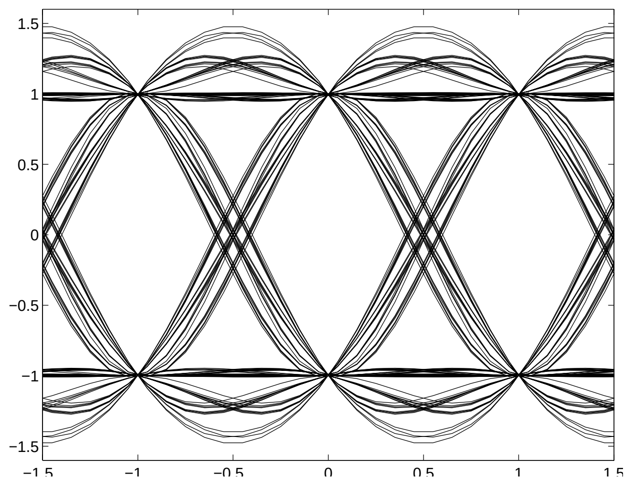
Example Eye Diagrams



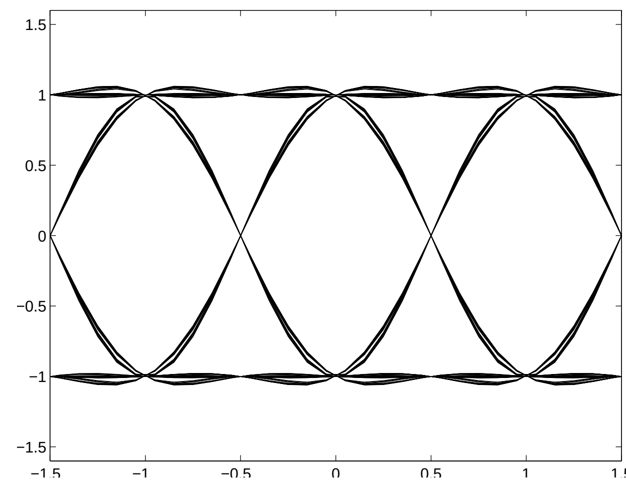
$\alpha = 0$ 0% excess bandwidth



$\alpha = 0.25$ 25% excess bandwidth

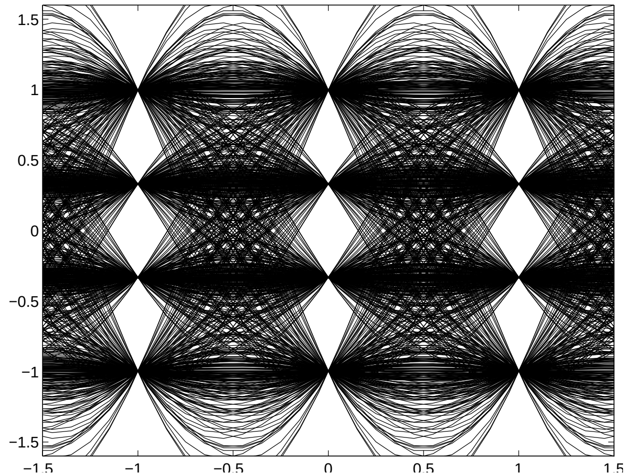


$\alpha = 0.5$ 50% excess bandwidth

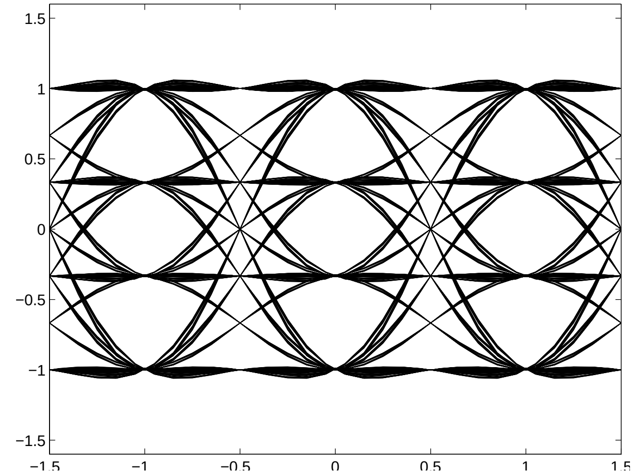


$\alpha = 1$ 100% excess bandwidth

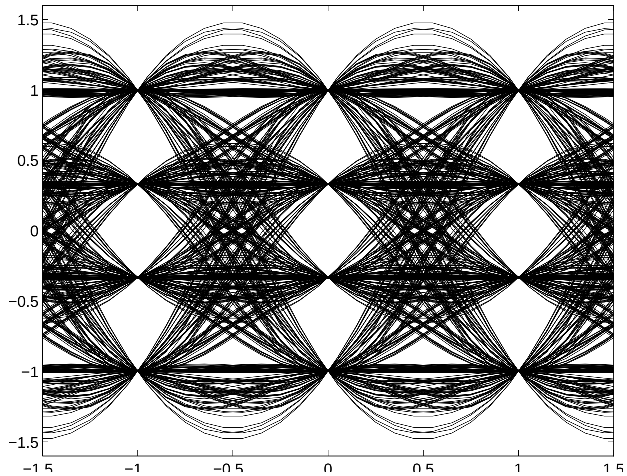
Example Eye Diagrams



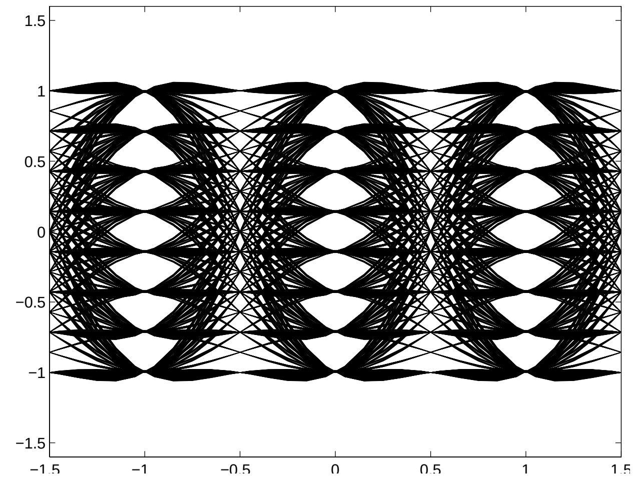
$\alpha = 0$ 0% excess bandwidth



$\alpha = 1$ 100% excess bandwidth

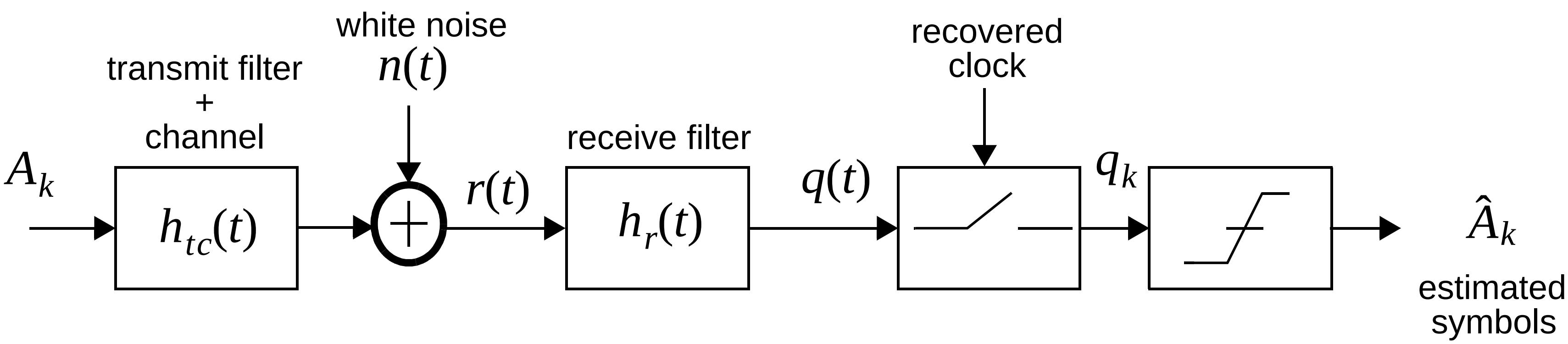


$\alpha = 0.5$ 50% excess bandwidth



$\alpha = 1$ 100% excess bandwidth

Matched-Filter



- For zero-ISI, $h_{tc}(t) \otimes h_r(t)$ satisfies Nyquist criterion.
- For optimum noise performance, $h_r(t)$ should be a ***matched-filter***.
- A matched-filter has an impulse response which is time-reversed of $h_{tc}(t)$

$$h_r(t) = K h_{tc}(-t) \tag{6}$$

where K is an arbitrary constant.

Matched-Filter (proof)

- Consider isolated pulse case (so no worry about ISI)

$$r(t) = A_0 h_{tc}(t) + n(t) \qquad (7)$$

$$q_0 = \int_{-\infty}^{\infty} r(\tau) h_r(t-\tau) d\tau \Big|_{t=0} = \int_{-\infty}^{\infty} r(\tau) h_r(-\tau) d\tau \qquad (8)$$

$$q_0 = A_0 \int_{-\infty}^{\infty} h_{tc}(\tau) h_r(-\tau) d\tau + \int_{-\infty}^{\infty} n(\tau) h_r(-\tau) d\tau \qquad (9)$$

- Want to maximize signal term to noise term
- Variance of noise is

$$\sigma_n^2 = N_0 \int_{-\infty}^{\infty} h_r^{\,2}(-\tau) d\tau \qquad (10)$$

Matched-Filter (proof)

- Assuming A_0 and $h_{tc}(t)$ fixed, want to maximize

$$\mathrm{SNR} = \frac{A_0^2\left[\int_{-\infty}^{\infty} h_{tc}(\tau)h_r(-\tau)d\tau\right]^2}{N_0^2\int_{-\infty}^{\infty} h_r{}^2(-\tau)d\tau} \tag{11}$$

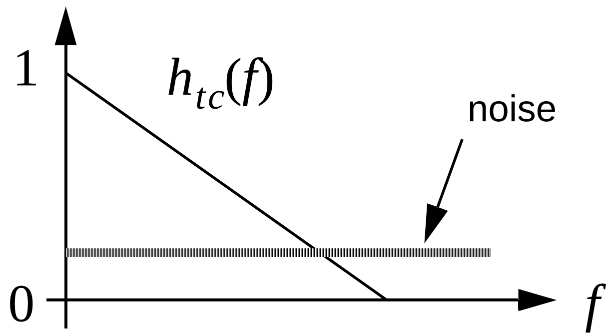
- Use Schwarz inequality

$$\left[\int_a^b f_1(x)f_2(x)dx\right]^2 \le \left[\int_a^b f_1^2(x)dx\right]\left[\int_a^b f_2^2(x)dx\right] \tag{12}$$

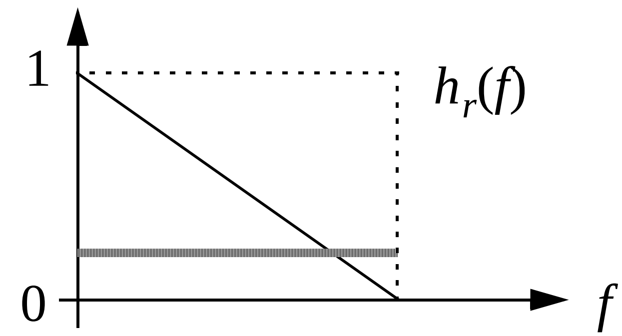
with equality if and only if $f_2(x) = Kf_1(x)$

- Maximizing (11) results in $h_r(t) = Kh_{tc}(-t)$ — QED

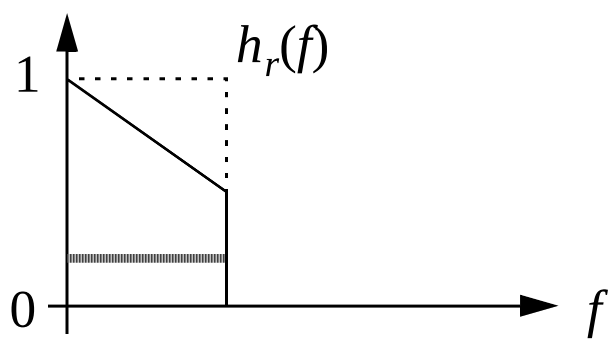
Matched-Filter — Why optimum?



