• Notes Set 23: Fiber optic communications systems
  
  • Receiver design in colored noise. Whitening filter. Intractability of problem
  • sketch of decision-feedback strategy.
  • Simplified practical (nonoptimum) receivers. Eye diagrams. Personik noise integrals. Optical receiver sensitivity
  • design of low-noise optical receiver front ends
Fiber optic digital communications:

- How it differs from prior material

- Smith & Pershing's method

- Performance Estimates.
We need to return to our prior material: optimax receiver strategy:

1) Message sent by message vector $\text{sequence}$ represented by basis functions $\phi_i$

2) Message now easily detected against white Gaussian noise.

3) Project signal against basis vectors, then project signal vector against set of possible signals to see which is closest.
What happens if noise is not white?

- All narrowband digital microwave receivers

\[ F = \text{noise figure} = F_0 = \text{constant across band} \]

\[ \Rightarrow \text{white system noise} \]

- Very broadband digital microwave receivers:

Fano's limit: cost meth for noise across whole bandwidth

\[ F = F_0 + K (f-f_0)^2 \]

- Noise is no longer white.

- Baseband PCM fiber optics

\[ \text{input noise current} \} = K_1 + K_2 f^2 \]

\[ \text{spectral density} \]  

\[ \Rightarrow \text{not white} \]
What does the theory tell us to do

- Pass the signal + noise through filter such that output noise spectrum is white "whitening filter"

- Theory of previous lecture

- In theory

... gives us the correct receiver architecture.
Problem:

\* After whitening filter, the message vectors are bit-by-bit no longer orthogonal.

\* Theory acknowledge no problem: architecture given previously is "optimum".

\* The problem: the "optimum" receiver structure is indeed optimum in terms of receiver sensitivity, but is highly non-optimum in terms of cost and difficulty of implementation.
Example to clarify this:

Fiber-optic receiver:

* input referred noise power spectral density

\[ N_0 = \frac{a}{f} + kT \]

* Signalling format: on-off keying.

* Basis function

The fiber people call this non-return to zero (NRZ)
Better name: 00kX with square symbol.
Note that the bits are orthogonal.

... so that, although we could in principle look at $n$ bits as $n$ boundaries in an $n$-dimensional signal space, we can in fact look at one bit at a time.
But the noise has P.S.D. like so:

\[ \log(N) \]

\[ f^2 - w = c + 6f^2 \]

\[ = g(1 + 4(\frac{f}{f_0})^2) \]

\[ f_0 = \sqrt{\frac{c}{\alpha}} \]

So an optimal receiver must pass the signal + noise first through a whitening filter:

[Diagram]

\[ \log(f) \]

in \( a + b \log^2 \) noise

White noise
TDL & GBBSec system: $T_0 = 25 \text{ ps}$

Bipolar receiver: $T_0 \sim 16\text{ ps}$, $(T_0 = 160 \text{ ps})$

**Input Pulse to Filter:**

**Impulse response of filter:**

\[ h(t) \approx \frac{1}{k} \left(1 - e^{-t/160\text{ ps}}\right) \text{ volt} \]

**Output Pulse from Filter:**

Decaying: $T = 160 \text{ ps}$
Problem:

6.t n:

6.t (n+1)

etc...

- data bits no longer orthogonal.

- Our theory did not demand orthogonal signals to give an optimum receiver. But it did demand orthogonal bits if the optimum receiver were to be simple.
Receiver structure "optimum"

Inputs are:

The output of matched filter is therefore:

Convolved with...
Response to single bit is \( \alpha \) approximately:

\[ \alpha \]

Next bit

\[ \alpha \]

etc.

Very strong intersymbol interference.
again, inter-symbol interference is ok:

\[ s_3 = s_1 + s_2 \]
\[ 0 \quad s_1 = s_1 + s_2 \]

\[ s_4 = s_1 - s_2 \]
\[ 0 \quad s_2 = s_1 - s_2 \]

no inter-symbol interference allows \( s_1 \) to decade on bit by bit basis (\( s_1, s_2, s_3, \ldots \))

these pictures pertain to the non-orthogonal signal case.
... but with 25 ps-duration bits spread in time to ±160 ps, the decision algorithm must consider \( 2 \times (160/25) = 13 \) successive outputs of the correlator / matched filter in order to determine a given bit's value.

Ideal hardware becomes impractical.

