Passband HDSL and ADSL Circuits and Systems

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Outline

Baseband Review and Limitations

- Cable Modeling
- Equalization and DFE
- dc Recovery and sinusoidal interference

Passband QAM/CAP HDSL and ADSL

- Basic Concepts
- Equalization
- Timing Recovery
- HDSL and ADSL Applications
- Line Interface Issues

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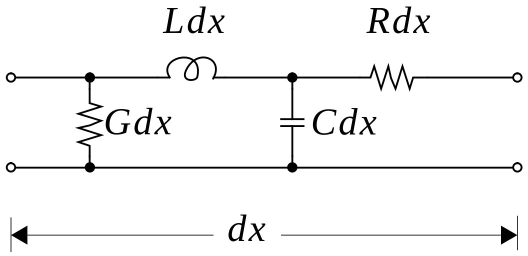
Cable Modeling

- Modeled as a transmission line.



Twisted-Pair Typical Parameters:

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

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Cable Attenuation

- Cable gain in dB is

$$\boxed{H_{dB}(d, \omega) \approx -k_R \times d \times \sqrt{\omega}} \quad (1)$$

- k_R — cable constant (typically 0.008)
  d — cable distance in km
  ω — frequency in rad/s
- Attenuation in dB is proportional to cable length
  — 2x distance doubles attenuation in dB
  — reduce atten by using larger diameter cable
- Attenuation also proportional to root-frequency
  — 4x frequency doubles attenuation in dB
  — fast rolloff once attenuation reaches 20dB

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Transformer Coupling

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

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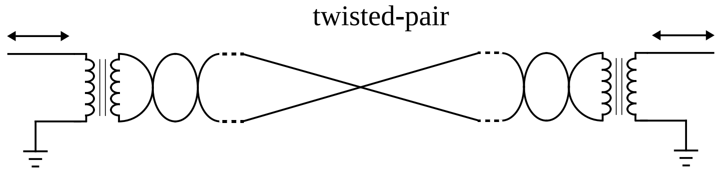
Transformer Coupling

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



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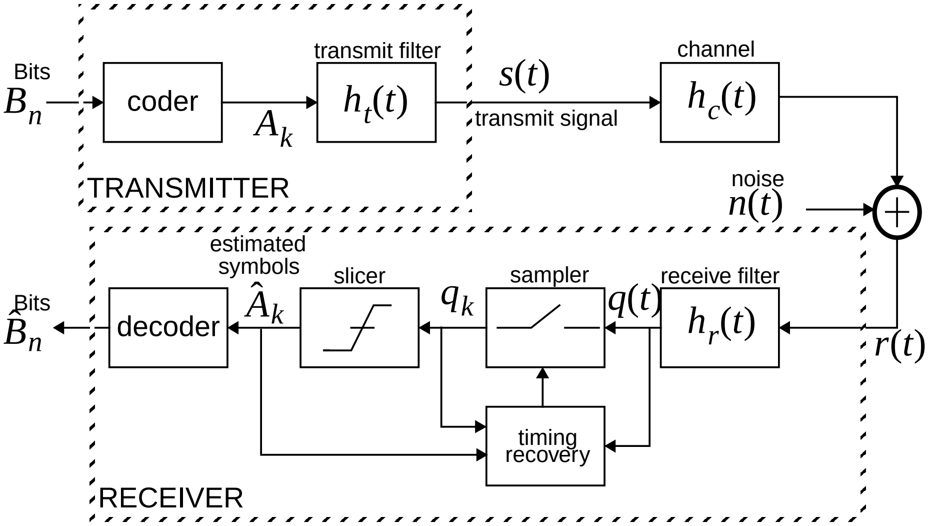
Basic Baseband System



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

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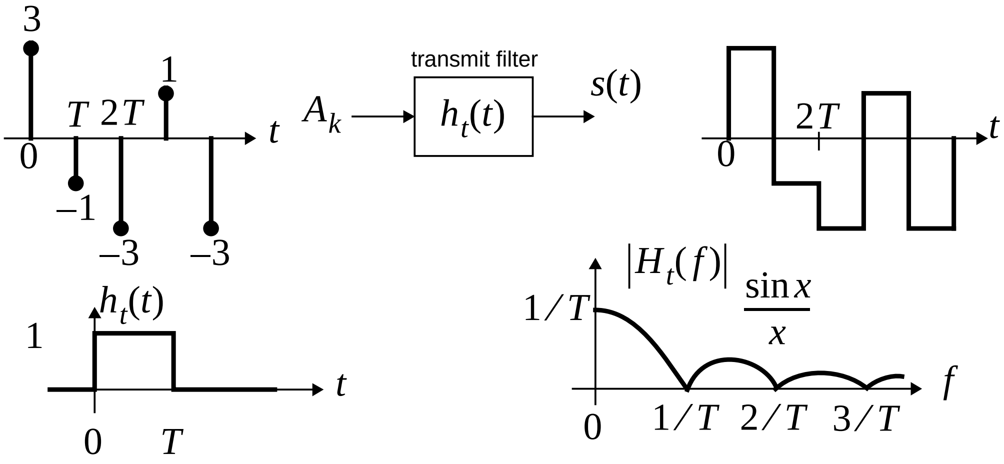
Rectangular Transmit Filter



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

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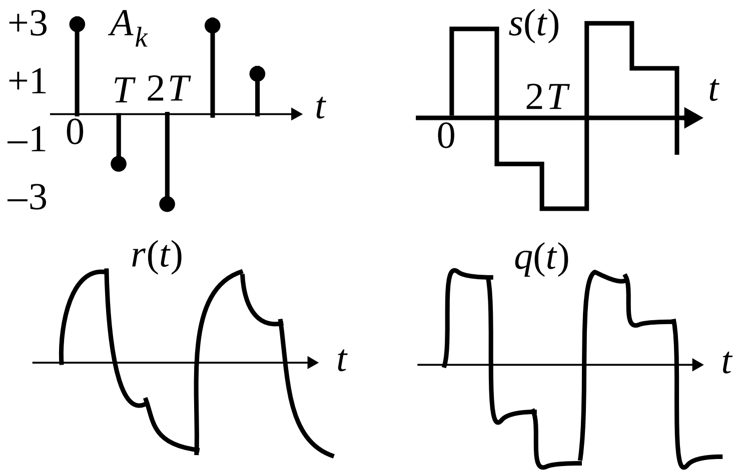
Multi-Level — Low-Noise, Large Bandwidth



4-PAM

2 bits
every
symbol

- Twice the bit information over same bandwidth!
- More susceptible to noise (but perhaps less noise)
- Commonly called PAM (here 2B1Q — 4-PAM)

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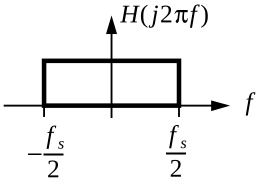
Nyquist Pulses

- For zero intersymbol interference, frequency domain criteria: ($f_s = 1/T$)

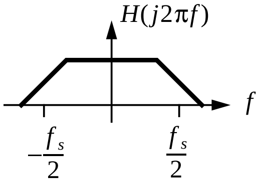
$$\frac{1}{T}\sum_{m=-\infty}^{\infty} H(j2\pi f + jm2\pi f_s) = 1 \tag{2}$$

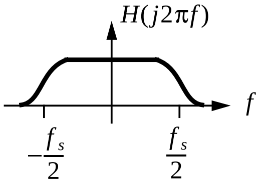
where $H(f) = H_t(f)H_c(f)H_r(f)$

Example Nyquist Pulses (in freq domain)