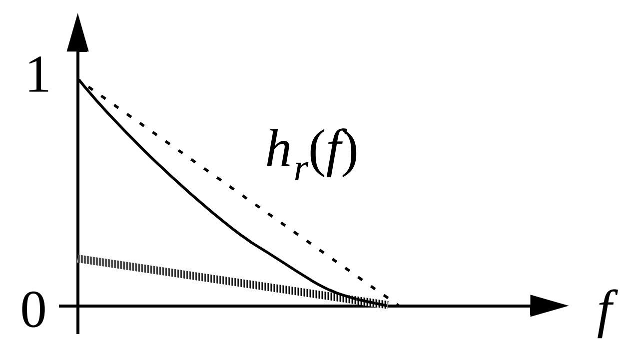
Transmit filter, channel and noise



Too much noise,
All of signal



Too little signal,
Less noise



Just right — max SNR

ISI and Noise

- In general, we need the output of a ***matched filter*** to obey Nyquist criterion
- Frequency response at output of matched filter is $|H_{tc}(j\omega)|^2$ leading to criterion

$$\frac{1}{T}\sum_{m=-\infty}^{\infty}|H_{tc}(j2\pi f + jm2\pi f_s)|^2 = 1 \tag{13}$$

Example

- Assume a flat freq resp channel and raised-cosine pulse is desired at matched-filter output
- Transmit filter should be $\sqrt{\text{raised-cosine}}$
- Receive filter should be $\sqrt{\text{raised-cosine}}$

Gaussian Noise and SNR Requirement

Probability Distribution Function

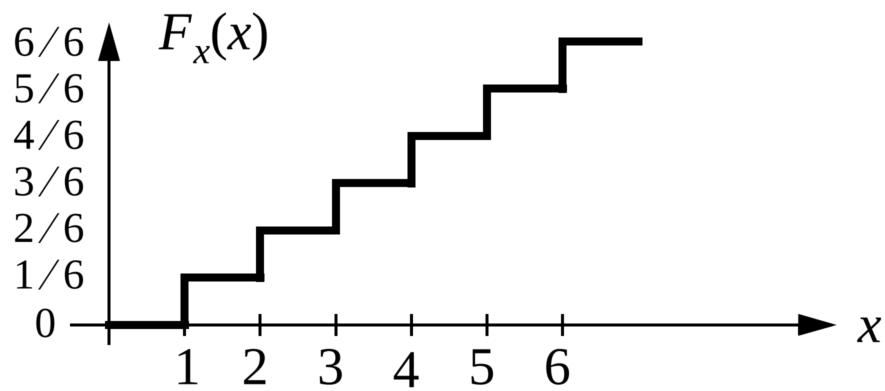
- Consider a random variable X
- Cumulative distribution function (c.d.f.) — $F_x(x)$

$$F_x(x) \equiv P_r(X \le x) \quad -\infty < x < \infty \tag{14}$$

$$1 \ge F_x(x) \ge 0 \tag{15}$$

Example

- Consider a fair die



Probability Density Function

- Derivative of $F_x(x)$ is p.d.f. defined as $f_x(x)$

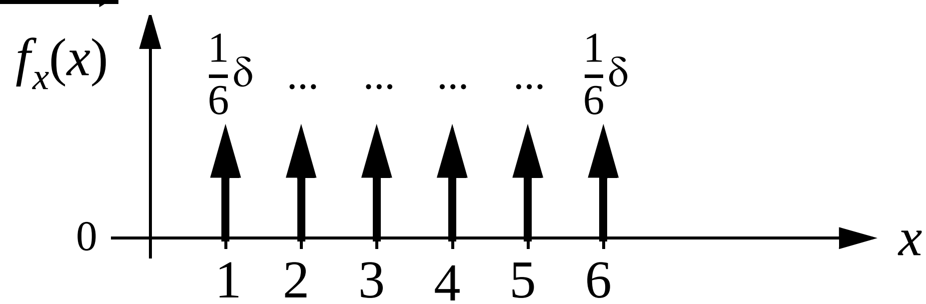
$$f_x(x) \equiv \frac{dF_x(x)}{dx} \qquad \text{or} \qquad F_x(x) = \int_{-\infty}^{\alpha} f_x(\alpha)d\alpha \tag{16}$$

- To find prob that X is between x_1 and x_2

$$P_r(x_1 < X \leq x_2) = \int_{x_1}^{x_2} f_x(\alpha)d\alpha \tag{17}$$

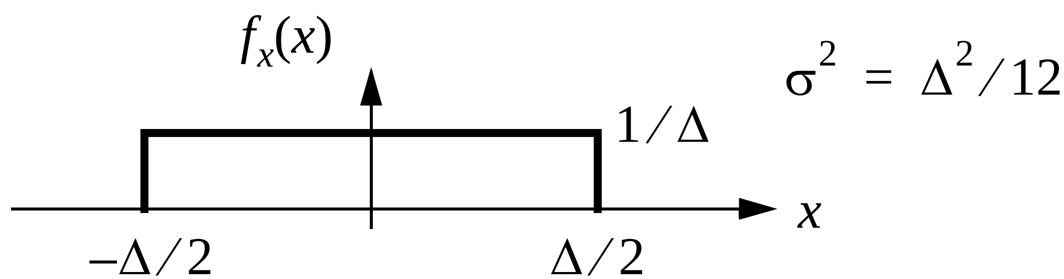
- It is the area under p.d.f. curve.

Example (fair die)



Uniform Distribution

- p.d.f. is a constant
- Variance is given by: $\sigma^2 = \frac{\Delta^2}{12}$ where Δ is range of random variables



- Crest factor: $CF \equiv \frac{\text{max}}{\sigma} = \frac{\Delta/2}{\Delta/\sqrt{12}} = \sqrt{3} = 1.732$

Example

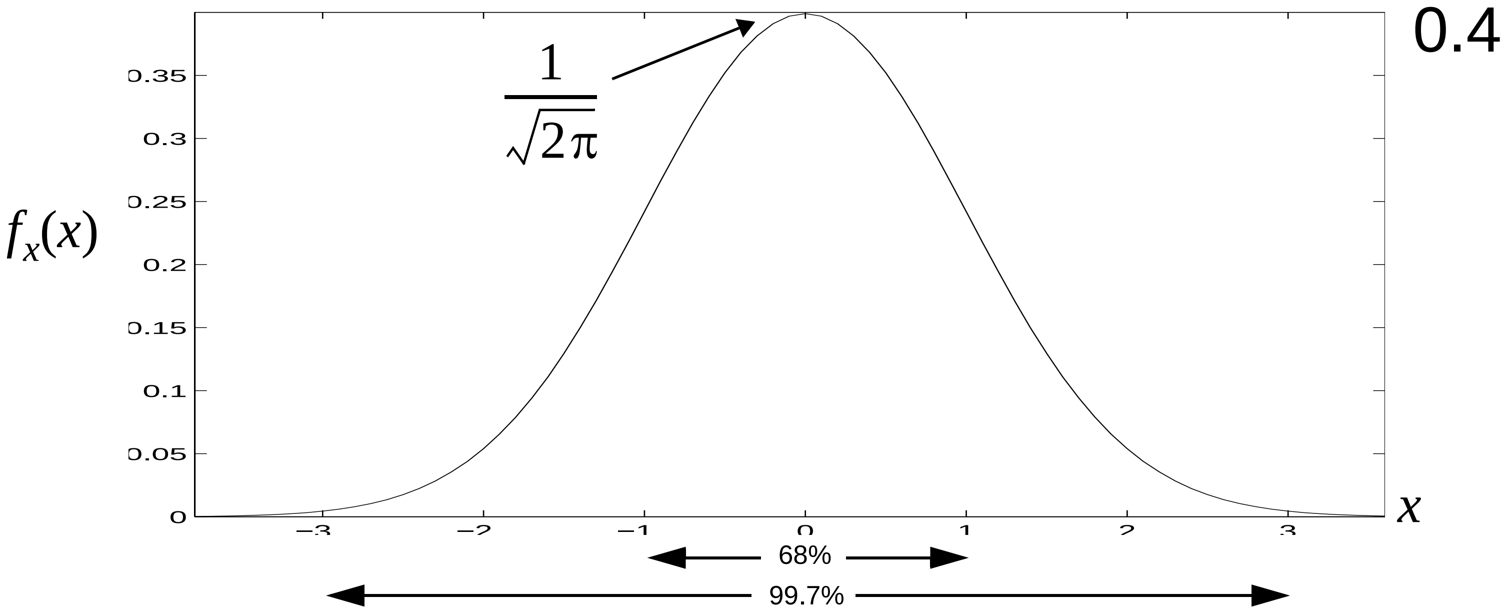
- A uniform random variable chosen between 0 and 1 has a mean, $\mu = 0.5$, and variance, $\sigma^2 = 1/12$

Gaussian Random Variables

Probability Density Function

- Assuming $\sigma^2 = 1$ (i.e. variance is unity) and $\mu = 0$ (i.e. mean is zero) then

$$f_x(x) = \frac{1}{\sqrt{2\pi}} e^{-x^2/2} \tag{18}$$

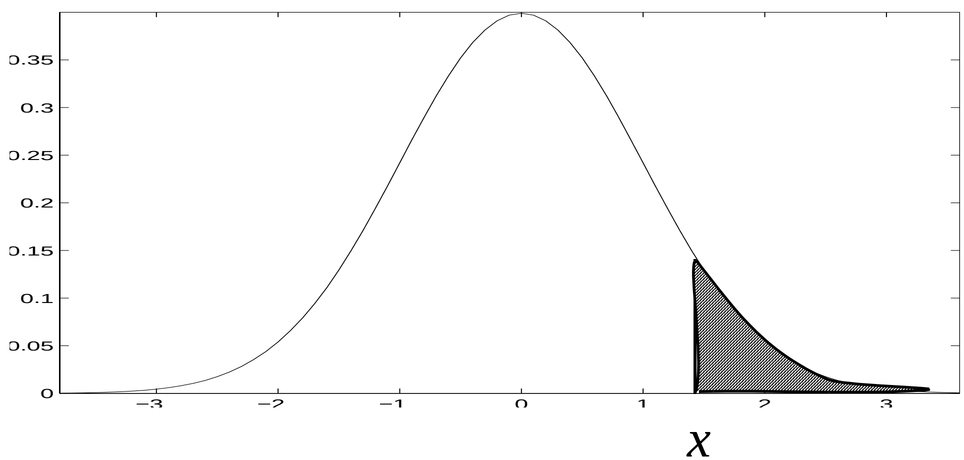


Gaussian Random Variables

- Often interested in how likely a random variable will be in tail of a Gaussian distribution

