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
Cable Modeling

- Modeled as a transmission line.

\[ L_{dx} \quad R_{dx} \]

\[ G_{dx} \quad C_{dx} \]

Twisted-Pair Typical Parameters:

- \( R(f) = (1 + j)\sqrt{f/4} \ \Omega/km \) 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)
- \( G = 0 \)

Cable Attenuation

- Cable gain in dB is

\[ H_{dB}(d, \omega) = -k_R \times d \times \sqrt{\omega} \]  

(1)

- \( k_R \) — cable constant (typically 0.008)
- \( d \) — cable distance in km
- \( \omega \) — frequency in rad/s

- Attenuation in dB is proportional to cable length
  — 2x distance doubles attenuation in dB
  — reduce attenuation by using larger diameter cable

- Attenuation also proportional to root-frequency
  — 4x frequency doubles attenuation in dB
  — fast rolloff once attenuation reaches 20dB
**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

![Twisted-pair transformer](image-url)
Basic Baseband System

- In 2B1Q, coder maps 2 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 \( H_t(f) \)
- dc component exists
**Multi-Level — Low-Noise, Large Bandwidth**

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

![Multi-Level Waveforms](image)

**Nyquist Pulses**

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

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

where \( H(f) = H_L(f)H_c(f)H_r(f) \)

**Example Nyquist Pulses (in freq domain)**

- **Sinc pulse**
- **Raised-cosine pulse**
Raised-Cosine Pulse

\[ f_s = 1/T \]

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

- \( \alpha \) determines excess bandwidth

Raised-Cosine Pulses

- More excess bandwidth — impulse decays faster.
**Raised-Cosine Pulse**

- $\alpha$ 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
- $1.5 \times (f_s/2) = 10$ kHz $\Rightarrow f_s = 13.333$ 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
• For zero-ISI, \( h_{tc}(t) \otimes h_r(t) \) satisfies Nyquist criterion.
• For optimum noise performance, \( h_r(t) \) \textit{matched-filter}.
• Matched-filter — time-reversed impulse resp \( h_{tc}(t) \)

\[
h_r(t) = K h_{tc}(-t)
\]

where \( K \) is arbitrary constant.
• Not usually best for zero-ISI equalization

\textbf{Matched-Filter — Wh y optimum?}

- Too little signal, Less noise
- Too much noise, All of signal
- 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)

Digital Adaptive Filters

- FIR tapped delay line is the most common
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** — \( p_i(n + 1) = p_i(n) + 2\mu e(n) \times \text{sgn}(x_i(n)) \)
- **Sign-error LMS** — \( p_i(n + 1) = p_i(n) + 2\mu \text{sgn}(e(n)) \times x_i(n) \)
- **Sign-sign LMS** — \( p_i(n + 1) = p_i(n) + 2\mu \text{sgn}(e(n)) \times \text{sgn}(x_i(n)) \)

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.
**FFE and DFE Combined**

Model as:

\[
\begin{align*}
    x(n) \pm 1 & \xrightarrow{H_{tc}(z)} H_1(z) + n_{\text{noise}}(n) \\
    & \xrightarrow{H_2(z)} y(n) \\
    & \xrightarrow{\text{FFE}} y_{\text{DFE}}(n) \\
    & \xrightarrow{\text{output data}} \pm 1 \\
    \delta(n) & = H_1 \\
    Y/N & = H_1 \\
    Y/X & = H_{tc}H_1 + H_2 \\
    & \text{When } H_{tc} \text{ small, make } H_2 = 1 \text{ (rather than } H_1 \to \infty) 
\end{align*}
\]

**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)
dc Recovery (Baseline Wander)

- 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)

Baseline Wander Correction

DFE Based

\[
\frac{z - 1}{z - 0.5} = 1 - \frac{1}{2}z^{-1} - \frac{1}{4}z^{-2} - \frac{1}{8}z^{-3} - \ldots \quad \text{STEP INPUT}
\]

0 1 0.5 0.25 0.125 0.0625 0.03125 0.015625 0.0078125 0.00390625 0.001953125 0.0009765625 0.00048828125 0.000244140625...