Sinc pulse





Raised-cosine pulse

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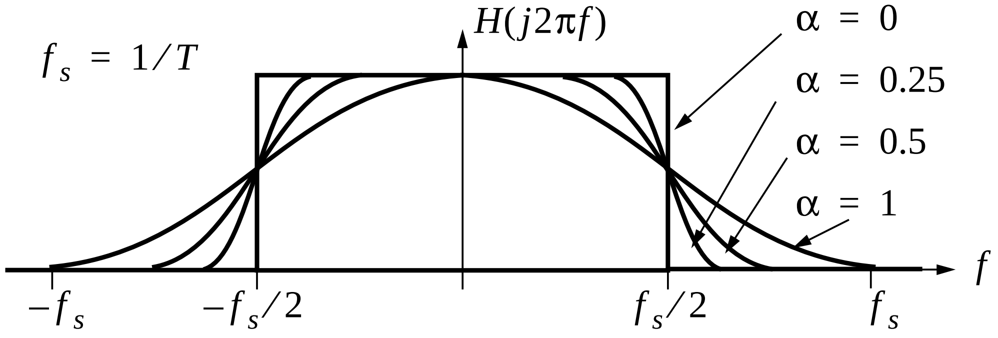
Raised-Cosine Pulse



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

- α determines ***excess bandwidth***

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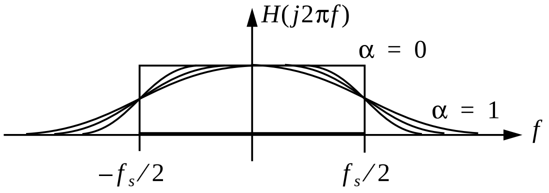
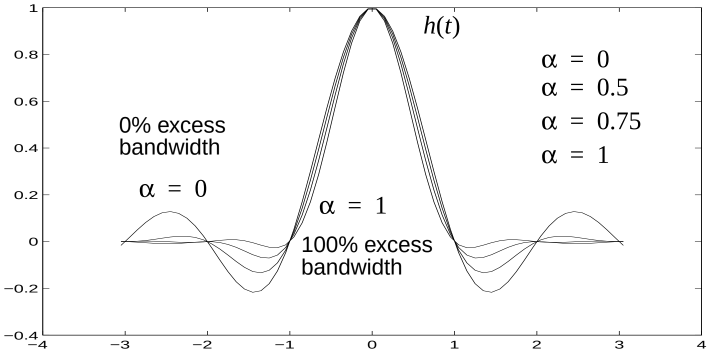
Raised-Cosine Pulses





- More excess bandwidth — impulse decays faster.

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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
- $1.5 \times (f_s/2) = 10\ \text{kHz} \ \Rightarrow \ f_s = 13.333\ \text{ksymbols/s}$

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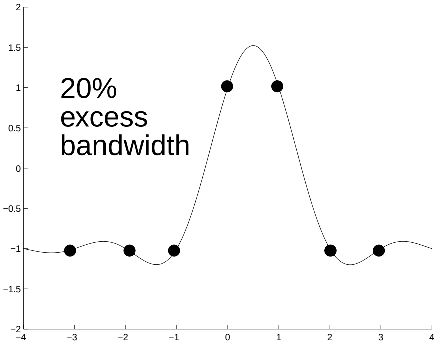
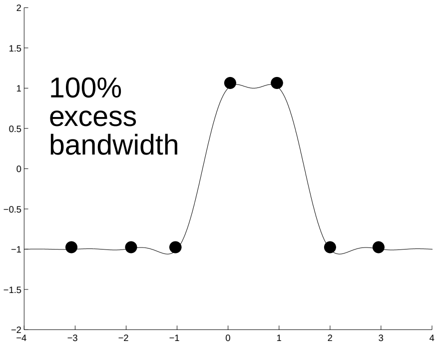
Example Waveforms





- Input: -1 -1 -1 +1 +1 -1 -1
- Crest factor: peak to rms ratio
  — higher crest factor with lower excess bandwidth

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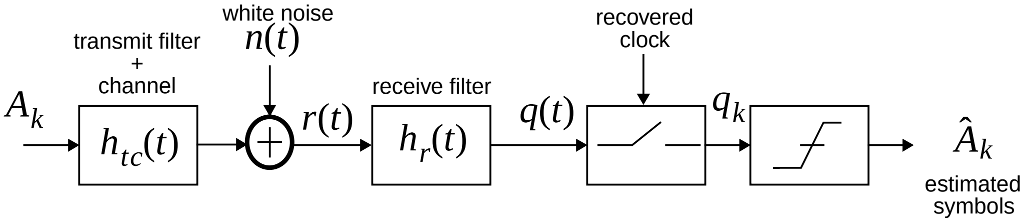
Matched-Filter

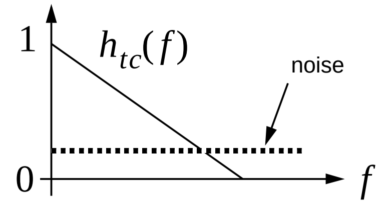

white noise
$n(t)$
transmit filter + channel
recovered clock
receive filter
A_k → $h_{tc}(t)$ → ⊕ → $r(t)$ → $h_r(t)$ → $q(t)$ → q_k → $\hat{A}_k$
estimated symbols

- For zero-ISI, $h_{tc}(t) \otimes h_r(t)$ satisfies Nyquist criterion.
- For optimum noise performance, $h_r(t)$ ***matched-filter***.
- Matched-filter — time-reversed impulse resp $h_{tc}(t)$

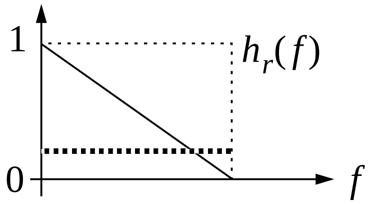
$$h_r(t) = K h_{tc}(-t) \tag{3}$$

where K is arbitrary constant.
- Not usually best for zero-ISI equalization

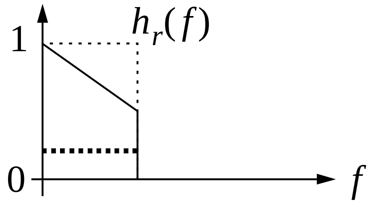
Matched-Filter — Why optimum?



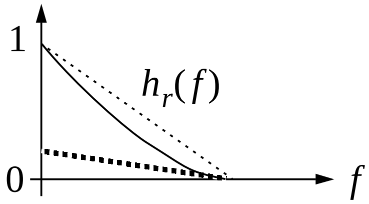
Transmit filter, channel and noise



Too much noise,
All of signal

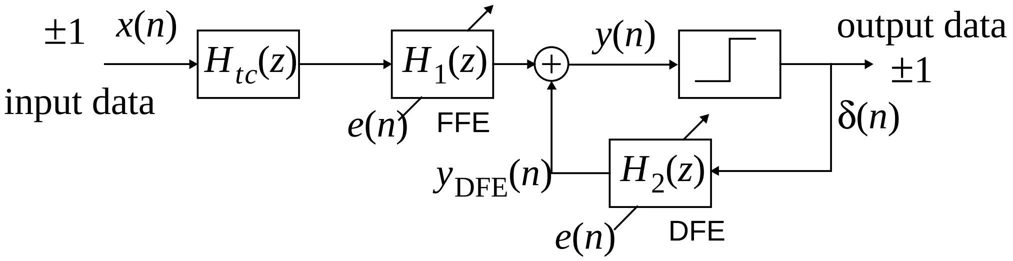


Too little signal,
Less noise



Just right — max SNR

Equalization — FFE and DFE Combined



- Assuming correct operation, output data = input data — otherwise error propagation in DFE
- $e(n)$ can be either:
  - — training: $e(n) = x(n - \text{delay}) - y(n)$
  - — decision directed: $e(n) = \delta(n) - y(n)$
- DFE less complex than FFE (trivial multiplies)

