• 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 reciever front ends

ECE — , Notes set 23

Fiber optic digital Communications:

- How it differs from prior material

- Smith & Personick's Method

- Performance Estimates.

We need to return to our prior material:

optimum receiver strategy:

1) Message sent by message vector / sequence

   - represented by basis functions φ_i

2) Message now usually 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?

ex// ★ narrowband digital Microwave receiver

F = noise figure = F_0 = constant across band.

⟹ white system Noise

★ Very broadband digital Microwave receiver:

Fano's limit: can't match for noise across whole bandwidth

$$F = F_0 + K(f - f_0)^2$$

f_0: channel center frequency.

- noise no longer white.

★ Baseband PCM fiber optics

$$\left.\begin{array}{l}\text{Input noise current}\\ \text{spectral density}\end{array}\right\} = K_1 + K_2 f^2$$

⟹ <u>Not</u> <u>white</u>

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 acknowledges 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 & difficulty of implementation.

Example to clarify this:

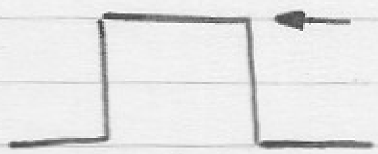
Fiber-optic receiver:

* input referred noise power spectral density

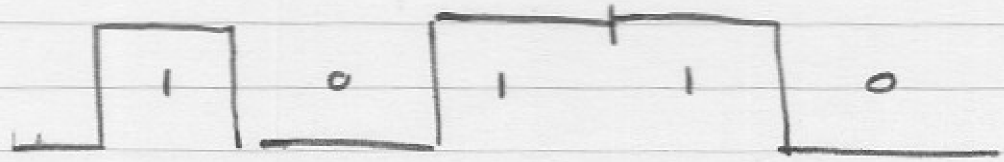
$$N_0 \simeq a + bf^2$$

* Signalling format: on-off keying.

* Basis function



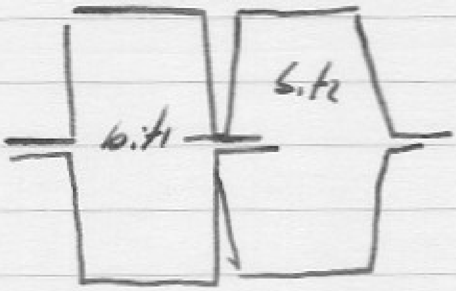
t_{bit}



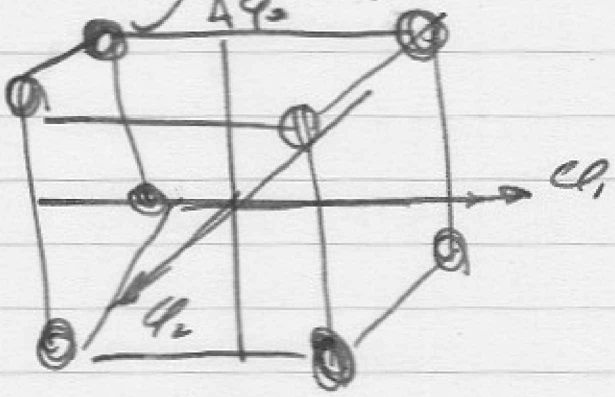
The fiber people call this Non-return-to-zero (NRZ)

Better Name: OOK with square symbol.

Note that the bits are orthogonal



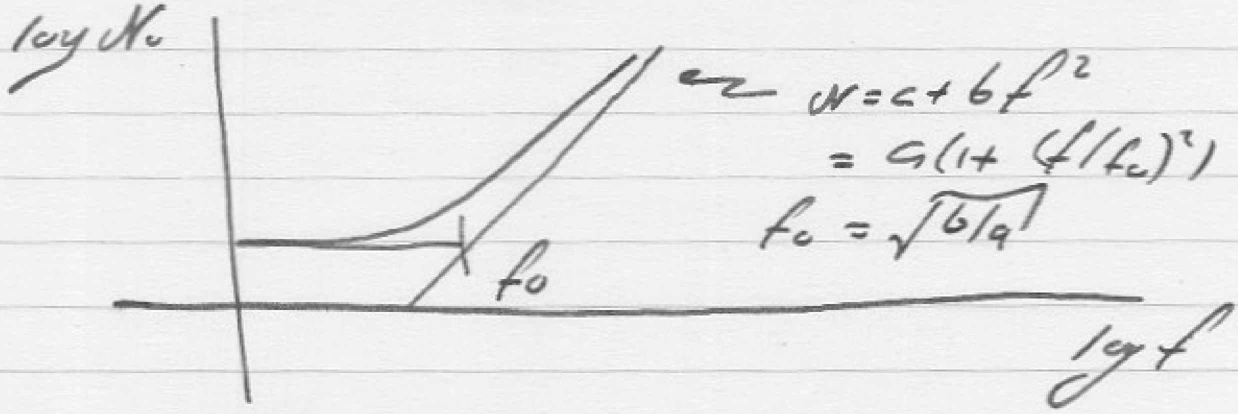
... so that, although we could in principle look at N bits as N boundary's in an N-dimensional signal space