Sacrifice some performance for some reduction in complexity: decision feedback method.

- or -

Sacrifice substantial performance for extreme hardware simplification: Nyquist-like method.
I will not discuss decision feedback here.

* Nyquist method:
  * Extremely simple hardware.
  * Replica whitening & matched filter correlator with simple noise-bandlimiting filter
  * Subject to constraints: zero intersymbol interference. E.g. orthogonal bit slots
Simplified receiver

\[ S_{\text{in}} \rightarrow \text{channel filter} \rightarrow \text{data} \rightarrow \text{level discriminator} \rightarrow \text{data} \]

\[ (1-0-0) \]

Channel  \( f \): No constraint.

\[ \text{Vout}_{nT_\text{bit}} \text{ depends only on value of } n^{th} \text{ bit} \]

- Zero ISI - intersymbol interference.

- Try to minimize noise...

At order of \{Sample\} and discriminator

Can & often is removed becomes comparator & clocked latch.
channel filter

[Drawing of an electronic circuit with labels and symbols]

bandlimited

dcok 0.1 Hz
Eye diagram:

Born on an oscilloscope:
- Scope triggered on clock
- Scope displays data stream
- All possible trajectories superimposed

Input Eye

Output Eye (a good one)

decision circuit looks at voltage at these points in time...
This demands that a bit input:

\[ T_s \]

...which is zero at the center of all other bits.

Finding filters that do this is a general problem solved by Nyquist; they are called Nyquist filters, although not pertaining to sampling.
Smith & Perzanik considered a far more restrictive set of (idealized, not real) filters.

\[
\frac{\sin \left( \frac{\pi T_b}{T_b} \right) \cos \left( \frac{\pi t}{T_b} \right)}{\left( \frac{\pi t}{T_b} \right) \left( 1 - \left( \frac{2 \epsilon}{T_b} \right)^2 \right)}
\]
How they do the analysis...

\[ I_{ph} \rightarrow \text{Front End} \rightarrow \text{Linear Channel} \rightarrow V_o \]

\[ I_{ph} = P_{optical} \cdot B \cdot \frac{n}{\hbar \nu} \]

\[ V_o = K \cdot Z_f(\nu) \cdot I_{ph} \]

Pick \( K \) so that \( Z_f(\nu) \to 1 \) at DC.
we analyze input network to find total input-referred noise current:

\[
\frac{d <i_i^*i_i>}{dt} = a + b f^2
\]

more on this in a minute...

the output noise voltage has variance

\[
\sigma_v^2 = k \int_{-\infty}^{\infty} \eta^2(\omega) \cdot (a + b f^2) \cdot df
\]

the output signal has magnitude:

\[
V_{oc} = I_o i_i \cdot K E_t (\omega=0)
\]
This has the key assumption of a Nyquist zero ISI channel filter.

... so that the amplitude of an isolated bit is indeed \( I = kE_0 \)
Then \( v_m = k \sqrt{\int_{0}^{\infty} \frac{1}{1 + (k + b z^2)^2} \, dz} \)

and \( V_m = k I_m Z(0) \)

\[ \text{OK} \]

\[ \text{boundary } k I_m Z(0)/2 \]

Write \( V_m = k X \) for a moment.
\[ P(\text{Error}) = Q \left[ \frac{\left( k I \cdot z(0) / 2 \right)}{k \chi} \right] \]

\[ = Q \left[ \frac{I \cdot z(0)}{2 \chi} \right] \]

As usual, \( P(\text{Error}) = \text{e.g.} = 10^{-9} \)

\[ \rightarrow I \cdot z(0) / 2 \chi = "Q" = 6 \]

\[ I : n = \frac{2 \chi \cdot Q}{\overline{Z}(0)} = 2Q \sqrt{\int_{0}^{\infty} \left| \frac{Z_{r}(f)}{Z_{r}(0)} \right|^{2} (c + 6f^{2}) \, df} \]

where we have assumed \( \Delta_{z}(f) = a + 6f^{2} \)

so and in general:

\[ I : n = 2Q \sqrt{\int_{0}^{\infty} \left| \frac{Z_{r}(f)}{Z_{r}(0)} \right|^{2} \Delta_{z}(f) \, df} \]
Receiver Sensitivity:

1 Sent \( P_{\text{tot}} = P_1 \implies I_{\text{in}} = \frac{hP_1}{\bar{v}} \)

0 Sent \( P_{\text{tot}} = 0 \implies I_{\text{in}} = 0 \)

\[ \bar{P} = \text{average power} = P_{1/2} \]

\[ \bar{P} = \frac{I_{\text{in}}}{h} \cdot \frac{1}{\bar{v}} \]
\[ \bar{P}_{\text{mi}} = \frac{\hbar \nu}{\mathcal{B}} Q \sqrt{\int_0^\infty \frac{z_r(f)}{z_r(0)} \left| S_z(f) \right|^2 \, df} \]

Now roughly: \[ S_z(f) = a + b f^2 \]

\[ z_r^2 \Delta = \int_0^\infty \frac{z_r(f)}{z_r(0)} \left| S_z(f) \right|^2 \, df \]

\[ = \int_0^\infty \left| \frac{z_r(f)}{z_r(0)} \right|^2 \cdot a \cdot df \]

\[ + \int_0^\infty \left| \frac{z_r(f)}{z_r(0)} \right|^2 \cdot b f^2 \, df \]
Let's normalize with respect to

Bit rate: $B$ bits/second.

$B$ bits/second $\rightarrow$ filter $Z_1(t)$

$2B$ bits/second $\rightarrow$ filter $Z_1(t/2)$

If we want to work with normalized filter frequency responses, e.g., normalized to the bit rate:

$Z_1(t/13) = Z_1(t)$
\[ <I^2> = \int_{0}^{\infty} \left\| \frac{\tilde{Z}_T(f,\beta)}{\tilde{Z}_T(0)} \right\|^2 \cdot a \cdot df \]

\[ + \int_{0}^{\infty} \left\| \frac{\tilde{Z}_T(f,\beta)}{\tilde{Z}_T(0)} \right\|^2 \cdot b \cdot \tilde{f} \cdot df \]

Write \( f = \beta \tilde{f} \)

\[ = aB \int_{0}^{\infty} \left\| \frac{\tilde{Z}_T(f,\tilde{f})}{\tilde{Z}_T(0)} \right\|^2 \cdot \tilde{f} \cdot df \]

\[ + 6B^3 \int_{0}^{\infty} \left\| \frac{\tilde{Z}_T(f,\tilde{f})}{\tilde{Z}_T(0)} \right\|^2 \cdot \tilde{f} \cdot \tilde{f} \cdot \tilde{f} \cdot df \]
\[ I_2 = \int_0^\infty \left\| \frac{\tilde{Z}_x(f)}{\tilde{\tau}_x(0)} \right\|^2 df \]

\[ I_3 = \int_0^\infty \left\{ \left\| \frac{\tilde{Z}_x(f)}{\tilde{\tau}_x(0)} \right\|^2 + f^2 \right\} df \]

are called Pearson's Coefficients.

They are the integrals of the filter's transfer functions in normalized frequency (normalized to the bit rate) and therefore depend only on the shape of the filtered pulse.
\[ P_{\text{min}} = \frac{h \nu}{8} Q \cdot \frac{1}{r^2} \sqrt{\langle i^2 \rangle} \]

where

\[ \langle i^2 \rangle = B \cdot I_2 \cdot a + 13^3 I_3 \cdot 6 \]

\[ S_i(t) = a + 6b^2 \]

\[ I_2 \simeq 0.69 \] \text{ parameters for a}
\[ I_3 \simeq 0.12 \] \text{ "real world" 5-pole L-C filter which approximates a raised cosine.}

\text{Raised cosine: } I_2 \text{ about the same}
\[ I_3 \simeq 0.06 \]
Optical Receiver Basics

Photodetector converts light into current, drives preamplifier

Preamplifier & photodetector set noise performance, hence attainable sensitivity

Nyquist filter bandlimits noise to $\approx B/2$.