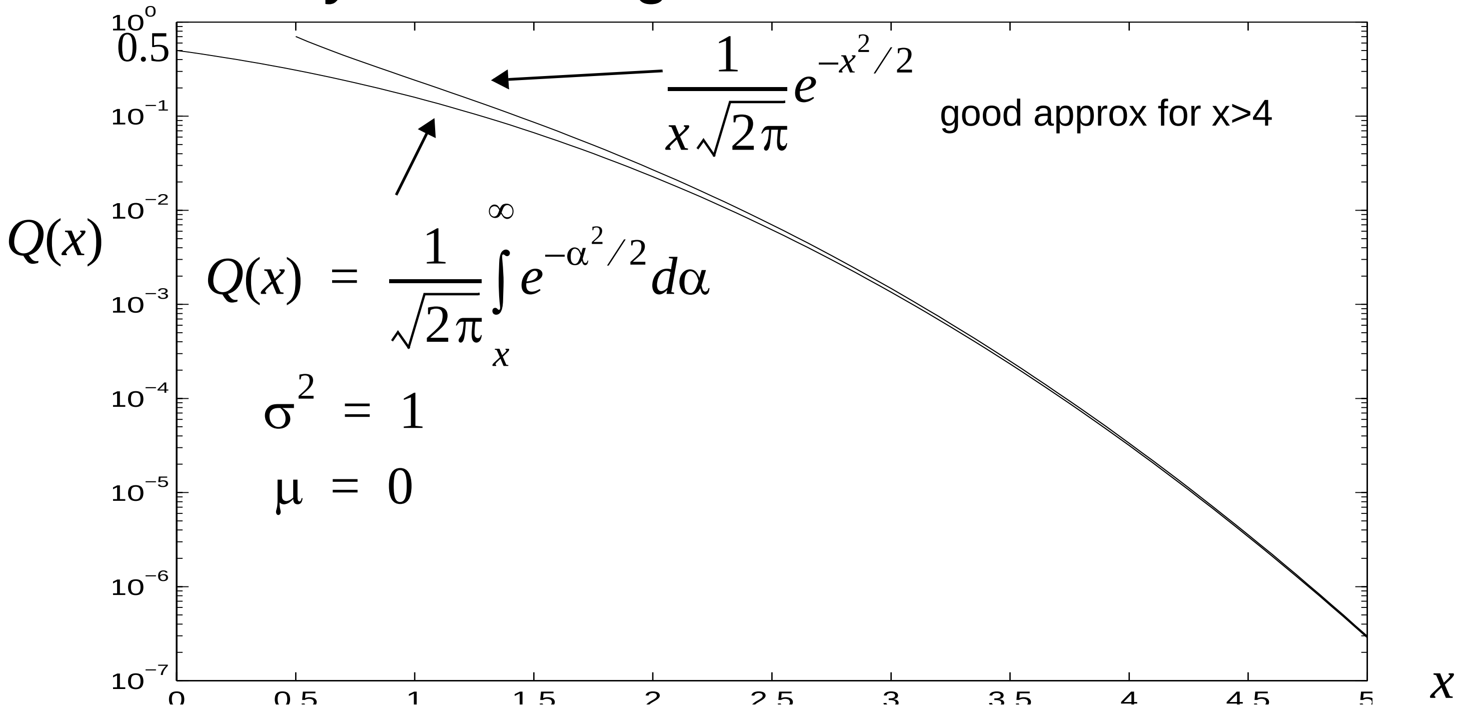
$$Q(x) \equiv P_r(X > x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-\alpha^2/2} d\alpha \qquad (19)$$

$$Q(x) = \frac{1}{2} \operatorname{erfc}(x/\sqrt{2}) \qquad (20)$$



Gaussian Random Variables

- Probability of x being in tail of Gaussian distribution

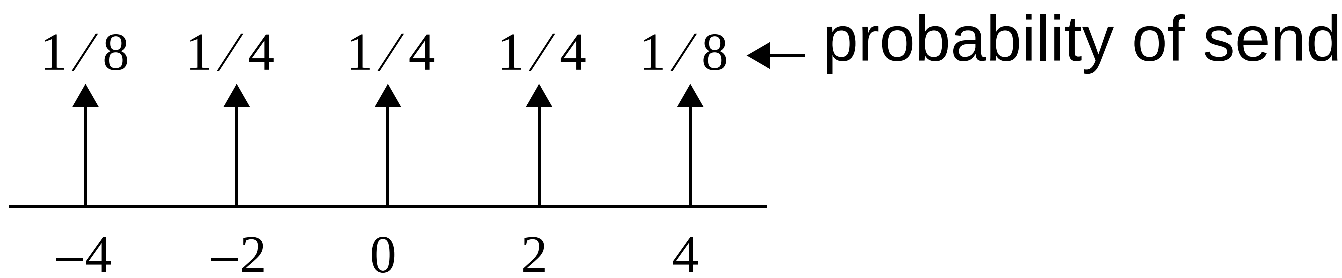


- If $\sigma^2 \neq 1$ or $\mu \neq 0$

$$P_r(X > x) = Q((x - \mu)/\sigma) \quad (21)$$

Example SNR Calculation

- 100Base-T2 for fast-ethernet uses 5-PAM
- Want to calculate the receive SNR needed for a symbol-error-rate of 10^{-10} (assume rest is ideal).



- Signal power, P_s

$$P_s = \frac{1}{4} \times 0W + \frac{1}{2} \times 4W + \frac{1}{4} \times 16W = 6W \tag{22}$$

- Using a reference of $1W$ as $0\mathrm{dB}$,

$$P_s = 10\log_{10}(6) = 7.78\mathrm{dB} \tag{23}$$

Example SNR Calculation

- Assume Gaussian noise added to receive signal.
- Since symbols are distance 2 apart, a noise value greater than 1 will cause an error in receive symbol.
- Want to find σ of Gaussian distribution such that likelihood of random variable greater than 1 is 10^{-10}.
- Recall

$$Q(x/\sigma) = 0.5\mathrm{erfc}((x/\sigma)/\sqrt{2}) \qquad (24)$$

- Let $x = 1$ and set

$$2Q(1/\sigma) = 10^{-10} \qquad (25)$$

(2 value because variable might be > 1 or < -1)

$$0.5 \times 10^{-10} = Q(1/\sigma) = 0.5\mathrm{erfc}(1/(\sigma\sqrt{2})) \qquad (26)$$

Example SNR Calculation

- Trial and error gives $1/(\sigma\sqrt{2}) = 4.57$ implying that $\sigma = 0.1547 = 1/6.46$
- Noise with $\sigma = 0.1547$ has a power of (ref to $1W$)

$$P_n = 10\log_{10}(\sigma^2) = -16.2\text{dB} \tag{27}$$

- Finally, SNR needed at receive signal is

$$\text{SNR} = 7.78\text{dB} - (-16.2\text{dB}) = 24\text{dB} \tag{28}$$

- Does not account that large positive noise on +4 signal will ***not*** cause symbol error (same on -4).
- It is slightly conservative
- BER approx same as symbol error rate if Gray coded

m-PAM

- For m bits/symbol $\Rightarrow$ 2^m levels
- Normalize distance between levels to 2 (so error of 1 causes a symbol error)
- $(m = 1) \Rightarrow \pm 1$ $(m = 3) \Rightarrow \pm 1, \pm 3, \pm 5, \pm 7$ etc.
- Noise variance of $(\sigma = 0.1547) \Rightarrow \mathrm{BER} = 10^{-10}$
- Symbols spaced $\pm 1, \pm 3, \pm 5, \ldots, \pm (2^m - 1)$
  — average power is: $S_m = (4^m - 1)/3$

$$\mathrm{SNR} = 10\log\left(\frac{S_m}{\sigma^2}\right) = 10\log\left(\frac{4^m - 1}{3\sigma^2}\right) \quad (29)$$

m-PAM

$$\text{SNR} = 10\log\left(\frac{S_m}{\sigma^2}\right) = 10\log\left(\frac{4^m - 1}{3\sigma^2}\right) \tag{30}$$

- equals 23.1 dB for $m = 2$, $\text{BER} = 10^{-10}$
- equals 28.2 dB for $m = 3$, $\text{BER} = 10^{-10}$ (approx +6dB)
- Can show $S_{m+1} = 4S_m + 1$
- Require 4 times more power to maintain same symbol error rate with same noise power (uncoded)
- In other words,
  — to send 1 more bit/symbol, need 6dB more SNR
  (but does not increase bandwidth)

Why Assume Gaussian Noise?

Central-Limit Theorem

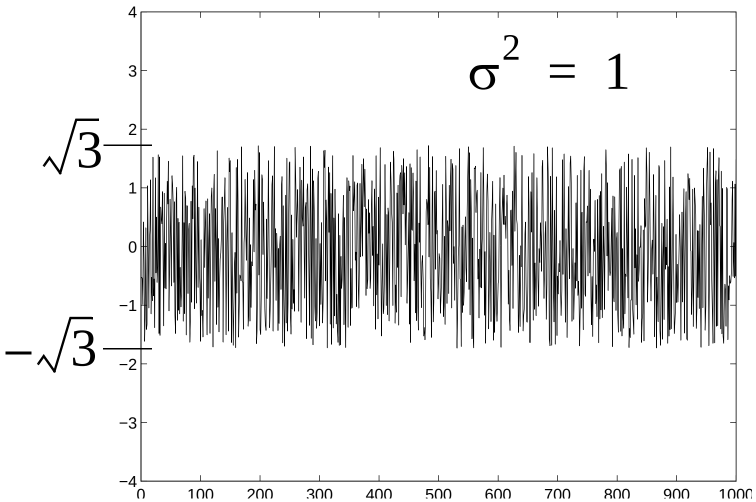
- Justification for modelling many random signals as having a Gaussian distribution

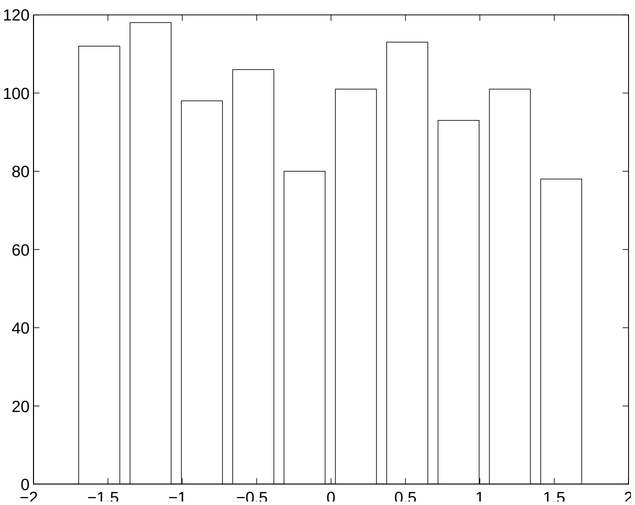
> Sum of independent random variables approaches Gaussian as sum increases

- Assumes random variables have identical distributions.
- No restrictions on original distribution (except finite mean and variance).
- Sum of Gaussian random variables is also Gaussian.