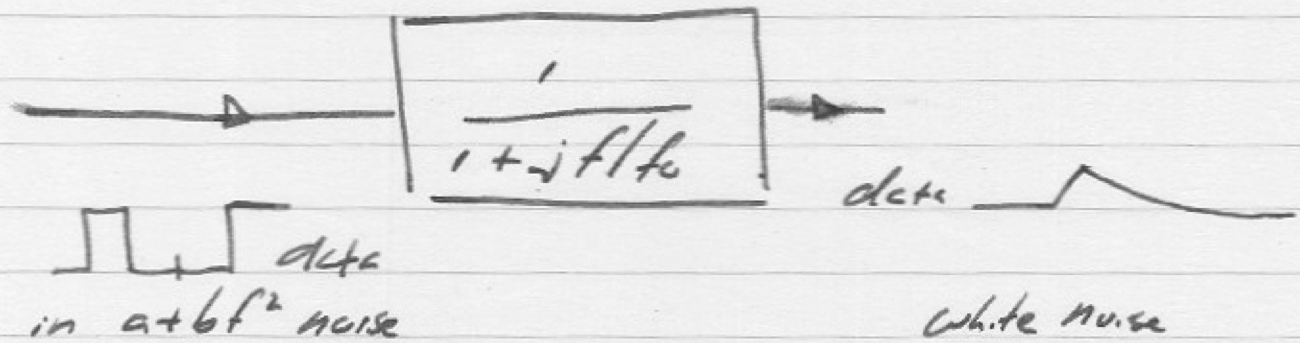
we can in fact look at one bit at a time

↳ nice & simple receiver

but the noise has p.s.d. like so:



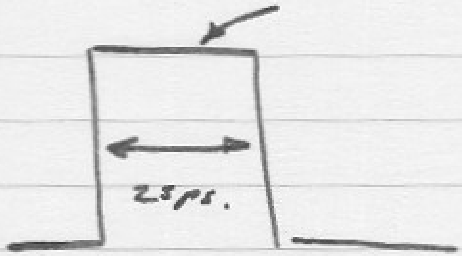
So an optimum receiver must pass the signal + noise first through a whitening filter:



TbL 40 Gb/sec system: $T_b = 25\,ps$

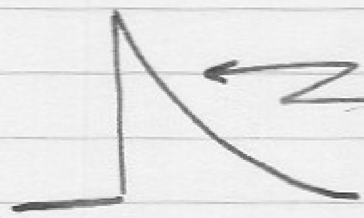
Bipolar receiver: $f_0 \sim 1$ GHz. ($\tau_0 = 160\,ps$)