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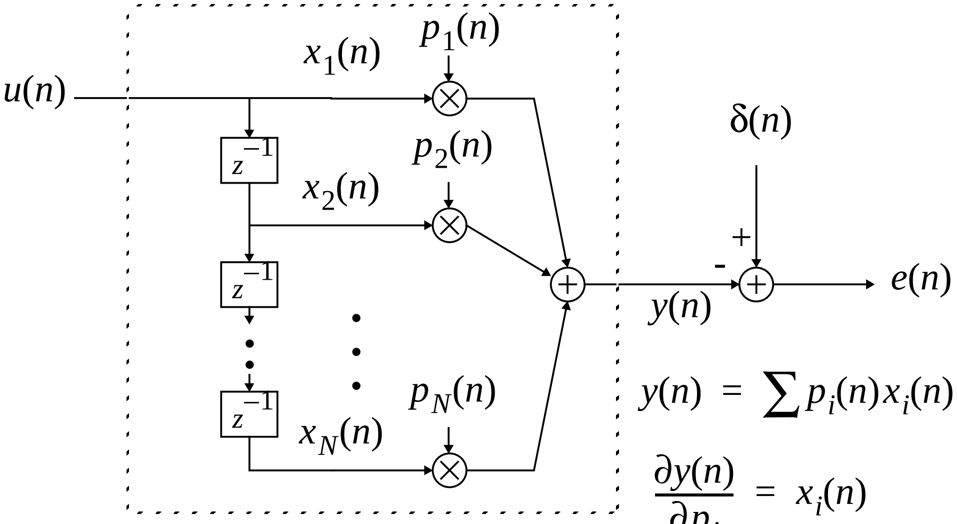
Digital Adaptive Filters

- FIR tapped delay line is the most common



$$y(n) = \sum p_i(n) x_i(n)$$

$$\frac{\partial y(n)}{\partial p_i} = x_i(n)$$

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LMS Algorithm (and variants)

- ***LMS*** — $p_i(n+1) = p_i(n) + 2\mu e(n) \times x_i(n)$

Variants to Reduce Complexity

- ***Sign-data LMS*** — $p_i(n+1) = p_i(n) + 2\mu e(n) \times \mathrm{sgn}(x_i(n))$
- ***Sign-error LMS*** — $p_i(n+1) = p_i(n) + 2\mu\,\mathrm{sgn}(e(n)) \times x_i(n)$
- ***Sign-sign LMS*** — $p_i(n+1) = p_i(n) + 2\mu\,\mathrm{sgn}(e(n)) \times \mathrm{sgn}(x_i(n))$
- However, the sign-data and sign-sign algorithms have gradient misadjustment — ***may not converge!***
- Might take a few bits (rather than just sign)

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Fractionally-Spaced FFE

- Feed forward filter is often a FFE sampled at 2 or 3 times symbol-rate — fractionally-spaced (i.e. sampled at $T/2$ or at $T/3$)

Advantages

— Allows the matched filter to be realized digitally and also adapt for channel variations (not possible in symbol-rate sampling)

— Also allows for simpler timing recovery schemes (FFE can take care of phase recovery)

Disadvantage

More costly to implement — full and higher speed multiplies, also higher speed A/D needed.

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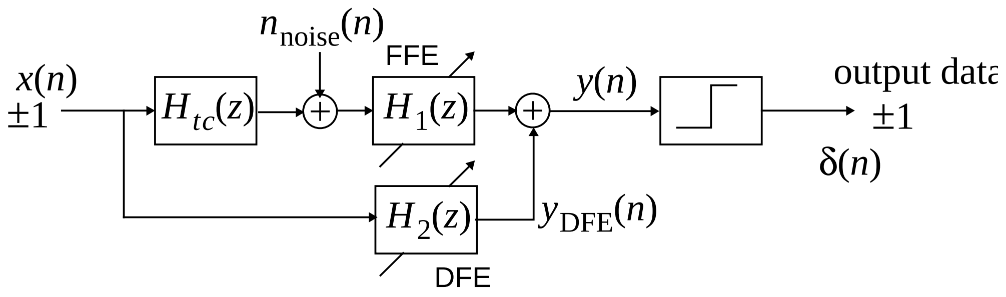
FFE and DFE Combined

Model as:



$$\frac{Y}{N} = H_1 \tag{4}$$

$$\frac{Y}{X} = H_{tc}H_1 + H_2 \tag{5}$$

- When H_{tc} small, make $H_2 = 1$ (rather than $H_1 \to \infty$)

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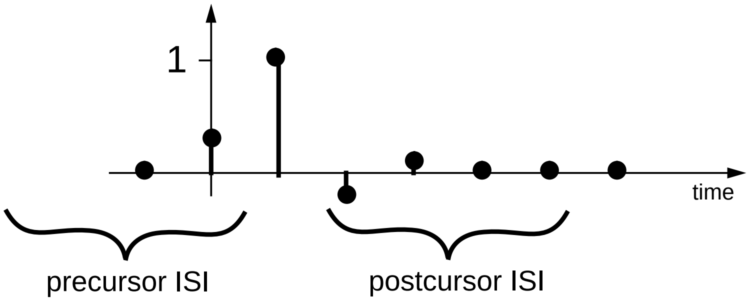
DFE and FFE Combined



- FFE can deal with precursor ISI and postcursor ISI
- DFE can only deal with postcursor ISI (cancellation)
- However, FFE enhances noise while DFE does not

When both adapt

- FFE adds little boost by pushing precursor into postcursor ISI (allpass)

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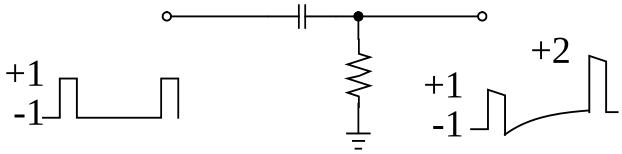
dc Recovery (Baseline W ander)

- Wired channels often ac coupled
- Reduces dynamic range of front-end circuitry and also requires some correction if not accounted for in transmission line-code



- Front end may have to be able to accommodate twice the input range!
- DFE can restore baseline wander - lower frequency pole implies longer DFE
- Can use line codes with no dc content — CAP/QAM, DMT, AMI (but not bandwidth efficient)

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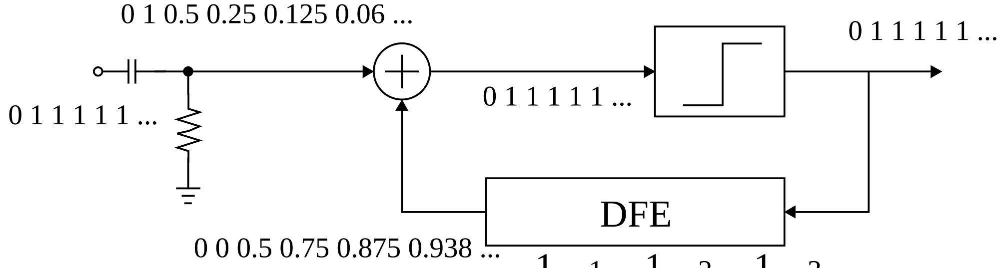
Baseline Wander Correction

DFE Based

$$\frac{z-1}{z-0.5} = 1 - \frac{1}{2}z^{-1} - \frac{1}{4}z^{-2} - \frac{1}{8}z^{-3} - \ldots$$

STEP INPUT



0 1 0.5 0.25 0.125 0.06 ...

0 1 1 1 1 1 ...

0 1 1 1 1 1 ...

0 1 1 1 1 1 ...