Uniform and Gaussian Signals

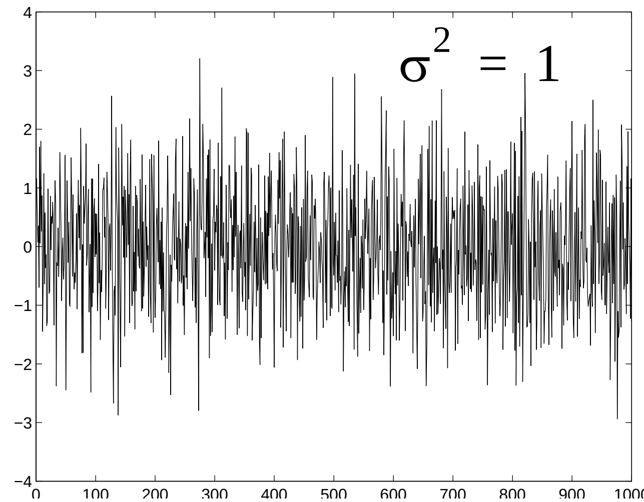
1000 samples of uniform random variables

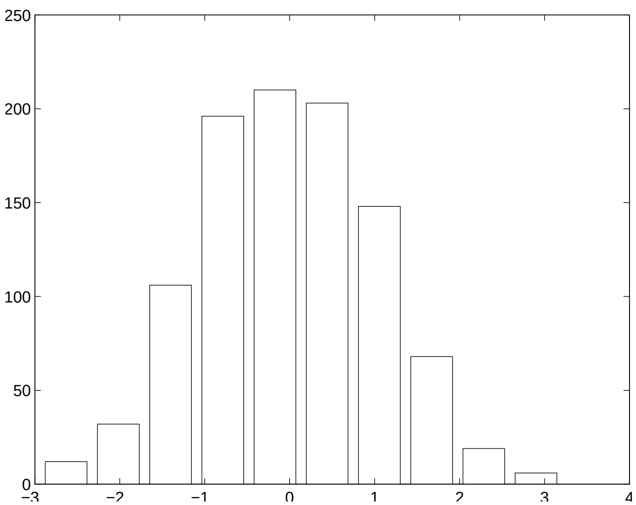




histogram of uniform variable

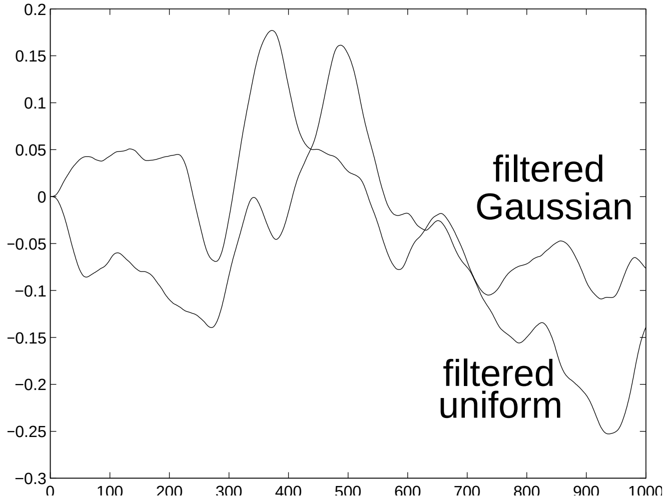
1000 samples of Gaussian random variables





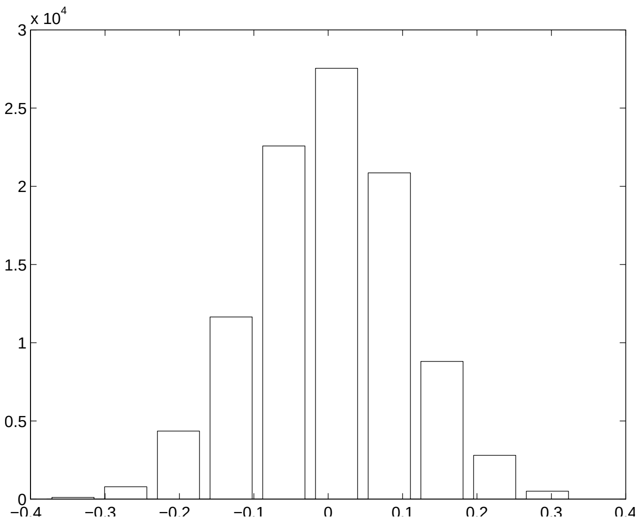
histogram of Gaussian variable

Filtered Random Signals

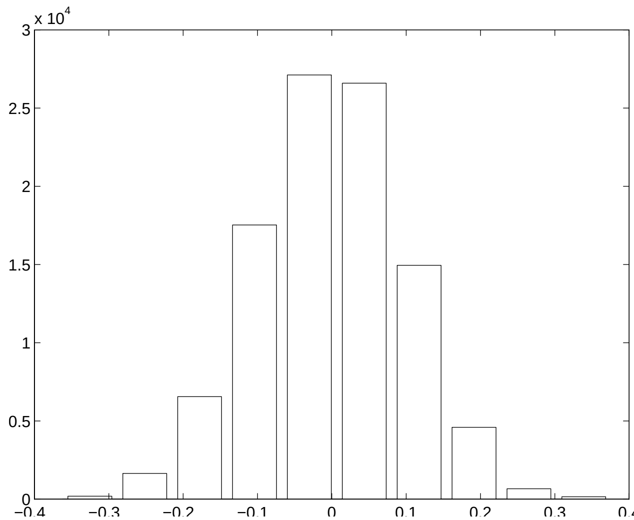


Filtered with 3’rd order Butterworth lowpass with cutoff $f_s / 200$

No longer independent from sample to sample



histogram of filtered uniform
(100,000 samples)



histogram of filtered Gaussian
(100,000 samples)

Wired Digital Communications

Wired Digital Transmission

Long Twisted-Pair Applications (1km - 6km)

- T1/E1 — 1.5/2Mb/s (2km)
- ISDN — Integrated Services Digital Network
- HDSL — High data-rate Digital Subscriber Line
- ADSL — Asymmetric DSL
- VDSL — Very high data-rate DSL

Short Twisted-Pair Applications (20m - 100m)

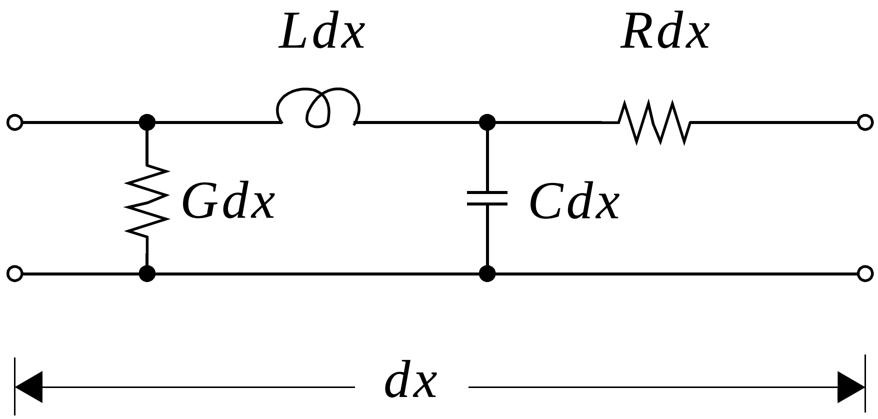
- 100Mb/s Fast-Ethernet — TX, T4, T2
- Gigabit Ethernet — Short haul, Long haul

Short Coax (300m)

- Digital video delivery — 300Mb/s - 1.5Gb/s

Cable Modelling

- Modelled as a transmission line.



Twisted-Pair Typical Parameters:

- $R(f) = (1 + j)\sqrt{f/4}\ \Omega/\text{km}$ due to the skin effect
- $L = 0.6\ \text{mH/km}$ (relatively constant above 100kHz)
- $C = 0.05\ \mu\text{F/km}$ (relatively constant above 100kHz)
- $G = 0$

Skin Effect

- "Resistance" is not constant with frequency and is complex valued.
- Can be modelled as:

$$R(\omega) = k_R(1 + j)\sqrt{\omega} \tag{31}$$

where k_R is a constant given by

$$k_R = \frac{1}{\pi d_c}\left(\frac{\mu}{2\sigma}\right)^{1/2} \tag{32}$$

- d_c is conductor diameter, μ is permeability, σ is conductivity
- Note resistance is inversely proportional to d_c.

• Jordan and Balmain, "Electromagnetic Waves and Radiating Systems", pg. 563, Prentice-Hall, 1968.

Characteristic Impedance

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tag{33}$$

- Making use of (31) and assuming $G = 0$

$$Z_0 = \left(\frac{k_R\sqrt{\omega}(1+j) + j\omega L}{j\omega C}\right)^{1/2} \tag{34}$$

$$Z_0 = \sqrt{\frac{L}{C}}\left(1 + \frac{k_R}{L\sqrt{\omega}}(1-j)\right)^{1/2} \tag{35}$$

Now using approx $(1+x)^{1/2} \approx 1 + x/2$ for $x \ll 1$

$$Z_0 \approx \sqrt{\frac{L}{C}} + \frac{k_R}{2\sqrt{\omega LC}}(1-j) \tag{36}$$

- At high freq, Z_0 appears as constant value $\sqrt{L/C}$

Characteristic Impedance

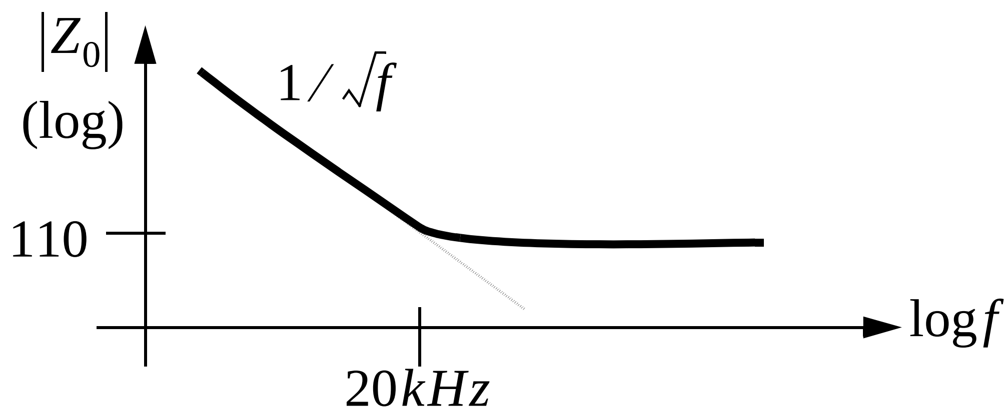
- From (33), when $\omega L \gg R$ (typically $\omega \gg 2\pi \times 16kHz$)

$$Z_{0h} = \sqrt{\frac{L}{C}} \qquad (37)$$

resulting in

$$Z_{0h} \approx 110\ \Omega \qquad (38)$$

- Thus, when terminating a line, a resistance value around $110\ \Omega$ should be used.



Cable Transfer-Function

- When properly terminated, a cable of length d has a transfer-function of

$$H(d, \omega) = e^{-d\gamma(\omega)} \tag{39}$$

where $\gamma(\omega)$ is given by

$$\gamma(\omega) = \sqrt{(R + j\omega L)(G + j\omega C)} \tag{40}$$

- Breaking $\gamma(\omega)$ into real and imaginary parts,

$$\gamma(\omega) \equiv \alpha(\omega) + j\beta(\omega) \tag{41}$$

$$H(d, \omega) = e^{-d\alpha(\omega)} e^{-jd\beta(\omega)} \tag{42}$$

- $\alpha(\omega)$ determines ***attenuation***.
- $\beta(\omega)$ determines ***phase***.