Decision circuit recovers binary data
# Key Component Models

## Field-Effect Transistors

\[ \frac{d}{df} \langle I_n I_n^* \rangle = 4kT g_m \]

Other FET parasitics at kT

1/f noise not relevant at rates above 10 GB/s

## Resistors

\[ \frac{d}{df} \langle I_n I_n^* \rangle = 4kT / R \]

## Photodiode

\[ I = \frac{n_g}{i_n} P_{\text{optical}} \]

\[ C_p \]
receiver noise analysis

Approximate input referred noise

\[
\frac{d\langle I_n I_n^* \rangle}{df} = \frac{4kT}{R_L} + \frac{4kT\Gamma}{g_m} (2\pi f)^2 (C_{gs} + C_p)^2
\]

but \( C_{gs} = \frac{g_m}{2\pi f_\tau} \), so choose \( C_{gs} = C_p \)

hence:

\[
\frac{d\langle I_n I_n^* \rangle}{df} = 32\pi kT f^2 C_p / f_\tau + 4kT / R_f
\]
Front-End Noise and Bandwidth

- Broadband circuit: cannot noise-match, noise does not approach $f_{\text{min}}$.

- Fundamental noise source is FET channel noise. Resistive loading added for bandwidth, degrades noise. Note: post-equalization often used.

$$
\frac{d\langle I_n^* I_n \rangle}{df} \approx 32\pi kT f^2 C_p / f_\tau + 4kT / R_f
$$

This assumes intelligent choice of FET size.

- Bandwidth is

$$
f_{3dB} \approx \frac{1 + A}{R_f (C_{\text{diode}} + C_{gs})}
$$

(set $A=1$ for simple stage)
### Common-Gate / Transimpedance Front-Ends

<table>
<thead>
<tr>
<th>same noise voltage and current in CS and CG modes.</th>
<th>transimpedance stage: voltage gain $A_V$</th>
</tr>
</thead>
<tbody>
<tr>
<td>CG stage gives input pole at $f_\tau / 2$.</td>
<td>common-gate stage</td>
</tr>
<tr>
<td>need high $f_\tau$, small $C_{\text{photodiode}}$.</td>
<td></td>
</tr>
<tr>
<td>input noise, given $C_{gs} = C_{\text{photodiode}}$:</td>
<td>$\frac{d\langle I_n I_n^* \rangle}{df} \approx 32 \pi kT f^2 C_p / f_\tau + 4kT / R_f$</td>
</tr>
<tr>
<td>bandwidth:</td>
<td>$f_{3dB} \approx \frac{1 + A}{R_f (C_{\text{diode}} + C_{gs})}$</td>
</tr>
</tbody>
</table>
Optical Receiver Sensitivity: Design Target

-20 dBm @ 40 Gbit/sec for BER~10^9
Key observations regarding noise:

Maximize HEMT $f_t$

Minimize detector and layout capacitance

Monolithic integration **NOT** imperative
  (well-designed bond pads add $\approx 25$ fF capacitance)

50$\Omega$ load would incur large noise penalty
Good transistors will make more sensitive receivers

\[ P \propto \sqrt{\frac{\Gamma C_P B^3}{f_\tau}} \]

-19 dBm @ 100 Gbit/s

Sensitivity can be greatly improved by using good low-noise HEMTs
Good hybrid integration: 25 fF pad capacitance, 2-3 dB penalty.
Calculating Sensitivity:

Channel filter shape with zero intersymbol interference (Nyquist filter) is assumed.

Input-referred noise current has a power spectral density of the form

\[
\frac{d\langle I_n I_n^* \rangle}{df} = k_0 + k_1 f + k_2 f^2
\]

Noise is amplified, and then integrated over filter bandwidth to determine RMS noise at filter output. For standard filter shapes, Personik has tabulated these integrals.

Dividing this by the receiver DC gain yields the input-referred RMS noise current.

Assuming Gaussian statistics, a 6:1 signal/noise ratio yields $10^{-9}$ error rate

The average input photocurrent at sensitivity is then 6 times the frequency-integrated input-referred noise current.
Sensitivity Expressions

Receiver sensitivity

\[ \bar{P} \cong 6 \frac{hv}{\eta q} \cdot \sqrt{\frac{32 \pi k T C_p B^3 I_3}{f_\tau}} + \frac{4kT B I_2}{R_f} \]

B is the bit rate. For a square optical pulse and a raised-cosine eye, \( I_2 = 0.68 \) and \( I_3 = 0.12 \). Sending impulses (solitons) will make \( I_3 \) somewhat smaller.

\[ I_2 = \left[ \frac{1}{ BH(0) } \right] \cdot \int H(f) df \]

\[ I_3 = \left[ \frac{1}{ B^3 H(0) } \right] \cdot \int f^2 H(f) df \]

Note at 100 GB/sec, a 5 \( \mu \)m \times 5 \( \mu \)m \times 0.2 \( \mu \)m detector, \( f_t = 200 \) GHz, \( \Gamma = 1.5 \), that \( R_L = 209 \Omega \) results in equal FET and resistor noise.