DFE

0 0 0.5 0.75 0.875 0.938 ...

$$\frac{1}{2}z^{-1} + \frac{1}{4}z^{-2} + \frac{1}{8}z^{-3} + \ldots$$

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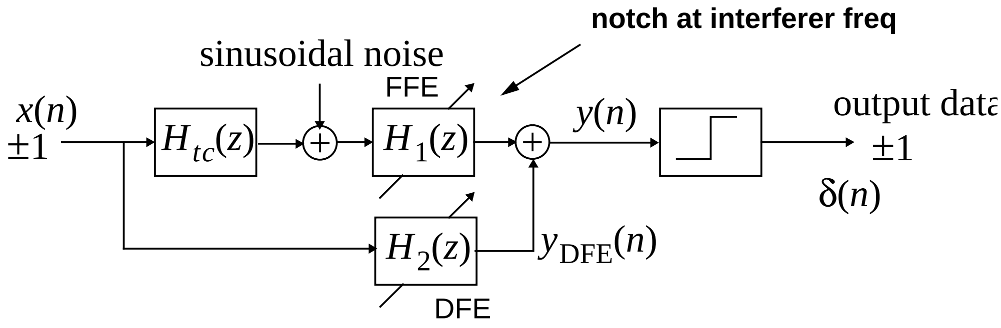
Sinusoidal Interf erence

- A sinusoidal interference can be notched out in FFE
- DFE can fill in missing frequency portion



- Effectiveness depends on FFE and DFE lengths
  — also good SNR so DFE error propagation is small

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Quadrature Amplitude Modulation (QAM)

In General

- Start with two independent real signals, $a(t)$, $b(t)$
  — call one real and one imag (for convenience)

$$u(t) = a(t) + jb(t) \tag{6}$$

- Modulate by $e^{j\omega_c t} = \cos(\omega_c t) + j\sin(\omega_c t)$ and keep real part

$$y(t) = \sqrt{2}\mathrm{Re}\left\{u(t) \times e^{j\omega_c t}\right\}$$
$$y(t) = \sqrt{2}a(t)\cos(\omega_c t) - \sqrt{2}b(t)\sin(\omega_c t) \tag{7}$$

- While QAM and single sideband have same spectrum efficiency, QAM does not need a phase splitter

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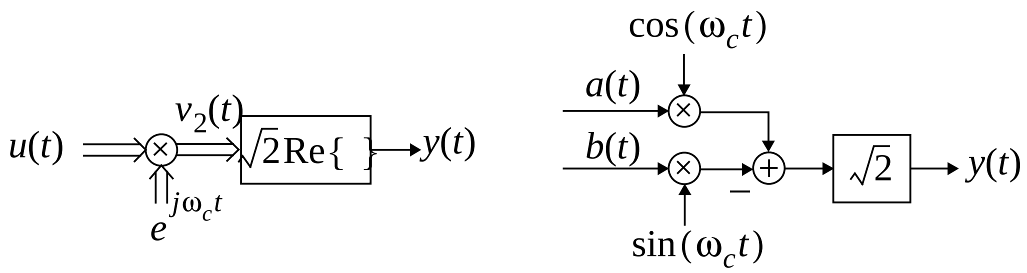
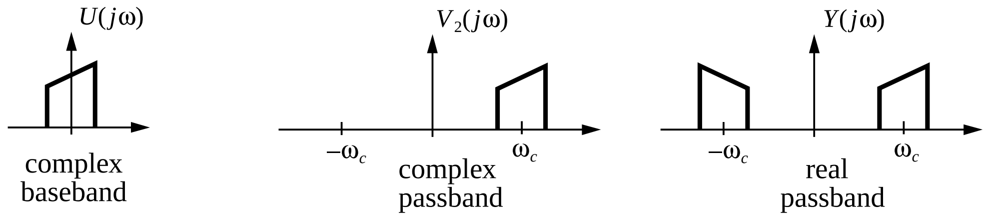
QAM Transmit





- Possibly not symmetrical around carrier frequency

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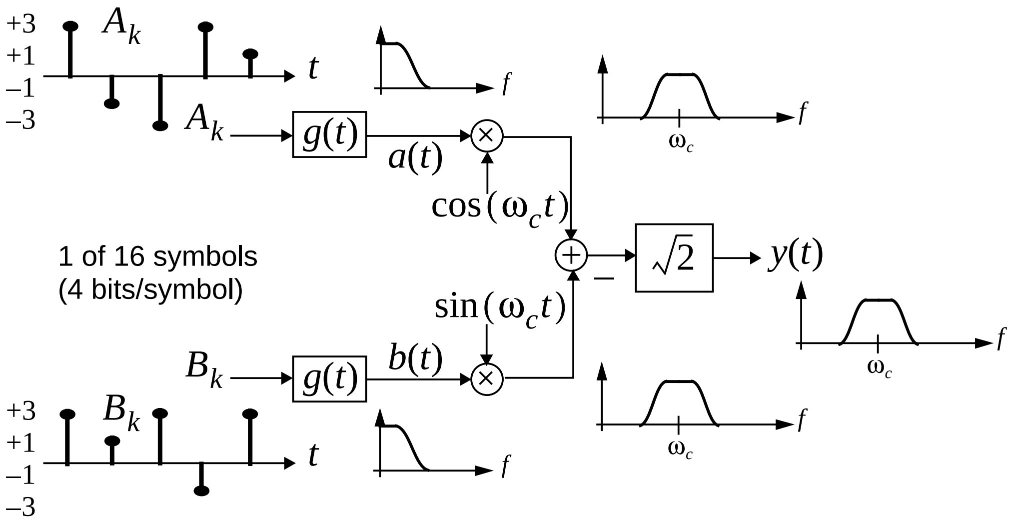
Digital QAM Transmit

- Let $a(t)$ and $b(t)$ be the output of two pulse shaping filters with multilevel inputs, A_k and B_k



1 of 16 symbols
(4 bits/symbol)

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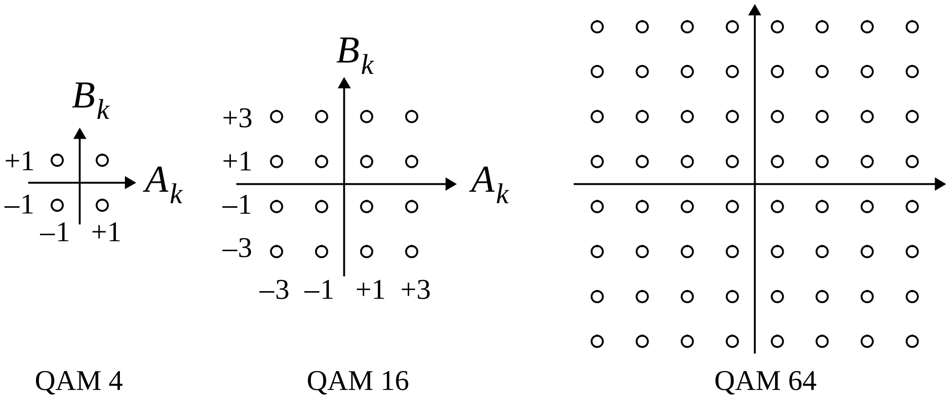
QAM

- PAM each independent data stream
- Signal constellations



- Gray encode so that if closest neighbor to correct symbol chosen, only 1 bit error occurs

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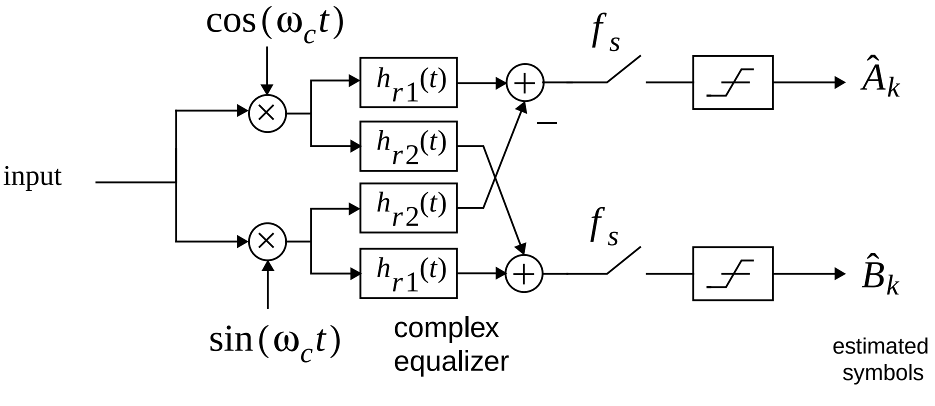
QAM Receiver