Cable Transfer-Function

- Assuming $G = 0$, then from (40)

$$\gamma = (j\omega CR - \omega^2 LC)^{1/2} \tag{43}$$

- Substituting in (31)

$$\gamma = (j\omega^{1.5} k_R C(1+j) - \omega^2 LC)^{1/2} \tag{44}$$

$$\gamma = j\omega\sqrt{LC}\left(1 + \frac{k_R}{L\sqrt{\omega}}(1-j)\right)^{1/2} \tag{45}$$

Now using approx $(1+x)^{1/2} \approx 1 + x/2$ for $x \ll 1$

$$\gamma \approx \frac{k_R}{2}\sqrt{\frac{\omega C}{L}} + j\left(\omega\sqrt{LC} + \frac{k_R}{2}\sqrt{\frac{\omega C}{L}}\right) \tag{46}$$

Cable Attenuation

- Equating (41) and (46)

$$\alpha(\omega) \approx \frac{k_R}{2}\sqrt{\frac{C}{L}} \times \sqrt{\omega} \tag{47}$$

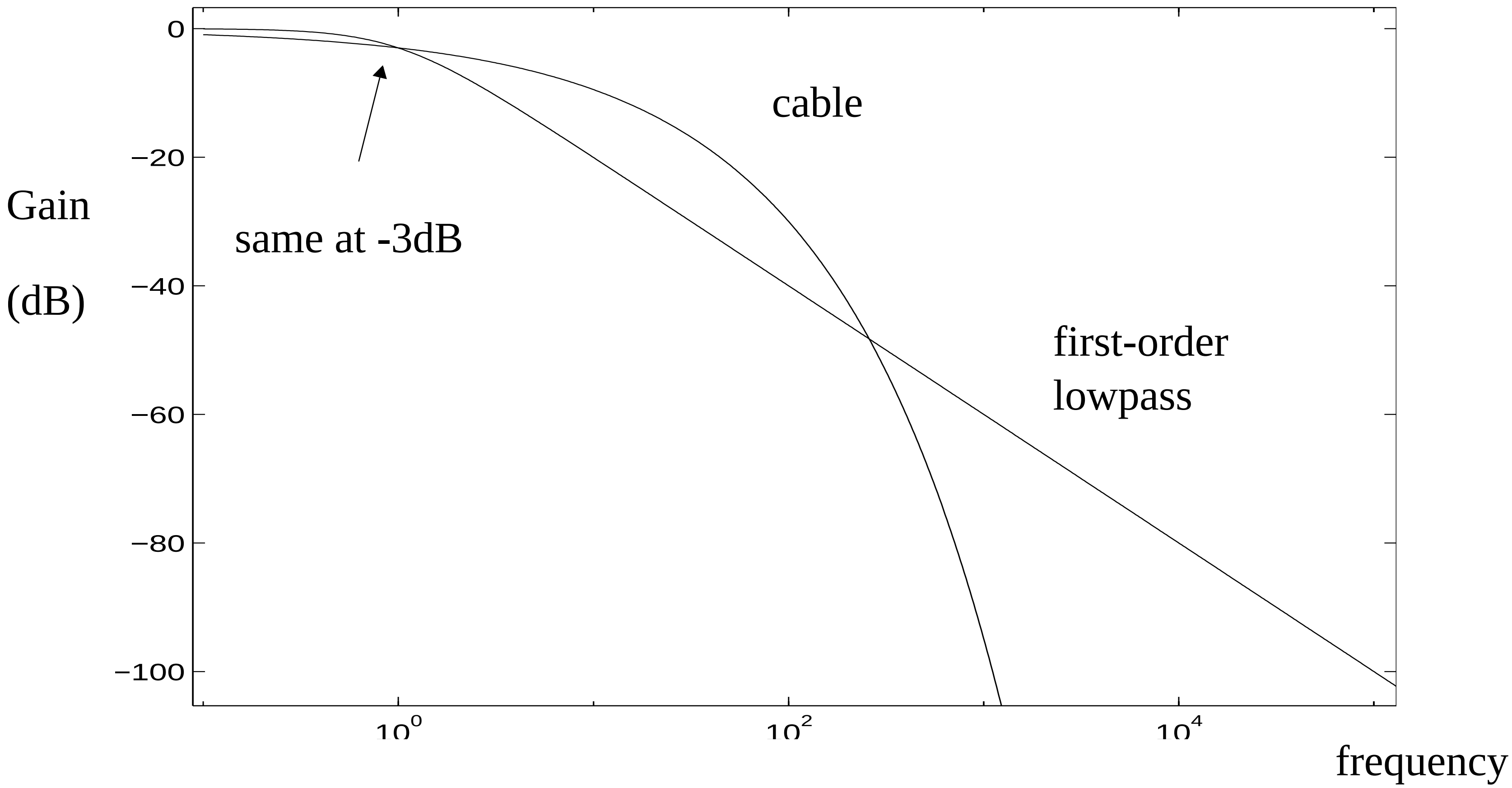
- Therefore gain in dB is

$$H_{dB}(d, \omega) \approx -8.68d \times \frac{k_R}{2}\sqrt{\frac{C}{L}} \times \sqrt{\omega} \tag{48}$$

- Note that attenuation in dB is proportional to cable length (i.e. 2x distance doubles attenuation in dB)
- Can reduce attenuation by using a larger diameter cable
- Attenuation proportional to root-frequency

Cable Attenuation

- Gain in dB is proportional to $\sqrt{f}$ due to skin effect.



- Do not confuse with 1/f noise slow frequency roll-off.

Cable Phase

- Equating (41) and (46)

$$\beta(\omega) \approx \omega\sqrt{LC} + \frac{k_R}{2}\sqrt{\frac{C}{L}} \times \sqrt{\omega} \qquad (49)$$

- The linear term usually dominates
- The linear term implies a constant group delay.
- In other words, the linear term simply accounts for the delay through the cable.
- Ignoring linear phase portion, remaining phase is proportional to $\sqrt{f}$.
- Note it has the same multiplying term as attenuation.

IIR Filter Cable Match using Matlab

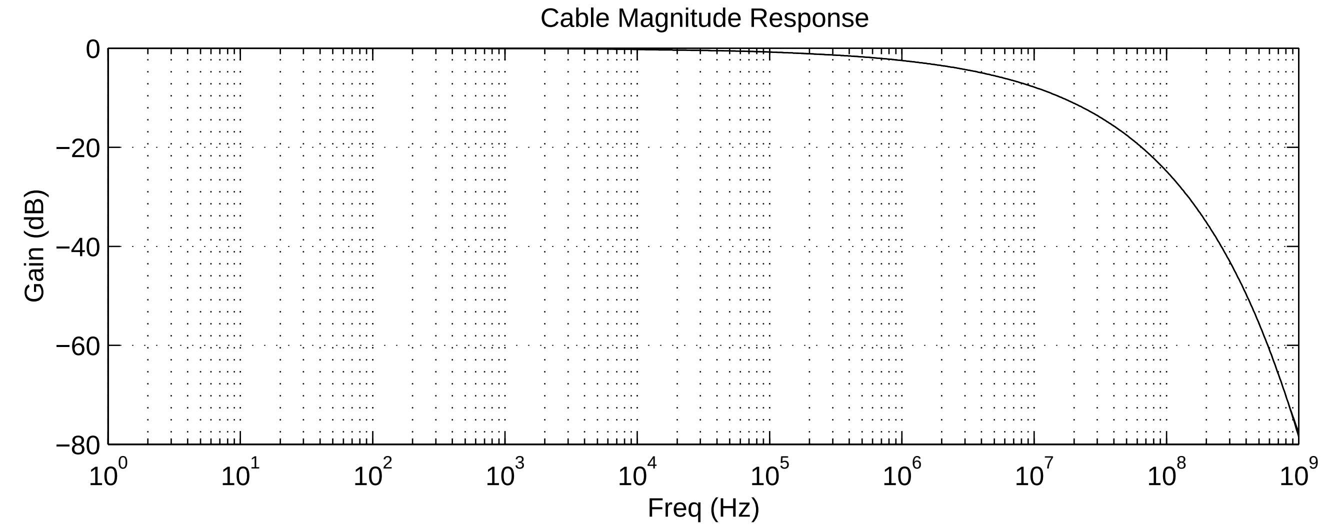
```
% this program calculates an iir num/den transfer-function
% approx for a transmission line with exp(sqrt(s)) type response.
clear;

% Order of IIR filter to match to cable
% nz is numerator order and np is denominator order
nz = 9;
np = 10;

% important parameters of cable
c = 0.05e-6 % capacitance per unit length in farads/km
l = 0.6e-3 % inductance per unit length in henries/km
kr = 0.25 % resistance per unit length in ohms/km (times (1+j)*sqrt(omega))
d = 0.1 % cable length in km
% above values adjusted to obtain -20dB atten for 100m at 125MHz
k_cable = (kr/2)*sqrt(c/l);

% the frequency range for finding tf of cable
fmin=1;
fmax=1e9;
```

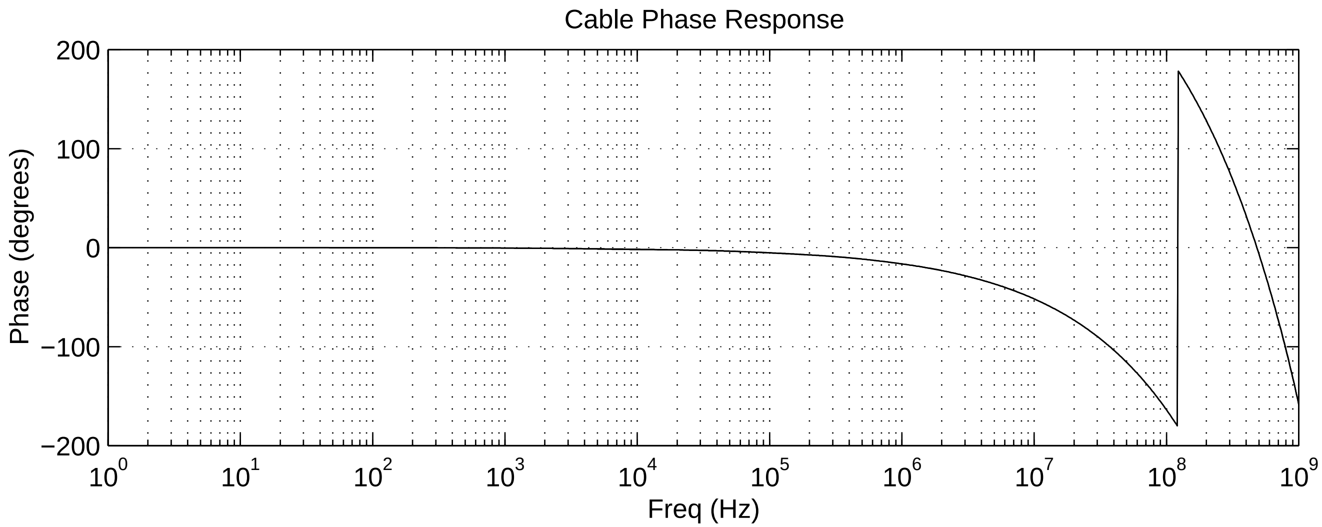
```
% specify frequency points to deal with
nmax=1000;
f=logspace(log10(fmin), log10(fmax), nmax);
w=2*pi*f;
s=j*w;

% 'cable' is desired outcome in exponential form
cable = exp(-d*k_cable*sqrt(2)*sqrt(s));

% Perform IIR approximate transfer-function match
% Since invfreqs miminizes (num-cable*den)
% first need an approximate den so that it can be used
% as a freq weighting to minimize (num/den - cable)
[num,den]=invfreqs(cable,w,nz,np, 1./w);
[denor]=freqs(den,1,w);
% re-iterate process with weighting for the denominator
% which now minimizes (num/den - cable)
[num,den]=invfreqs(cable,w,nz,np, (1./denor).^2);
[denor]=freqs(den,1,w);
[num,den]=invfreqs(cable,w,nz,np, (1./denor).^2);

% find approximate transfer function 'cable_approx' to 'cable'
[cable_approx]=freqs(num,den,w);
```

```
% also find pole-zero model
[Z,p,k]=tf2zp(num,den);


% PLOT RESULTS
clf;
figure(1);
subplot(211);
semilogx(f,20*log10(abs(cable)),'r');
hold on;
semilogx(f,20*log10(abs(cable_approx)),'b');
title('Cable Magnitude Response');
xlabel('Freq (Hz)');
ylabel('Gain (dB)');
grid;

hold off;
subplot(212);
semilogx(f,angle(cable)*180/pi,'r');
hold on;
semilogx(f,angle(cable_approx)*180/pi,'b');
title('Cable Phase Response');
xlabel('Freq (Hz)');
```

```
ylabel('Phase (degrees)');
grid;

hold off;
figure(2);
subplot(211);
semilogx(f,20*log10(abs(cable)./abs(cable_approx)));
title('Gain Error Between Cable and Cable_approx');
xlabel('Freq (Hz)');
ylabel('Gain Error (dB)');
subplot(212);
semilogx(f,(angle(cable)-angle(cable_approx))*180/pi);
title('Phase Error Between Cable and Cable_approx');
xlabel('Freq (Hz)');
ylabel('Phase Error (degrees)');
grid;
```