At 40 GB/sec, the point of equal contribution is \( R_L = 1.3 \, k\Omega \).
Data Rate, bits/sec

FET f = 200 GHz
FET L = 1.5
5 µm x 5 µm x 0.2 µm detector

Optical Receiver Sensitivity for Square and 3-Pulses

receiver sensitivity, nP, dBm

Impulses
Square pulses (normal)
Noise Analysis

Simplify for now to focus on key issues.
More Notes on C.G. opt. recur:

FET2 noise, input-referred: \* \( R_f \to \infty \)

\[
\frac{4kT \Pi (2\pi f)^2}{9m_2} \left( C_{gs2} + C_{gd1} \right)^2 = \frac{\sqrt{\langle I_I^* \rangle}}{I_{ref}} = A
\]

(1 + \( \frac{2f}{f_m} \)) \* \( \frac{1}{2f} \) \* \( I_{ref} \)

FET1 noise, input-referred.

\[
\frac{4kT \Pi (2\pi f)^2}{9m_1} \left( C_{gs1} + C_p \right)^2 = \frac{\sqrt{\langle I_I^* \rangle}}{I_{ref}} = B
\]

\[
[\frac{4kT \Pi (2\pi f)^2 C_{gs2}^2}{9m_1}]^4 \quad = \quad \frac{4}{B} \quad \text{because } C_{gs1} = C_p
\]

Minimizes \( B \)

Examine the ratio of second to first-stage noise; assuming \( C_{gs2} >> C_{gd1} \)

\[
\frac{A}{B} = \left( \frac{C_{gs2}}{C_{gs1}} \right)^2 \cdot \frac{1}{4} \cdot \frac{1}{9m_2} = \frac{1}{4} \left( \frac{W_{g2}}{W_{g1}} \right)
\]

Note that \( W_{g2} \leq W_{g1} \) will ensure negligible FET2 noise contribution.

\*(drops for now - but check the effect later)\*
So, is that the end of the design effort?

No - we must now add the load resistor $R_f$ and try to get adequate bandwidth without sacrificing sensitivity.

\[ C_{gs1} + C_p = 2C_{gs1} \]

The pole at "X" is at \( f_1/2 \), while at "Y" it is at \( f_2 = \frac{(1-A_v)}{2\pi R_f (C_{gs2} + C_{gd1})} \).
Compare now to the simple transimpedance stage:

\[ C_{gs1} + C_p = 2 C_{gs1} = 2 C_p \]

For equal values of \( R_f \), this equal noise penalties from \( R_f \), the trans-\( z \) has larger capacitance (hence lower B.W.) by the ratio

\[
\frac{2 C_p}{C_{gs2} + C_{gd1}} = \frac{2 C_{gs1}}{C_{gs2} + C_{gd1}} \approx \frac{2 W_{g1}}{W_{g2}}
\]

If we make \( W_{g2} / W_{g1} = 1/2 \), we get fair improvement in bandwidth with the \( C_g \) stage added.
So, it would seem that a reasonable workable solution would be to make 
FET 2 have $\frac{1}{2}$ the gate width 
of FET #1. We will find that the 
noise penalty of FET 2 is only $\frac{1}{8}$ 
($= \frac{1}{16}$ in $\sqrt{2}$), hence $\sim \frac{1}{16}$ in $\sqrt{2}$.) of the 
noise of FET 1, and that for equal 
transimpedance - stage & bandwidth, 
that the resistor $R_f$ cap key is 

\underline{four times larger}. 

Why not simply make $Q_2$ much smaller than $Q_1$, given that an "exact" analysis gives minimum $Q_2$ noise, and bigger stage-2 bandwidth, if $Q_2$ is very small—in fact equal to $W_{q_1} \times \frac{C_{q_{11}}}{C_{q_{61}}}$?

Because:

1) We can't make $C_{q_{11}}$ that small
   
   ($W_{q_2} = W_{q_1} \times \frac{C_{q_{11}}}{C_{q_{61}}}$)

2) The second-stage gain becomes vanishingly small—putting a considerable burden on the noise performance of the third stage.
Instead, we should content ourselves with making the $CG$ preamplifier four times better.