- Treat as two independent streams though they are synchronized in time
- Can use FFE, DFE on each stream as in baseband case.
- Timing recovery shared between two streams

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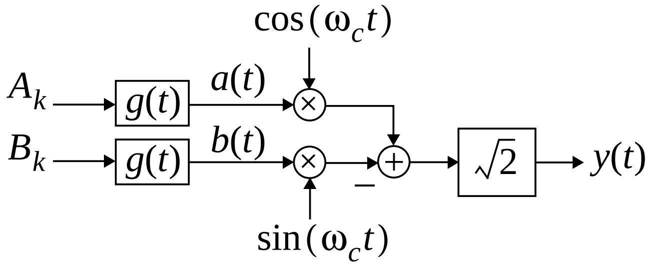
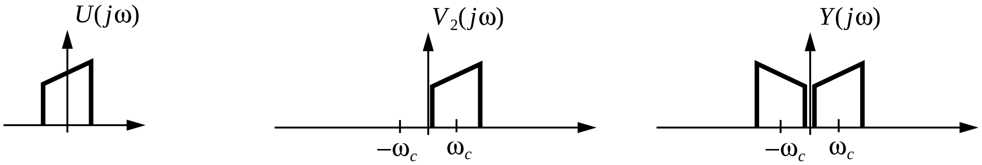
QAM Low Freq Modulation

- Modulate to a low freq f_c just so no dc occurs — or perhaps a bit more





complex baseband

complex passband

real passband

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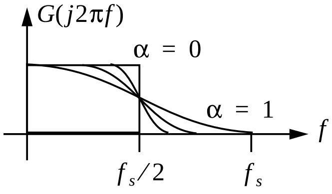
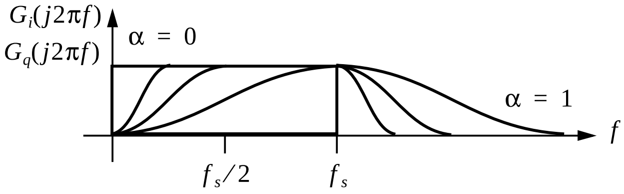
QAM Low Freq Modulation

- The choice for f_c depends on excess bandwidth



lowpass prototype



passband

- Excess bandwidth naturally gives a notch at dc
- For 100% excess bandwidth $f_c = f_s$
- For 20% excess bandwidth $f_c = 1.2 \times f_s/2$

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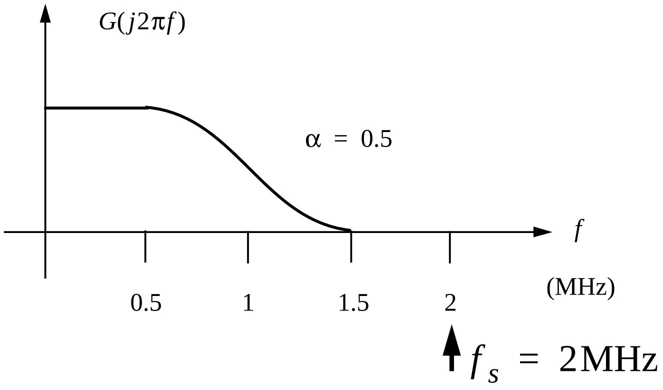
Example — Baseband PAM

- Desired Rate of 4Mb/s — Freq limited to 1.5MHz
- Use 50% excess bandwidth ($\alpha = 0.5$)
- Use 4-level signal (2-bits) and send at 2MS/s



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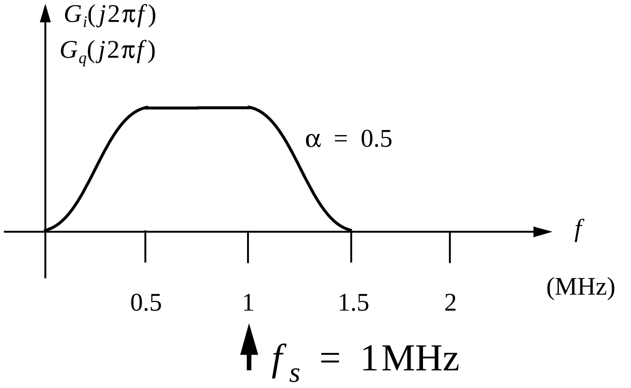
Example — QAM

- Desired Rate of 4Mb/s — Freq limited to 1.5MHz
- Use 50% excess bandwidth ($\alpha = 0.5$)
- Use QAM-16 signalling and send at 1MS/s



- Area under two curves same

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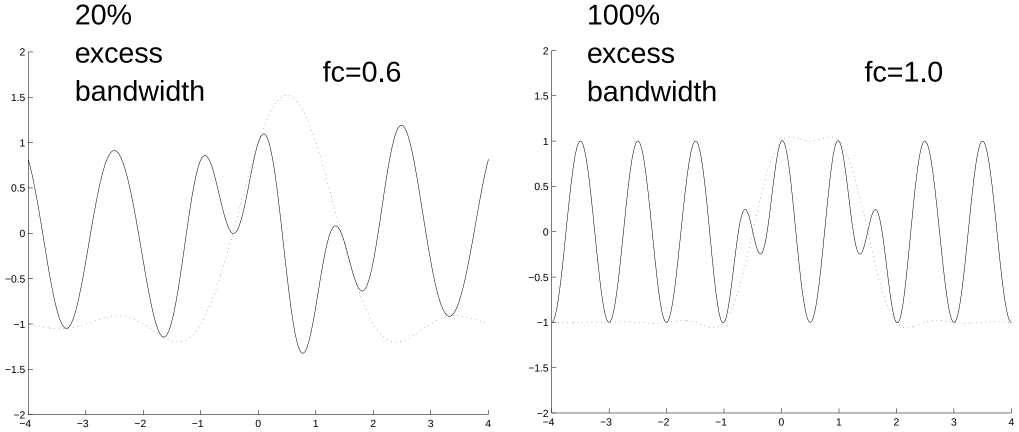
Example QAM Waveforms



- Only “cos” modulated waveform shown
- QAM waveform always within baseband envelope

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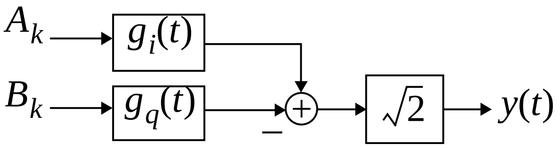
CAP (Carrierless AM/PM)

- Can directly create impulse response of two QAM-like signals.



$$g_i(t) = g(t)\cos(\omega_c t) \tag{8}$$

$$g_q(t) = g(t)\sin(\omega_c t) \tag{9}$$

- Not feasible if ω_c is much greater than symbol freq
- Two impulse responses are orthogonal

$$\int_{-\infty}^{\infty} g_i(t) g_q(t) dt = 0 \tag{10}$$

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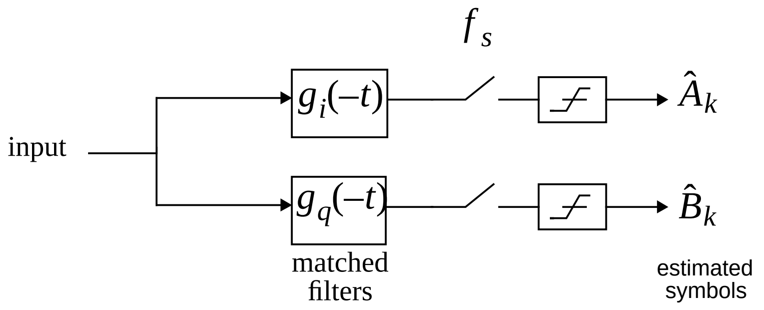
CAP