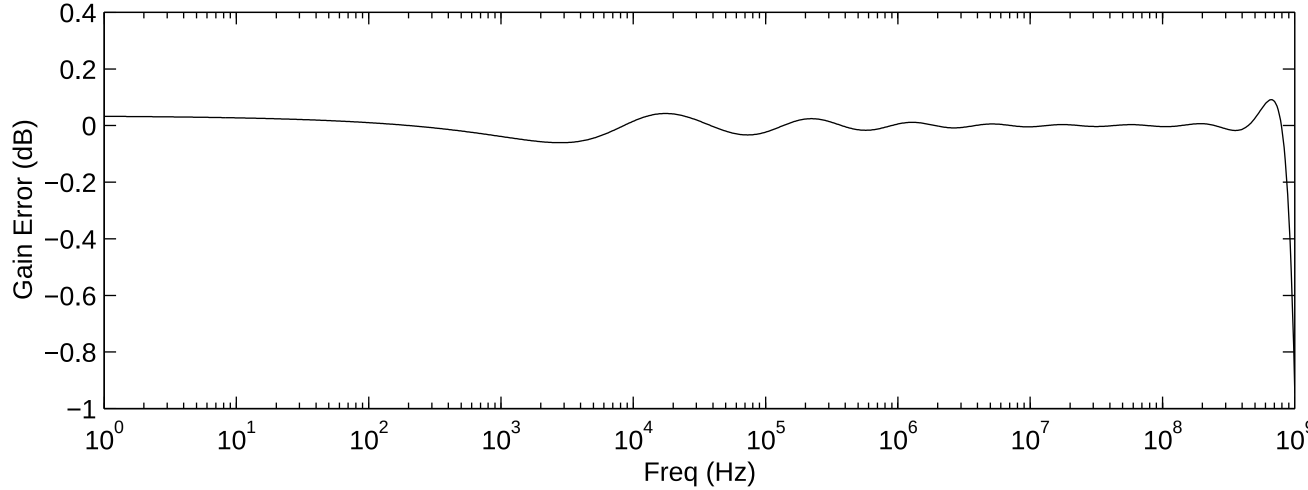
Cable Response



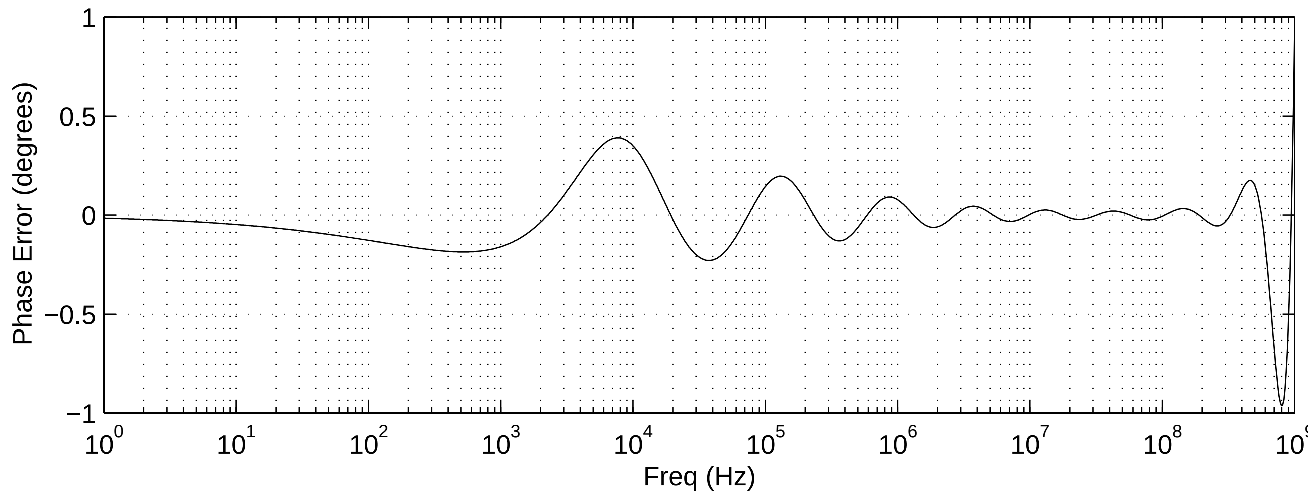


IIR Matching Results

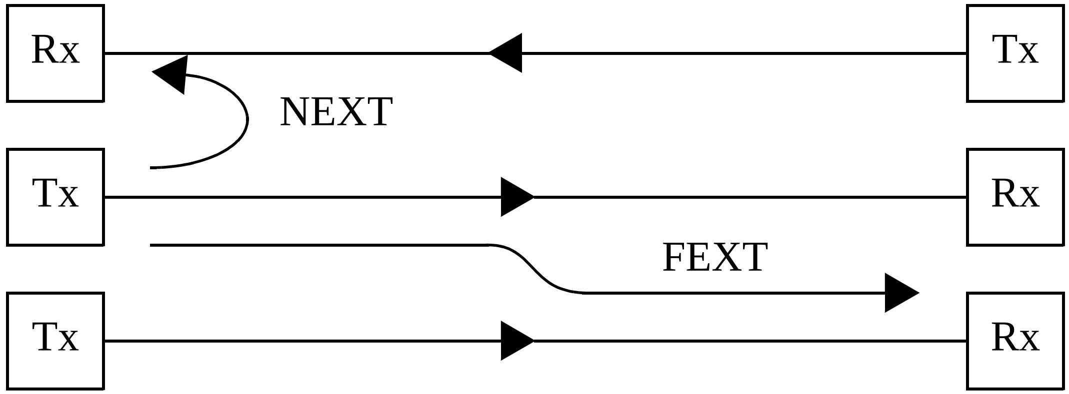
Gain Error Between Cable and Cable$_a$pprox



Phase Error Between Cable and Cable$_a$pprox



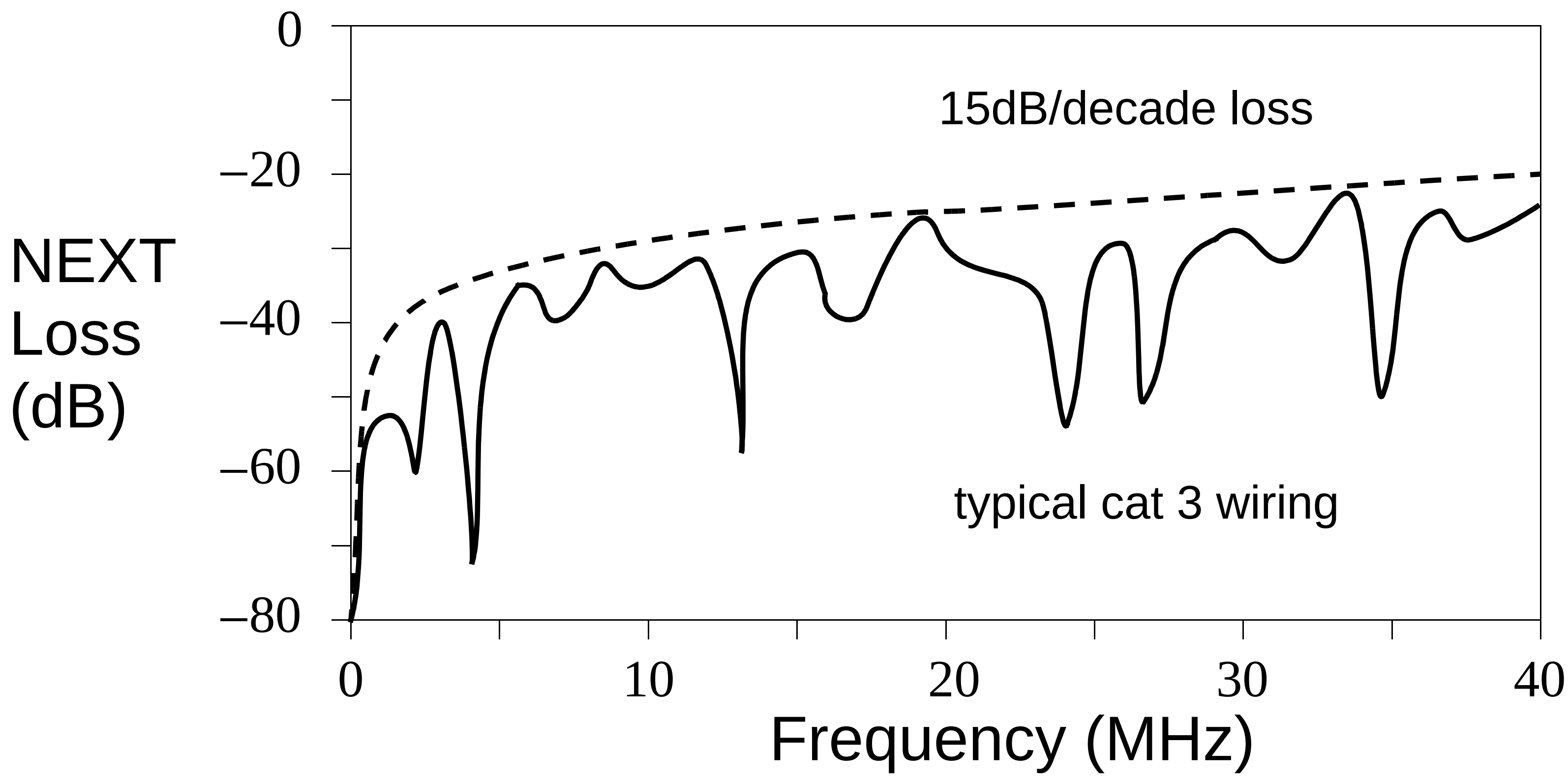
Near and Far End Crosstalk



- In FEXT, interferer and signal both attenuated by cable
- In NEXT, signal attenuated but interferer is coupled directly in.
- When present, NEXT almost always dominates.
- Can cancel NEXT if nearby interferer is known.
- Envelope of squared gain of NEXT increases with $f^{1.5}$