**Proposed solution #1**

Q1: $C_{g_1} = \frac{C_p}{Wg_1}$

Q2: $Wg_2 = Wg_1 / 2$

Pick $R_f$ so that

\[
\frac{4kT}{R_f} \frac{LB}{I_2} = \left(\frac{1}{2}\right) \times \frac{32\pi^4 kT C_p}{L} \frac{LB^3 I_3}{R_f}
\]

the factor of $\frac{1}{2}$ being negotiable...
Note that $I_2 = 0.68$ $I_3 = 0.12$

Now a 5 µm x 5 µm x 0.2 µm detector has $C_{p} = 14 fF$, but we'll use $25 fF$ to allow for stray C.

$C_{p} = 25 fF$ $f_T = 100$ GHz

$C_{gs1} = 25 fF$ $g_{m1} = 20 C_{gs1} f_T = 15.7 ms = 164.2$

$C_{gs2} = 12.5 fF$, $g_{m2} = g_{m1}/2 = 1128.2$

\[
\frac{32\pi k T \sqrt{2} C_{p} I_3^2}{f_T} = 1.87 \times 10^{-4} \text{ A}^2 \quad \Theta \quad 100 \text{ Gals.}
\]

− allowing the noise contribution of $R_f$ to be 50% of this gives us $R_f = \sqrt{120 \Omega}$ ($B = 100$ Gbps)

750 Ω ($B = 40$ Gbps)

These are minimum values.
Recall that:

\[ R_{in} = \frac{RF + RL}{1 + gmRL} \]

\[-Z_T = \frac{RF}{1 - gmRL} - \frac{1}{1 + gmRL} \]

To this point the circuit looks:

100 GB/sec. Choose \( RL = \infty \)!

& choose \( RF = 240 \Omega \)!

\[ Z_T = RF - \frac{1}{gm^2} = 240 \Omega - 128 \Omega = 112 \Omega \]

\[ R_{in} = \frac{1}{gm^2} = 128 \Omega \]

Second pole position: (neglecting \( C_{gs2} \))

\[ f_{p2} = \frac{1}{2\pi (128 \Omega) C_{gs2}} \]

\[ f_T \approx 100 \text{ GHz}! \]

Performance gets better & better as \( RL \) gets bigger -- maybe not so.
\[ a_1 = \frac{(C_{gs3} + C_{gs2})}{g_{m2}} \]
\[ a_2 = \frac{C_{gs3} C_{gs2} R_f}{g_{m2}} \]
\[ \frac{a_2}{a_1} = \frac{C_{gs3} C_{gs2} R_f}{g_{m2}} \]

- Underdamped response will be observed if \( g_{m2} R_f < 1 \)
- A trouble.
Optimum Noise Design of Multi-Gigabit Common-Gate Optical Receivers

B. Agarwal and M.J.W. Rodwell

Department of Electrical and Computer Engineering,
University of California, Santa Barbara, CA 93106.

Abstract

1. Introduction

Ultra-wideband photoreceivers are needed for multi-gigabit optical transmission systems. Flat photoreceiver front-end response is desired over several decades of frequency to avoid post-equalization problems. High sensitivity at high data rates require low front-end noise at these high frequencies. The standard approach used is a front-end with a photodiode in conjunction with a transimpedance or integrating preamplifier. Several such monolithic [2] and hybrid [3] optical receivers with a p-i-n/MSM photodiode and HEMT/HBT for operation at 10-15 Gbits/sec have been reported. Excluding the use of inductive peaking [4,5] which can provide moderate noise reduction for most receiver configurations, integrating optical preamplifiers attain the lowest noise but have low bandwidth and require equalization. At low data rates, transimpedance amplifiers can attain similar noise performance. At very high data rates (>10 Gbits/sec), a transimpedance amplifier is extremely difficult to realize with the present transistor technology, as very large transistor gain-bandwidth products are demanded. Addition of a common-gate stage to the input of the transimpedance preamplifier with appropriate selection of circuit element values allows
the receiver bandwidth to be increased to one-half of the HEMT current-gain cutoff frequency, while attaining the same noise performance as the integrating front-end. Common-gate receivers were demonstrated earlier [6], but the authors presented no noise or gain-bandwidth analysis and concluded that the common-gate stage degraded the noise performance. Here, we show that, if appropriately designed, the common-gate/transimpedance based receiver can attain both wide bandwidth and very low noise. By combining this circuit configuration with advanced low-capacitance InGaAs photodetectors and AlGaAs/InGaAs/GaAs PHEMTs, optical receivers with -22 dBm sensitivity at 25 Gbits/sec and -20 dBm at 40 Gbits/sec are attainable.