- Two matched filters used for receiver



- No need for demodulation by cos and sin
- Need to adapt each one to separate impulse — should ensure they do not converge to same impulse

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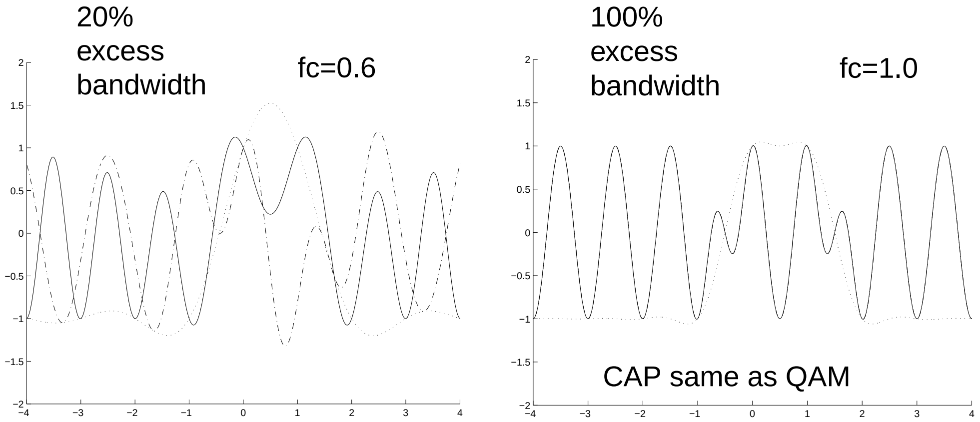
CAP and QAM



- CAP same as QAM if carrier is a multiple of fs
- Not same if non-multiple (rotating QAM signal)
- CAP waveform might not fall within envelope of baseband signal

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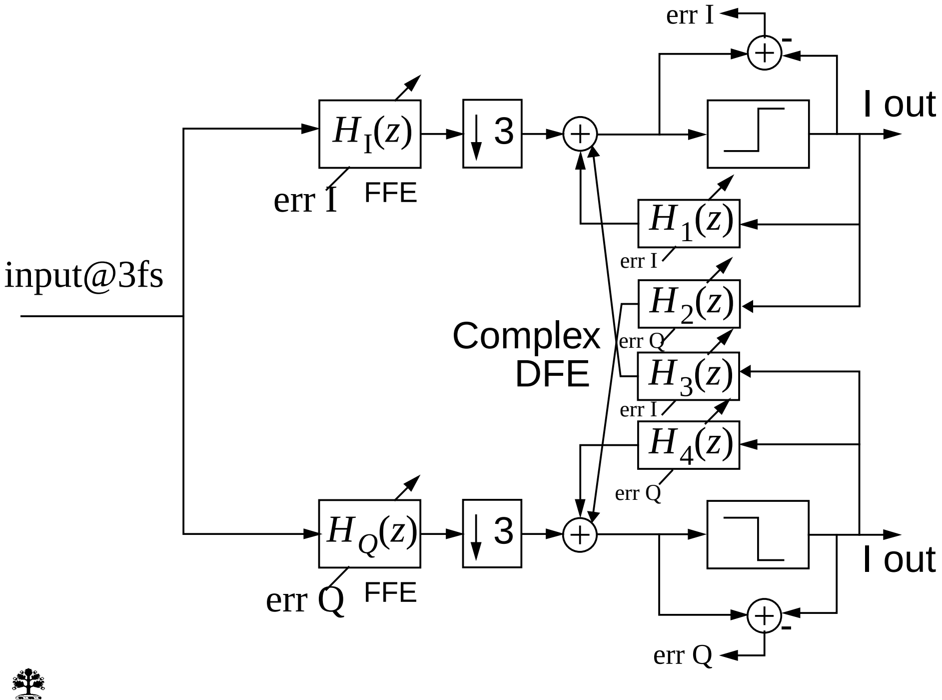
CAP/QAM vs. PAM

- Both have same spectral efficiency

- CAP is a passband scheme and does not rely on signals near dc
- More natural for channels with no dc transmission
- Freedom of modulating signal to desired band

- Can always map a PAM scheme into CAP
  — 2-PAM ↔ 4-CAP   4-PAM ↔ 16-CAP
  8-PAM ↔ 64-CAP
- Cannot always map CAP scheme into PAM
  — cannot map 32-CAP since $\sqrt{32}$ not an integer

CAP Equalization



CAP Equalization

FFE operates at 3Fs

- 3 times to satisfy Nyquist sampling
- matched filtering is adaptive
- phase adjustment possible (timing recovery need only find frequency)

FFE are polyphase filters

- Outputs of FFE are immediately downsampled by 3
- N tap filter requires N multiply/accumulates at ***downsampled rate***

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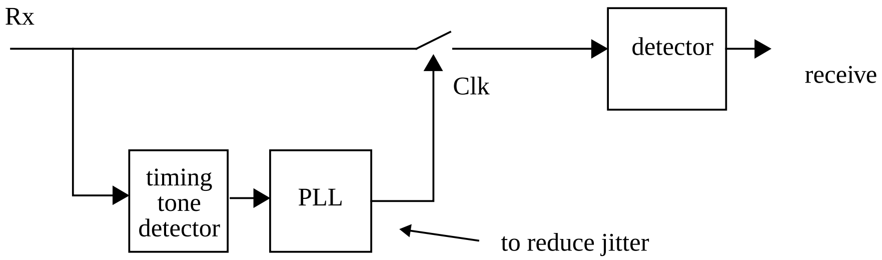
Deductive Timing Recovery



deductive

- Apply non-linearity to generate fs tone.
- Common non-linearity is absolute value
- ***Ensemble average*** of non-linear circuit output is periodic in T (i.e. tone at fs)
- Thus, f_s component exists (with scrambled data) although not present before non-linearity

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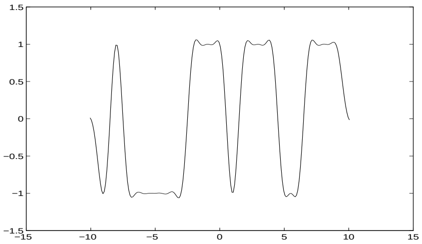
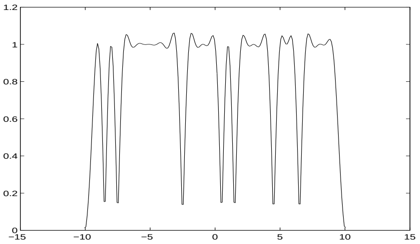
Baseband Example (100% excess BW)

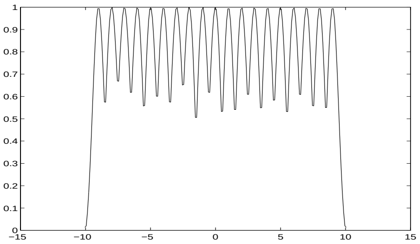


receive signal

average NOT in time but
over transmit sequences
(100 sequences in this case)



abs(receive signal)



average(abs(receive signal))

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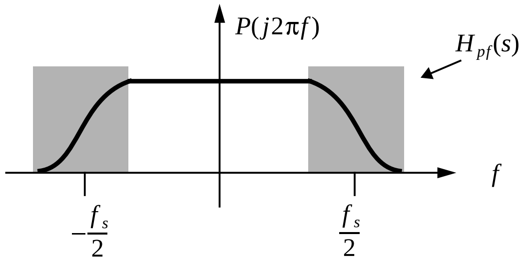
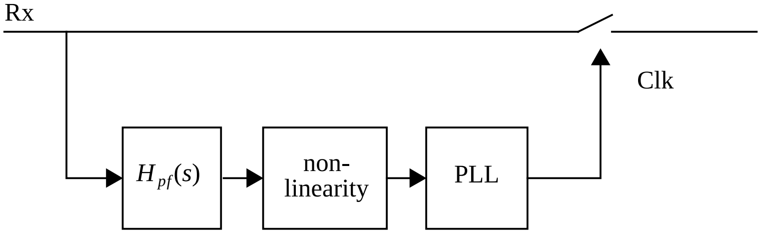
Deductive Timing

- Can pre-filter receive signal to only non-flat portion to reduce jitter — eliminate portion that does not contribute to timing tone.