Twisted-Pair Crosstalk

- Crosstalk depends on turns/unit length, insulator, etc.
- Twisted-pairs should have different turns/unit length within same bundle



Transformer Coupling

- Almost all long wired channels (>10m) are AC coupled systems
- AC coupling introduces ***baseline wander*** if random PAM sent
- A long string of like symbols (for example, +1) will decay towards zero degrading performance
- Requires baseline wander correction (non-trival)
- Can use passband modulation schemes (CAP, QAM, DMT)
- ***Why AC couple long wired channels??***

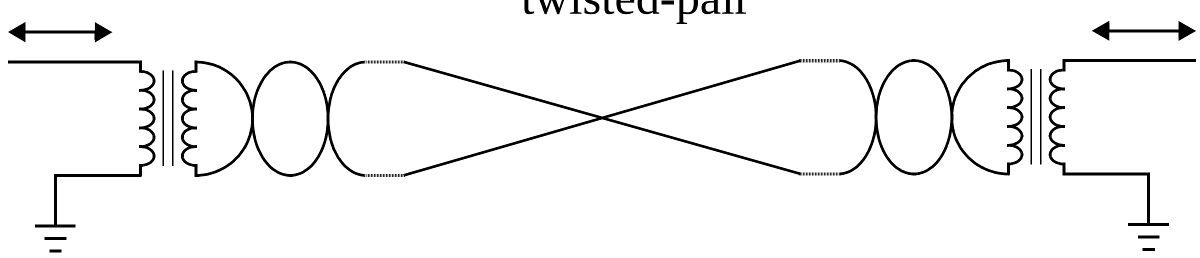
Transformer Coupling

Eliminates need for similar grounds

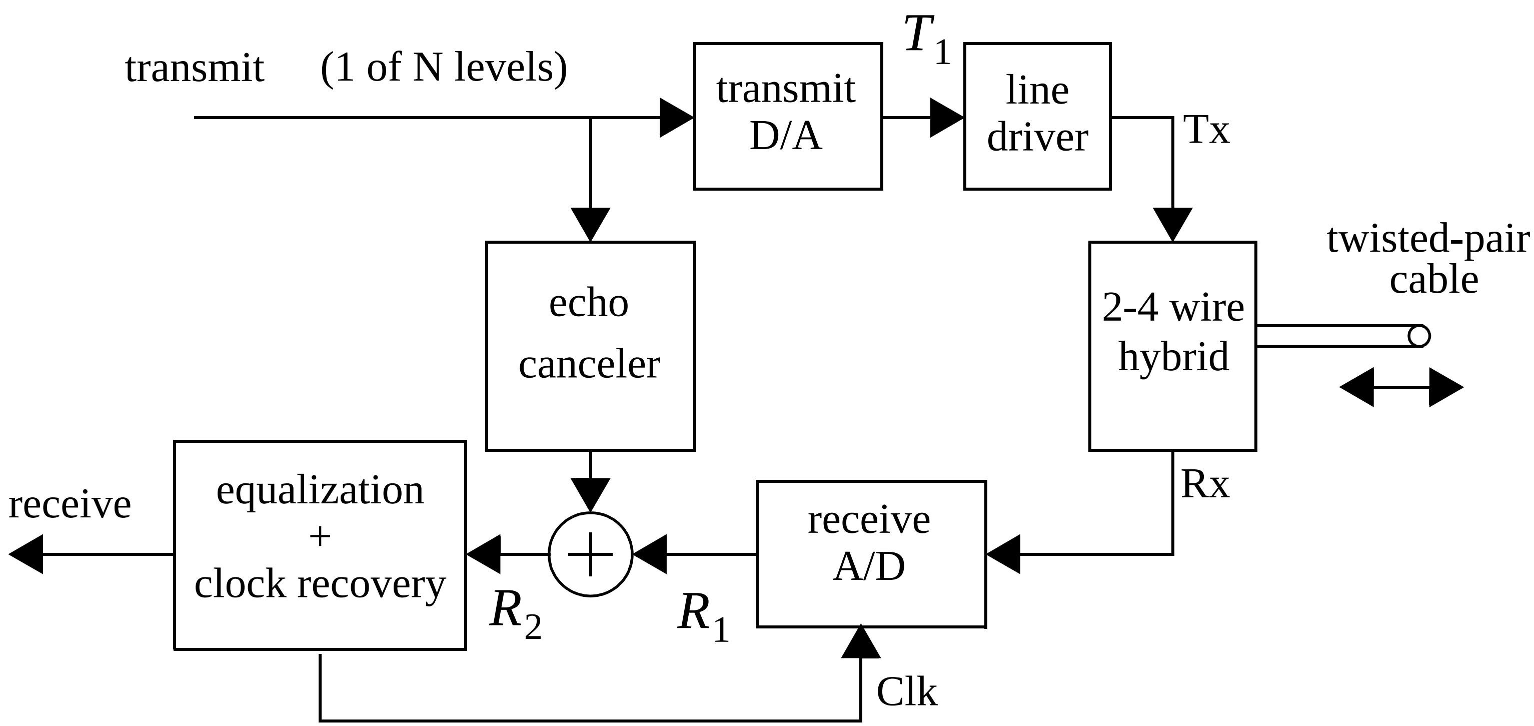
- If ground potentials not same — large ground currents

Rejects common-mode signals

- Transformer output only responds to differential signal current
- Insensitive to common-mode signal on both wires



Generic Wired PAM Transceiver



- Look at approaches for each block

HDSL Application

- 1.544Mb/s over 4.0km of existing telephone cables.
- Presently 4-level PAM code (2B1Q) over 2 pairs (a CAP implementation also exists).
- Symbol-rate is 386 ksymbols/s

Possible Bridged-Taps

- Can have unterminated taps on line
- Modelling becomes more complicated but DFE equalizes effectively
- Also causes a wide variation in input line impedance to which echo canceller must adapt — difficult to get much analog echo cancellation

HDSL Application

- Symbol-rate is 386 ksymbols/s

Received Signal

- For $d = 4km$, a 200kHz signal is attenuated by $40dB$.
- Thus, high-freq portion of a 5Vpp signal is received as a 50mVpp signal — ***Need effective echo cancellation***

Transmit Path

- Due to large load variations, echo cancellation of analog hybrid is only 6dB
- To maintain 40dB SNR receive signal, linearity and noise of transmit path should be better than 74dB.

ISDN Application

- Similar difficulty to HDSL but lower frequency
- 160kb/s over 6km of 1 pair existing telephone cables
- 4-level PAM coding — 2B1Q
- Receive signal at 40kHz atten by 40dB
- Requires highly linear line-drivers + A/D converters for echo cancellation (similar to HDSL)

Fast-Ethernet Application

CAT3	CAT5
$H_{dB}(f) = 2.32\sqrt{f} + 0.238f$	$H_{dB}(f) = 1.967\sqrt{f} + 0.023f + 0.05/\sqrt{f}$
$12.5MHz \leftrightarrow 11dB$	$12.5MHz \leftrightarrow 7dB$
crosstalk worse	crosstalk better

100Base-T4

- 4 pair CAT3 — 3 pair each way, 25MS/s with coding

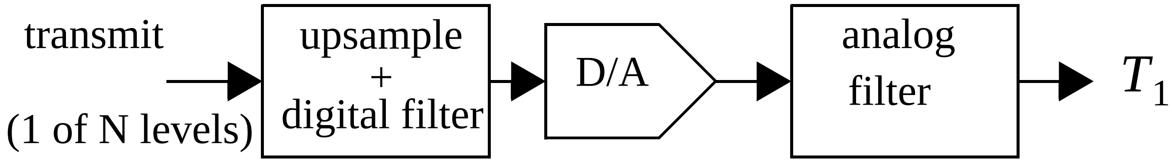
100Base-TX

- 2 pair CAT5 — 3 level PAM to reduce radiation

100Base-T2

- 2 pair CAT3 — 5x5 code, 25MS/s on each pair

Typical Transmit D/A Block



- Polyphase filter to perform upsampling+filtering

HDSL

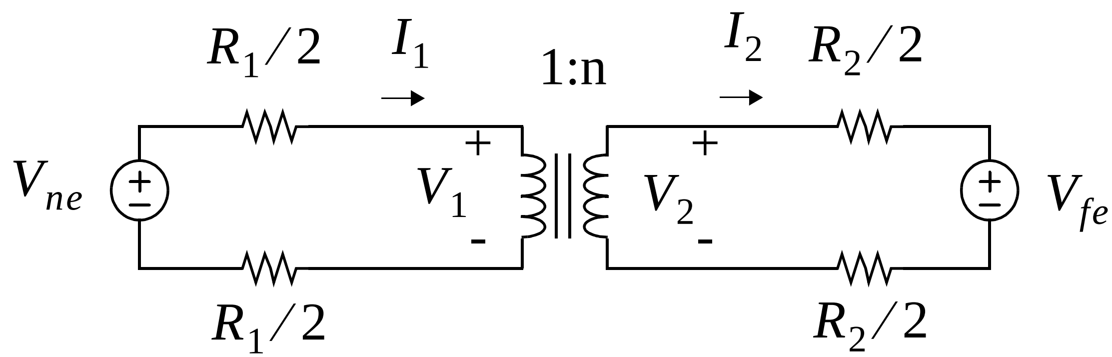
- D/A and filter needs better than 12-bit linearity
- Might be an oversampled 1-bit DAC
- One example: $\uparrow 16$; 48 tap FIR; $\uparrow 4$; $\Delta\Sigma$ DAC

Fast-Ethernet

- Typically around 35 dB linearity + noise requirement
- 100Base-T2 example: $\uparrow 3$; simple FIR; 75MHz 4-bit DAC; 3’rd-order LP cont-time filter

Line Drivers

- Line driver supplies drive current to cable.
- Commonly realized as voltage buffers.
- Often the most challenging part of analog design.
- Turns ratio of transformer determines equivalent line impedance.



$$V_{ne} = \frac{2}{n}V_2$$

$$V_1 = V_2/n$$

$$I_1 = nI_2$$

$$R_1 = R_2/n^2$$

Typical Values

$$R_2 = 100\Omega$$

$$V_2 = \pm 2.5\text{V}$$

$$I_2 = \pm 25\text{mA}$$

Line Driver Efficiency

- Efficiency improves as power supply increased

Example (assume can drive within 1V of supplies)

- From typical values, max power delivered by line driver is $P_{\mathrm{line+R}} = 2 \times 2.5 \times 25\mathrm{mA} = 125\mathrm{mW}$

12V Case

- Consider 12V supply — use $n = 0.5$, $V_{ne,\,\max} = 10V$, $I_{1,\max} = 12.5\mathrm{mA}$ leading to $P = 12 \times 12.5\mathrm{mA} = 150\mathrm{mW}$ (and drive an 800 ohm load)