II. Analysis of the Transimpedance amplifier

In Fig. 1(a) is shown a simple transimpedance amplifier as is predominantly used in photoreceivers [7]. $R_f$ is the feedback resistor, while the input HEMT Q1, the gain stage $A$, and $R_f$ form an inverting amplifier with gain $A = -g_{m1}R_fA$. A simple first-order noise model of the HEMT (Fig. 1(b)) is generally used in optical receiver design. Neglecting all secondary noise sources and dropping non-dominant terms, the input referred equivalent noise current spectral density is given by

$$\frac{d\langle I_eI_e\rangle}{df} = \frac{4kT}{g_{m1}^2}(2 \pi f)^2 \left(C_p + C_{gs}\right)^2 + 4kT / R_f$$  (1)

where $C_p$ is the photodiode capacitance, $f_c$ is the HEMT current-gain cutoff frequency and $\Gamma$, its channel noise coefficient. Using the optimum HEMT size (that which gives $C_p = C_{gs}$), this becomes
\[ \frac{d\left(1-A^*\right)}{dt} = 32\pi kT f^2 C_p / f_z + 4kT / R_f \] (2)

At the gate of Q1, the node impedance is \( R_f / (1 - A) \) and the node capacitance is \( C_p + C_{g\text{m}1} = 2C_p \). Hence, the preamplifier 3 dB bandwidth is given by

\[ f_{3\text{db}} = (1 - A) / 4\pi C_p R_f \] (3)

We would like to make \( R_f \) large so that it contributes negligible noise. In the limit of very large \( R_f \), we are left with the integrating front-end, which results in low front-end bandwidth and introduces significant difficulties with post-equalization. To attain large bandwidth, while still having a large \( R_f \), the preamplifier should have a large gain \( A \). To avoid loop instability, \( A(j\omega) \) should have no poles below the loop bandwidth \( f_{3\text{db}} \). Both, very high gain and wide bandwidth is difficult to attain, given finite transistor cut-off frequency \( f_c \). Hence, multi-gigabit receivers are generally designed with \( R_f \) reduced to the point where a front-end bandwidth of about 50-75% of the data rate is attained, and sensitivity is sacrificed.

III. Analysis of common-gate/transimpedance preamplifier

These difficulties are overcome by the common-gate/transimpedance preamplifier (Fig. 1(c)). At the source of Q1, the node impedance is \( 1 / g_{\text{m}1} \) and the node capacitance is \( C_p + C_{g\text{m}1} = 2C_{g\text{m}1} \). Hence the input pole frequency is \( f_z / 2 \). The dominant pole is at the gate of Q2, where the node impedance is \( R_f / (1 - A) \) and the node capacitance is \( C_{g\text{m}1} + C_{g\text{m}2} \).

\[ f_{3\text{db}} = (1 - A) / 2\pi(C_{g\text{m}1} + C_{g\text{m}2}) R_f \] (4)
As before, taking only the dominant noise terms, the input referred equivalent noise current spectral density is given by

\[
\frac{d\langle I_n^X \rangle}{df} = \frac{4kT}{g_m^1} (2\pi f)^2 (C_p + C_{sd1})^2 \\
+ \frac{4kT}{R_f} (1 + (2\pi f)^2 (C_p + C_{sd1})^2 / g_m^2) \\
+ \frac{4kT}{g_m^2} (1 / R_f^3 + (2\pi f)^2 (C_{sd1} + C_{sd2})^2)
\] (5)

The optimum gate widths of Q1 and Q2 are those which give \( C_{gs1} = C_p \) and \( C_{gs2} = C_{sd2} \) simplifying (5) to

\[
\frac{d\langle I_n^X \rangle}{df} = \frac{32\pi kT f^2 C_p / f_i + 4kT}{R_f} (1 + (2f / f_i)^2) + \frac{4kT}{g_m^2 R_f^2}
\] (6)

where the approximation \( C_{gs} >> C_{gs} \), which is true for a typical HEMT, is used. From equations (2) and (6), the channel noise of HEMT Q1 is same for both configurations.