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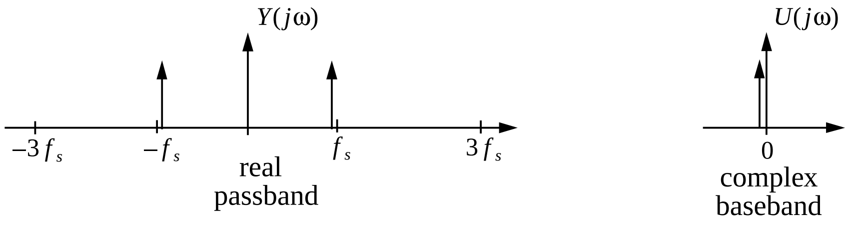
Digital PLL

- Complex modulate signal by fs (down to dc)
  — Mult by sin(fs) and cos(fs) (clock at 3fs)
- Adjust 3fs until frequency is precisely at dc
  — if positive freq, speed clock up
  — if negative freq, slow clock down



- Sinusoid output tells whether speed up or down
- Use a digital controlled oscillator to adjust freq

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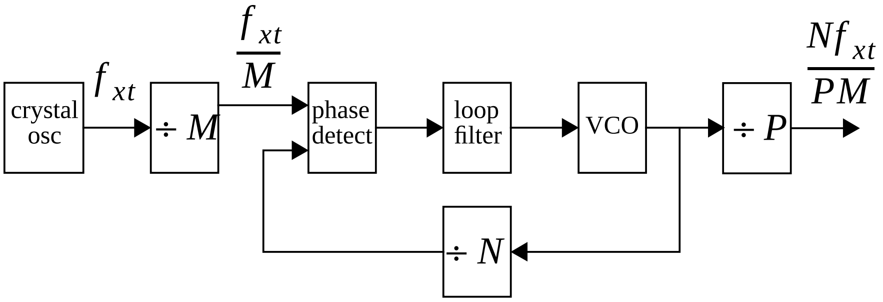
A Fractional-N Frequency Synthesizer

- Often need a low jitter clock that can have arbitrary frequency.
- A voltage-controlled crystal oscillator is expensive.
- Use oversampling within a PLL



$N = \{k, k+1\}$

A digital controlled oscillator

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HDSL and ADSL Applications

HDSL Goal

- Transmit 1.544Mb/s over 5.5km of telephone cables
- Symmetric and full-duplex operation
- Baseband and Passband line codes in use today
- Presently two wire pairs (i.e. 4 wires)

ADSL Goal

- Rate-adaptive
- Downstream transmit — 640kb/s to 7Mb/s
- Upstream transmit — 270kb/s to 1Mb/s
- One wire pair — length depends on rate

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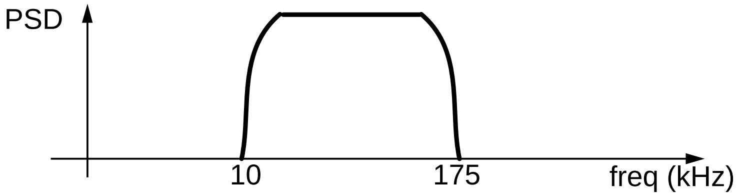
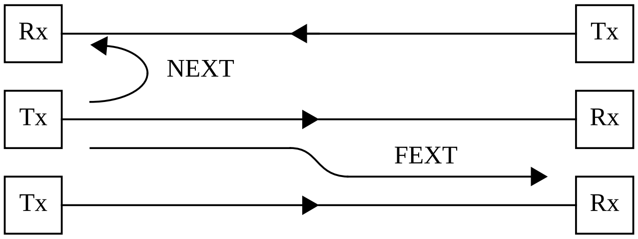
CAP/QAM HDSL



- Downstream and upstream use same freq band
- Requires effective echo cancellation — high linearity is major challenge
- NEXT limits data rate



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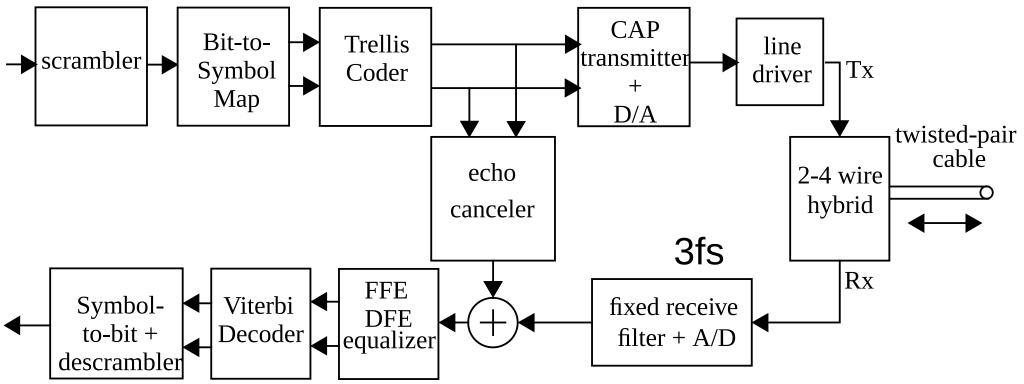
CAP HDSL Transceiver



scrambler

Bit-to-Symbol Map

Trellis Coder

CAP transmitter + D/A

line driver

Tx

echo canceler

2-4 wire hybrid

twisted-pair cable

3fs

Rx

fixed receive filter + A/D

FFE DFE equalizer

Viterbi Decoder

Symbol-to-bit + descrambler

- Some echo cancellation done in hybrid
- Downsample by 3 done after FFE (polyphase filters)

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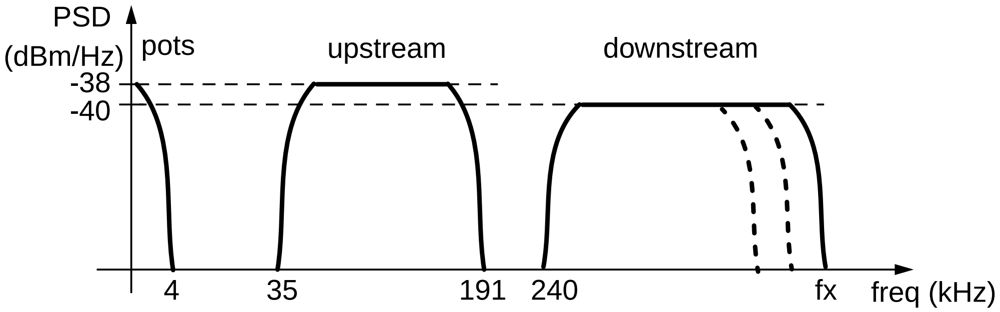
CAP/QAM RADSL



- FDM used for downstream and upstream
- Requires more bandwidth but no NEXT limitation — FEXT limits data rate
- Major challenge is to build high performance bandsplit filters

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CAP/QAM RADSL

Upstream

- Baud rate fixed at 136 kBaud
- Vary bits/symbol to achieve various data rates
- 3 bits/symbol (272kb/s) to 8 bits/symbol (952kb/s)
- Also coding to achieve 4 dB of coding gain

Downstream

- fx varies from 396 to 1491 kHz
  — 136 kBaud ⇒ fx = 396 kHz
  — 340 kBaud ⇒ fx = 631 kHz
  — 680 kBaud ⇒ fx = 1022 kHz
  — 952 kBaud ⇒ fx = 1335 kHz
  — 1088 kBaud ⇒ fx = 1491 kHz
- 3 bits/symbol to 8 bits/symbol (4 dB coding gain)

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RADSL Line Interface Issues

Line Driver

- Transmit launch levels near 20V pp
  (since self next is not limit and higher freq)
- Bipolar line drivers to obtain linearity and drive
  — presently separate chip
- Crest factor around 4 (higher for DMT)

Bandsplit Filters

- Often external RLC filtering for linearity reasons
- Might have some internal integrated filtering

Echo Cancellation

- New systems looking at full-duplex over lower band

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Echo Cancellation

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.