3V Case

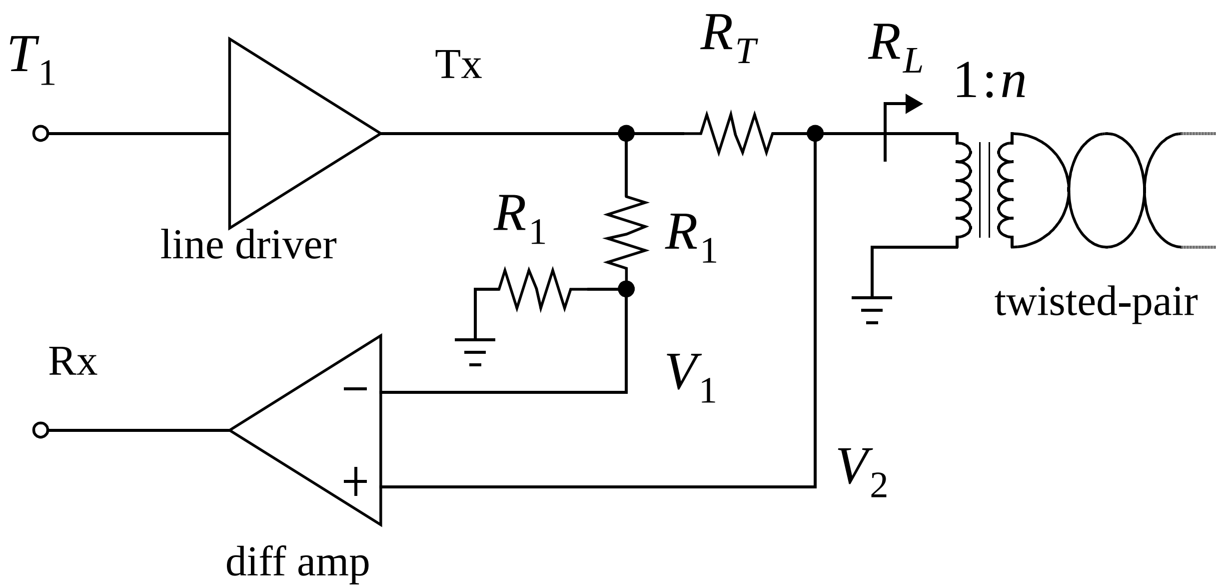
- Consider 3V supply — use $n = 5$, $V_{ne,\,\max} = 1V$, $I_{1,\max} = 125\mathrm{mA}$ leading to $P = 3 \times 125\mathrm{mA} = 375\mathrm{mW}$ (and drive an 8 ohm load!!!)

Line Driver

- In CMOS, W/L of output stage might have transistors on the order of 10,000!
- Large sizes needed to ensure some gain in final stage so that feedback can improve linearity — might be driving a 30 ohm load
- When designing, ensure that enough phase margin is used for the wide variation of bias currents
- Nested Miller compensation has been successfully used in HDSL application with class AB output stage
- Design difficulties will increase as power supplies decreased

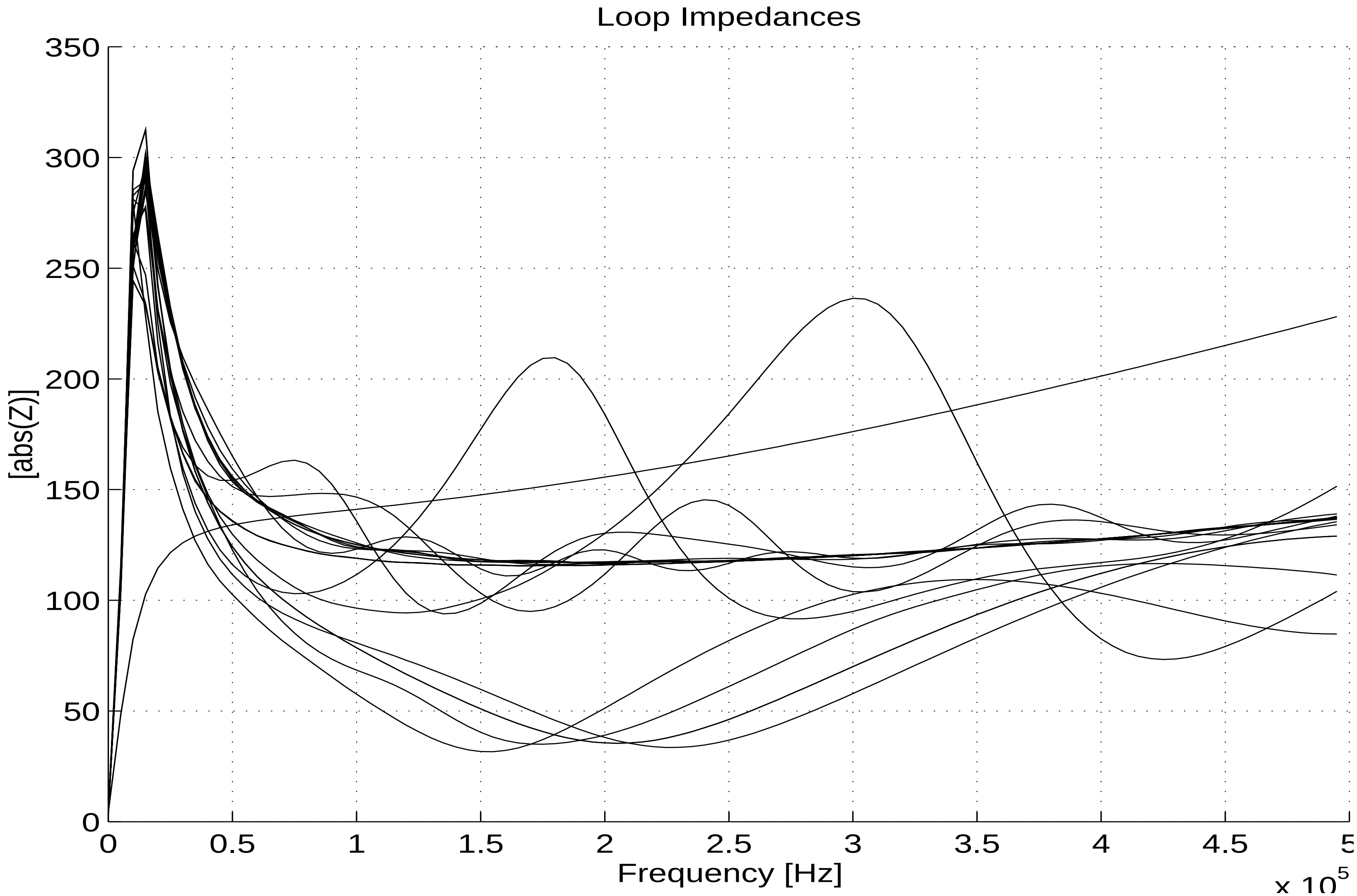
2-4 Wire Hybrids

- Dual-duplex often used to reduce emission.
- However, dual-duplex requires hybrids and echo cancellation.



- If $R_L = R_T$, no echo through hybrid
- Can be large impedance variation.

Typical HDSL Line Impedances



Hybrid Issues

- Note zero at dc and pole at 10kHz.
- Low frequency pole causes long echo tail (HDSL requires 120 tap FIR filter)

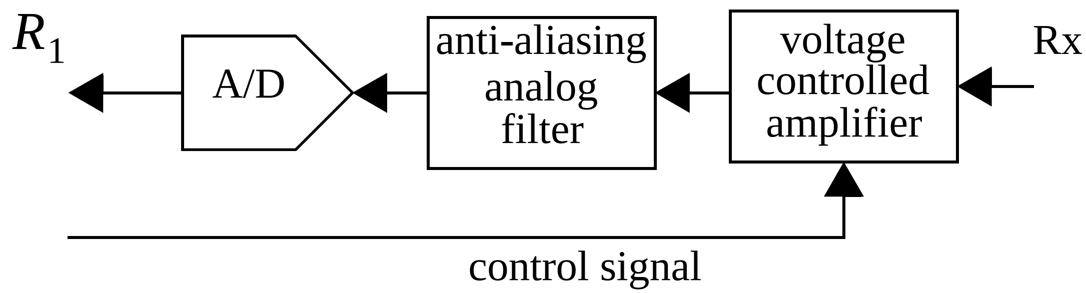
Alternatives

- Could eliminate R_1 circuit and rely on digital echo cancellation but more bits in A/D required.

OR

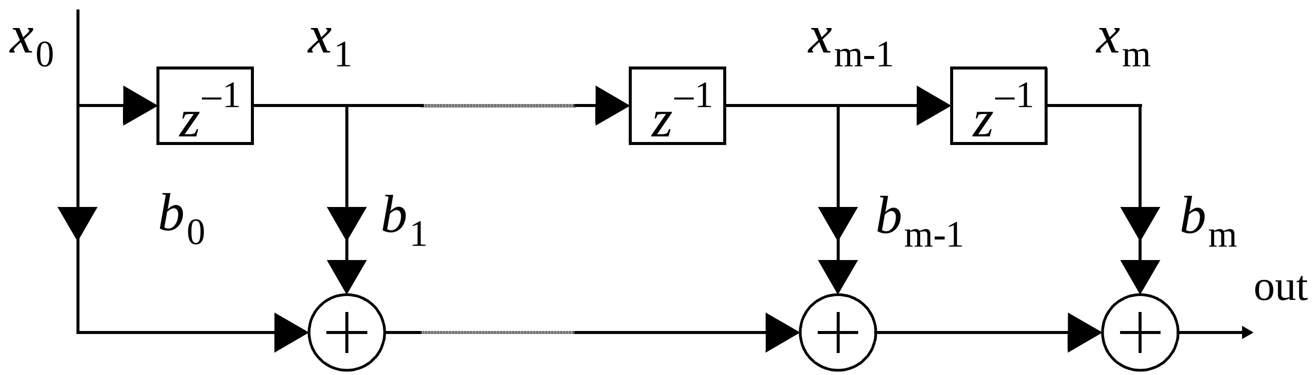
- Can make R_1 circuit more complex to ease A/D specs.
- Less echo return eases transmit linearity spec.
- Might be a trend towards active hybrids with or without extra A/D and D/A converters (particularly for higher speeds).

Typical Receive A/D



- Often, VGA is controlled from digital signal.
- Anti-aliasing can be simple in oversampled systems.
- Continuous-time filters are likely for fast-ethernet
- Example: 100Base-T2 suggests a 5'th order cont-time filter at 20MHz with a 6-bit A/D at 75MHz.
- Challenge here is to keep size and power of A/D small.

Echo Cancellation



$$b_i(k+1) = b_i(k) + \mu e(k) x_i(k) \quad \text{LMS}$$

- Typically realized as an adaptive FIR filter.
- Note input is transmit signal so delay lines and multiplies are trivial.
- HDSL uses about a 120 tap FIR filter
- Coefficient accuracy might be around 20 bits for dynamic range of 13 bits.

Echo Cancellation

- Fast-ethernet might be around 30 taps and smaller coefficient accuracy
- Can also perform some NEXT cancellation if signal of nearby transmitter is available (likely in 100Base-T2 and gigabit ethernet)

Alternatives

- Higher data rates may have longer echo tails.
- Might go to FIR/IIR hybrid to reduce complexity.
- Non-linear echo cancellation would be VERY useful in reducing transmit linearity spec.
- However, these non-linearities have memory and thus Volterra series expansions needed.

Equalization

HDSL

- Echo canceller required ***before*** equalization so fractional spaced equalizer not practical
- Typically 9 tap FFE and 120 tap DFE
- Long DFE also performs dc recovery (baseline wander)

Fast Ethernet

- Often fractional-spaced EQ - 30 taps
- DFE — 20 taps (dc recovery)