If both are designed for the same bandwidth, \( R_f \) for the common-gate configuration can be much larger (by the ratio of \( C_{gs1} \) to \( C_{gs2} \)) than for the simple transimpedance amplifier, thus giving a noise performance similar to that of an integrating front-end. For equal noise currents from both (i.e. same \( R_f \)), the common-gate receiver will have a larger bandwidth by the ratio of \( C_{gs1} \) to \( C_{gs2} \). In either case, the common-gate receiver works better than the simple transimpedance receiver.

IV. Design Illustration

We illustrate here, a design for 20-40 Gbits/sec reception (Fig. 2). Complete small-signal and noise models for the 0.3 μm gate length pseudomorphic HEMTs are taken from [8] while a 100 GHz 5μm x 5μm GaInAs/lnP p-i-n photodiode [9] is assumed. The widths of
the HEMTs Q1 and Q2 are chosen for low input noise. The feedback resistors are chosen for large bandwidth while not degrading the noise performance. Q1 is the common-gate stage while Q2 and $R_f$ form the transimpedance stage. Q3 and Q4 provide both, voltage gain and buffering to drive a 50 $\Omega$ load.

Fig. 3 shows SPICE simulations of transimpedance gain and input noise current using complete device models. The simulation includes 25 fF parasitic assembly capacitance consistent with MMIC assembly. The 3 dB bandwidth is 20 GHz and the input referred noise current is 14 pA/sqrt. Hz at 20 GHz. The predicted receiver sensitivity [7] at 10$^{-9}$ BER (Fig. 4) is -19 dBm at 40 Gbits/sec. For comparison, the predicted sensitivity of an integrating front-end is also shown. It can be seen that the common-gate receiver approximates the minimum attainable noise of an integrating front-end.

V. Conclusions

We have analyzed the simple transimpedance preamplifier and the common-gate/transimpedance preamplifier for use in optical receiver front-ends. Our analysis shows that, for the same bandwidth of the two circuits, the common-gate-transimpedance preamplifier is less noisy and hence, more sensitive. On the other hand, if both are designed for the same noise performance, the common-gate/transimpedance preamplifier has a higher gain-bandwidth. In the case of the common-gate/transimpedance preamplifier, the trade-off between noise and bandwidth is less than in the case of the simple transimpedance amplifier. SPICE simulations of our front-end design using state-of-the-art devices predict satisfactory performance for 20-40 Gbits/sec. In conclusion, we would like to say that the common-gate/transimpedance preamplifier is a promising candidate for high bitrate photoreceivers.
References

[1] Brian Kaspar


Figure Captions

Fig. 1 Simple transimpedance amplifier (a), Simplified HEMT noise model (b), and Common-gate/transimpedance preamplifier (c)

Fig. 2 Schematic diagram of common-gate/transimpedance photoreceiver

Fig. 3 SPICE simulations of transimpedance gain and input referred equivalent noise current spectral density

Fig. 4 Predicted sensitivity of common-gate/transimpedance preamplifier and integrating front-end using Personick coefficients I_2=0.68 and I_3=0.12
Receiver Implementation: In Summary

100 GB/sec Optical Receiver

40 GB/sec Optical Receiver
Block diagram.

Photodetector and preamplifier

"Linear channel": high gain broadband AGC amplifier

Nyquist Filter

Decision circuit

Timing Recovery (phase-locked-loop)

Data output

Voltage input

Timing input
Sensitivity Comparison, HBT phototransistor vs HBT-PIN, PIN-FET and APD-FET

![Graph showing sensitivity comparison between different photodetector types.]

HBT and HBT phototransistor:
F<sub>T</sub>=80 GHz, F<sub>max</sub>=44 GHz, β=100

- Front-illuminated HBT phototransistor
- HBT and base-collector photodiode
- FET F<sub>T</sub>=20 GHz, 0.1 pF detector
- FET F<sub>T</sub>=60 GHz, 0.1 pF detector
- FET F<sub>T</sub>=200 GHz, 0.1 pF detector
- FET F<sub>T</sub>=20 GHz, 0.5 pF detector

NRZ data rate, bits/second

Figure 9. HBT "linear-channel" AGC amplifier
Phase-locked loop.

Diagram showing the components of a phase-locked loop, including an untimed data input, a phase detector, and a voltage-controlled oscillator.
AGC AMPLIFIER

Gain control

Outputs

ACG detector