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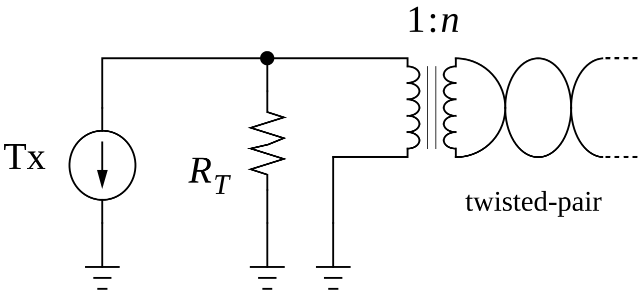
Line Drivers

- Line driver supplies drive current to cable.
- Often current drive in ethernet case



- Not practical for high-linearity (no feedback)
  — large non-linear capacitance affects current out
- Most xDSL line drivers realized as voltage buffers
- High crest factor makes line drivers more challenging

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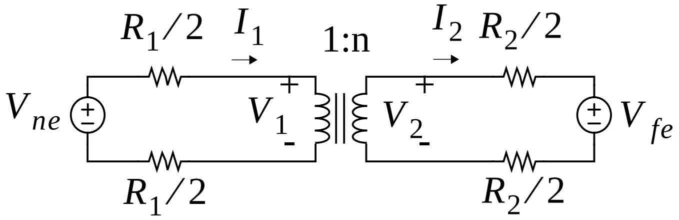
Line Driver

- Can be the most challenging part of analog design.
- Turns ratio of transformer determines equivalent line impedance.



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

$$V_1 = V_2/n$$
$$I_1 = nI_2$$
$$R_1 = R_2/n^2$$

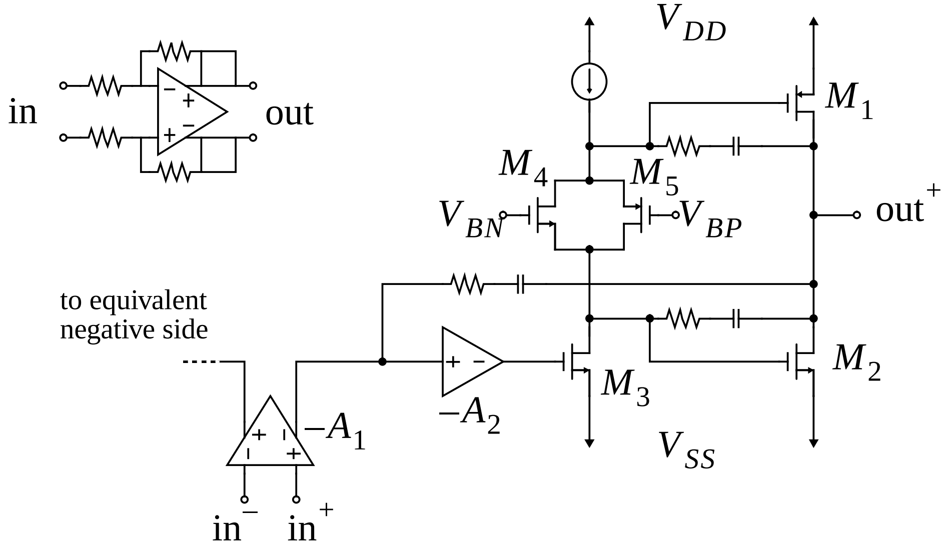
Typical Values

$$R_2 = 100\Omega$$
$$V_2 = \pm 2.5\text{V}$$
$$I_2 = \pm 25\text{mA}$$

Line Driver

- 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
- Efficiency improves as power supplies increase
- Design difficulties will increase as power supplies decreased

Example CMOS Line Driver



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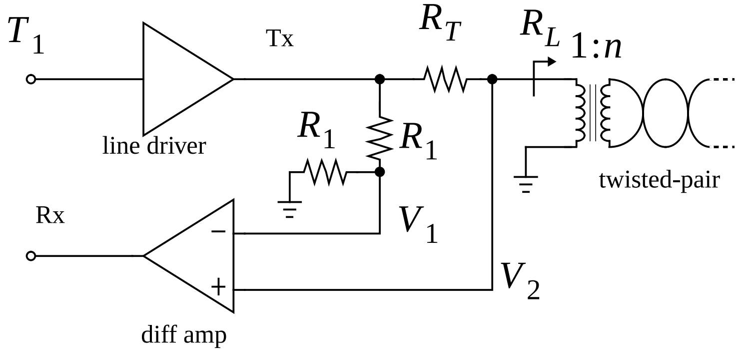
2-4 Wire Hybrid



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

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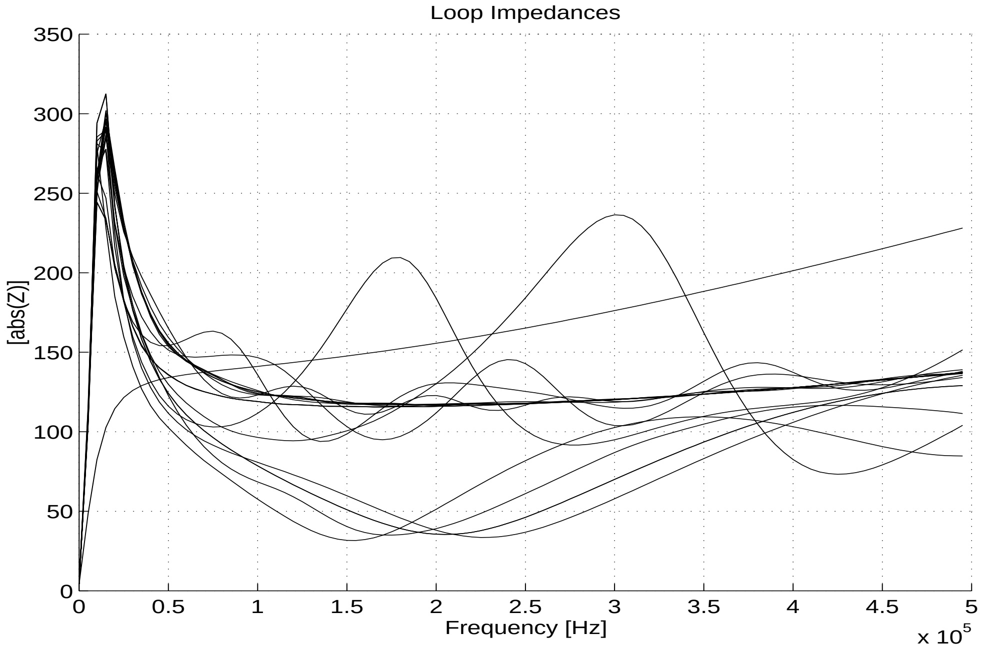
Typical Line Impedances



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Hybrid Issues

- Low frequency pole causes long echo tail in baseband system
  (Baseband 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
  — Extra D/A to relax A/D converter
  — perhaps 2 A/D converters to relax line driver

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