DFE

\[
\frac{1}{2}z^{-1} + \frac{1}{4}z^{-2} + \frac{1}{8}z^{-3} + \ldots
\]
**Sinusoidal Interference**

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

![Diagram of sinusoidal interference notched out in FFE and DFE]

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

**Quadrature Amplitude Modulation (QAM)**

**In General**

- Start with two independent real signals, \( a(t), b(t) \) — call one real and one imag (for convenience)
  \[
  u(t) = a(t) + j b(t)
  \]  
  (6)

- Modulate by \( e^{j \omega_c t} = \cos(\omega_c t) + j \sin(\omega_c t) \) and keep real part
  \[
  y(t) = \sqrt{2} \text{Re}\left\{ u(t) \times e^{j \omega_c t} \right\}
  \]
  \[
  y(t) = \sqrt{2} a(t) \cos(\omega_c t) - \sqrt{2} b(t) \sin(\omega_c t)
  \]  
  (7)

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

- Possibly not symmetrical around carrier frequency

Digital QAM Transmit

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

- PAM each independent data stream
- Signal constellations

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

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
**QAM Low Freq Modulation**

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

$$\begin{align*}
  \cos(\omega_c t) \\
  A_k \rightarrow g(t) \rightarrow a(t) \rightarrow y(t) \\
  B_k \rightarrow g(t) \rightarrow b(t) \rightarrow \sqrt{2} \rightarrow \sin(\omega_c t)
\end{align*}$$

- The choice for $f_c$ depends on excess bandwidth

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

\[ G(j2\pi f) \]
\[ \alpha = 0.5 \]
\[ f_s = 2\text{MHz} \]

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

\[ G(j2\pi f) \]
\[ G_e(j2\pi f) \]
\[ \alpha = 0.5 \]
\[ f_s = 1\text{MHz} \]

- Area under two curves same
Example QAM Waveforms

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

CAP (Carrierless AM/PM)

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

\[
\begin{align*}
A_k &\rightarrow g_i(t) \\
B_k &\rightarrow g_q(t) \quad \boxplus \quad \sqrt{2} \quad y(t)
\end{align*}
\]

\[
g_i(t) = g(t) \cos(\omega_c t) \quad (8)
\]

\[
g_q(t) = g(t) \sin(\omega_c t) \quad (9)
\]

- Not feasible if \( \omega_c \) is much greater than symbol freq
- Two impulse responses are orthogonal

\[
\int_{-\infty}^{\infty} g_i(t) g_q(t) dt = 0 \quad (10)
\]
**CAP**

- Two matched filters used for receiver

\[
\begin{align*}
&\text{input} \\
&g_f(-t) \quad \mathcal{F} \rightarrow \hat{A}_k \\
g_q(-t) \quad \mathcal{F} \rightarrow \hat{B}_k
\end{align*}
\]

- 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

**CAP and QAM**

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

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**CAP Equalization**

![Diagram of CAP Equalization with mathematical expressions and filter blocks](image-url)
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**

- 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
Baseband Example (100% excess BW)

- Receive signal
- Absolute value of receive signal
- Average of absolute value of receive signal

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

Deductive Timing

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

\[ H_{pf}(s) \]

PLL

Rx

\[ H_{pf}(s) \]

non-linearity

PLL

Clk
**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

\[
\frac{f_{xt}}{M} \quad \leadsto \quad \frac{f_{xt}}{PM} \quad \div \quad \frac{Nf_{xt}}{P}
\]

\[N = \{k, k+1\}\]

A digital controlled oscillator
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

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

![PSD Diagram](slide 48 of 62)
**CAP HDSL Transceiver**

- Some echo cancellation done in hybrid
- Downsampling 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
**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
Echo Cancellation

Received Signal

- For \( d = 4\, km \), a 200kHz signal is attenuated by 40\( dB \).
- 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 6\( dB \)
- To maintain 40\( dB \) SNR receive signal, linearity and noise of transmit path should be better than 74\( dB \).

Line Drivers

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

![Diagram of line driver](image)

- 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
Line Driver

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

\[ V_{ne} = \frac{2}{n} V_2 \]
\[ V_1 = \frac{V_2}{n} \]
\[ I_1 = nI_2 \]
\[ R_1 = \frac{R_2}{n^2} \]

Typical Values

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_2 )</td>
<td>100Ω</td>
</tr>
<tr>
<td>( V_2 )</td>
<td>±2.5V</td>
</tr>
<tr>
<td>( I_2 )</td>
<td>±25mA</td>
</tr>
</tbody>
</table>

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

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

2-4 Wire Hybrid
Typical Line Impedances

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
References

General

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CAP/QAM Info

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CAP/QAM Comparisons

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• G.H. Im et al, “51.84 Mb/s 16-CAP ATM LAN standard,” *IEEE Journal on Selected Areas in Communications*, vol. 13, pp. 620-632, May. 1995

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