<u>Input Pulse to filter:</u>



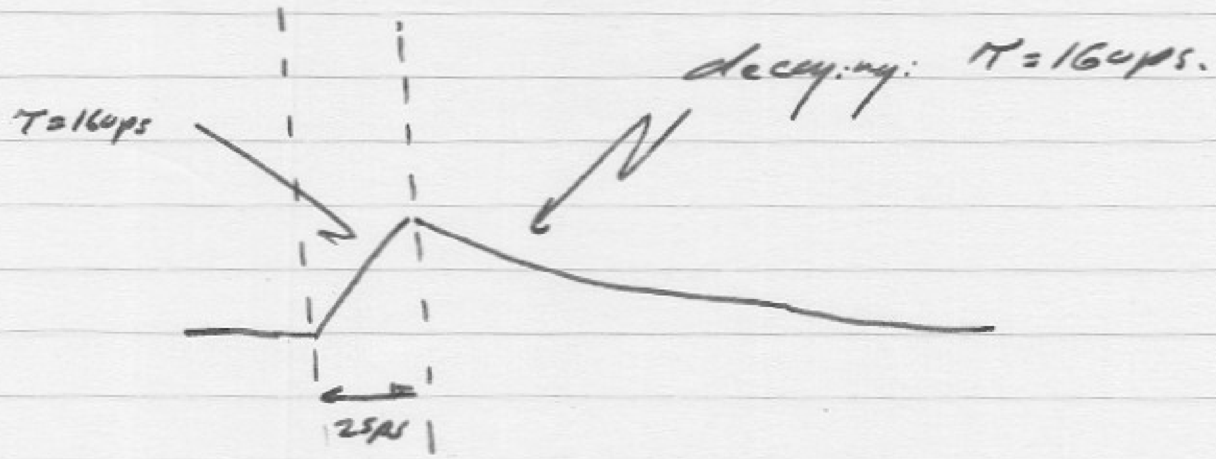
~~output~~

<u>Impulse response of filter:</u>



$k \cdot (1 - e^{-t/(160ps)}) \cdot u(t)$

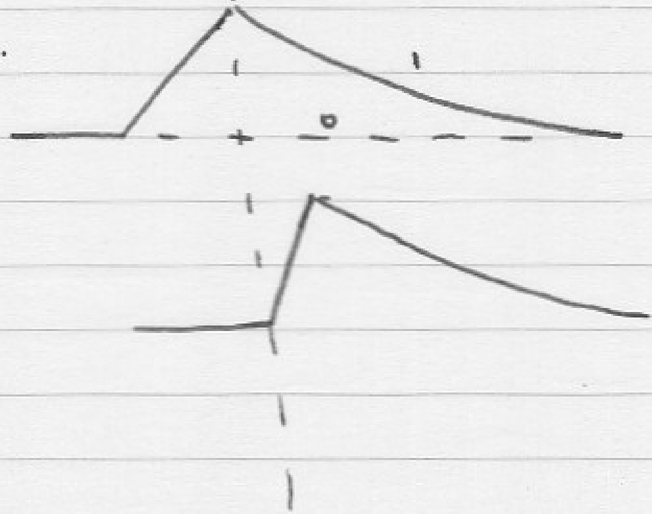
output pulse from filter



Problem:

bit n:

bit $(n+1)$

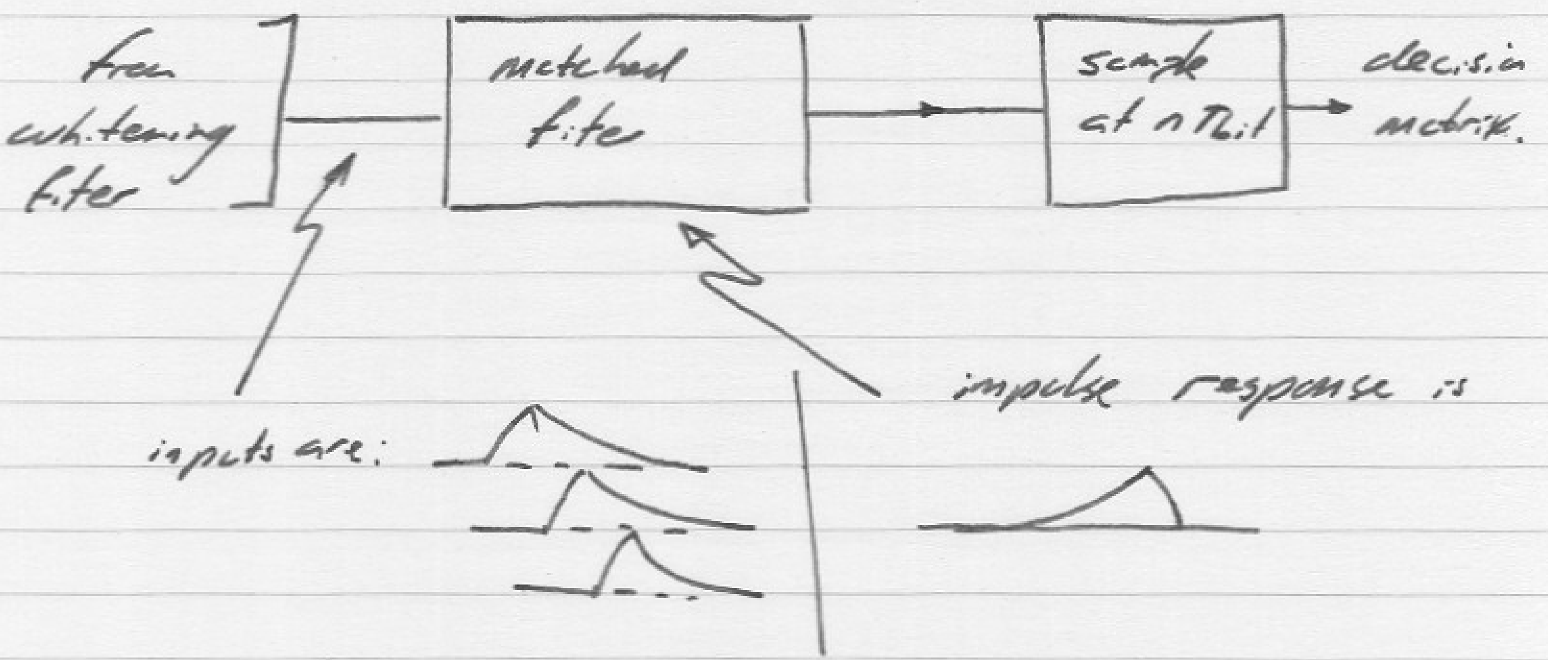


etc....

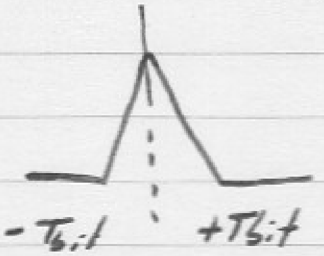
⇒ data bits no longer orthogonal.

⇒ our theory did not demand orthogonal signals to give an optimum receiver, but it did demand orthogonal bits if that optimum receiver were to be simple

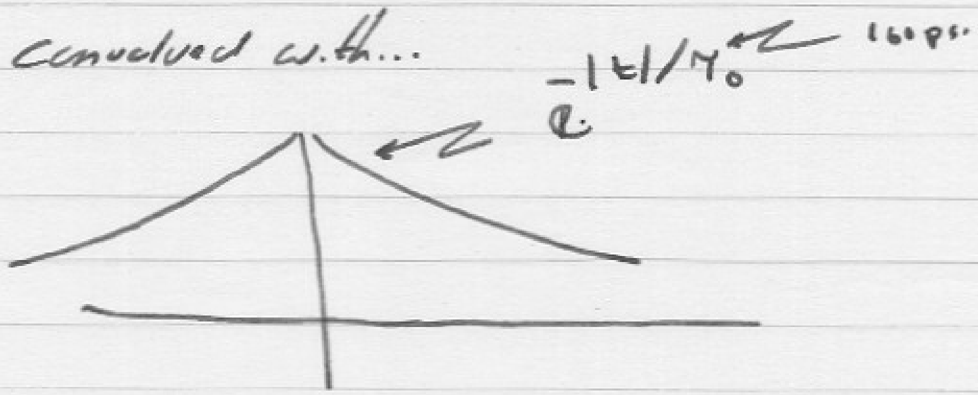
Receiver structure "optimum"



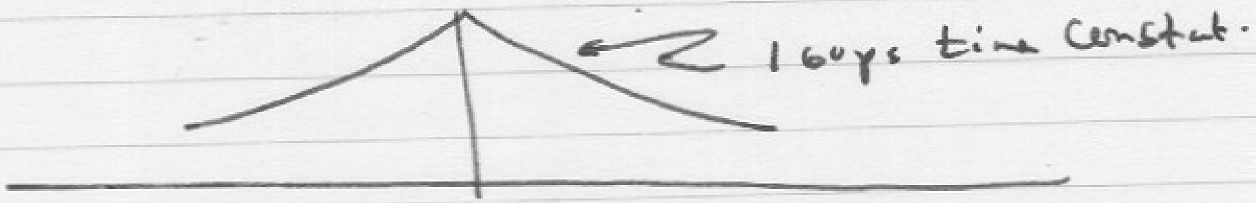
⇒ outputs of matched filter is therefore:



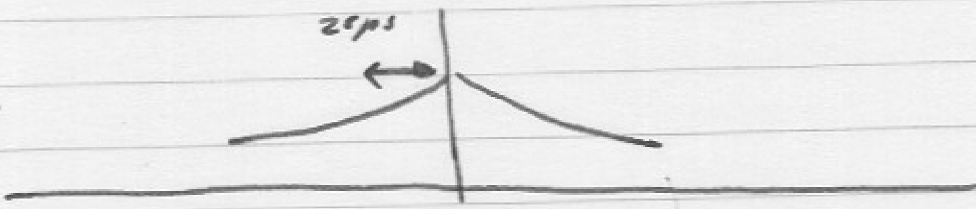
convolved with...



Response to single bit is ∴ approximately:



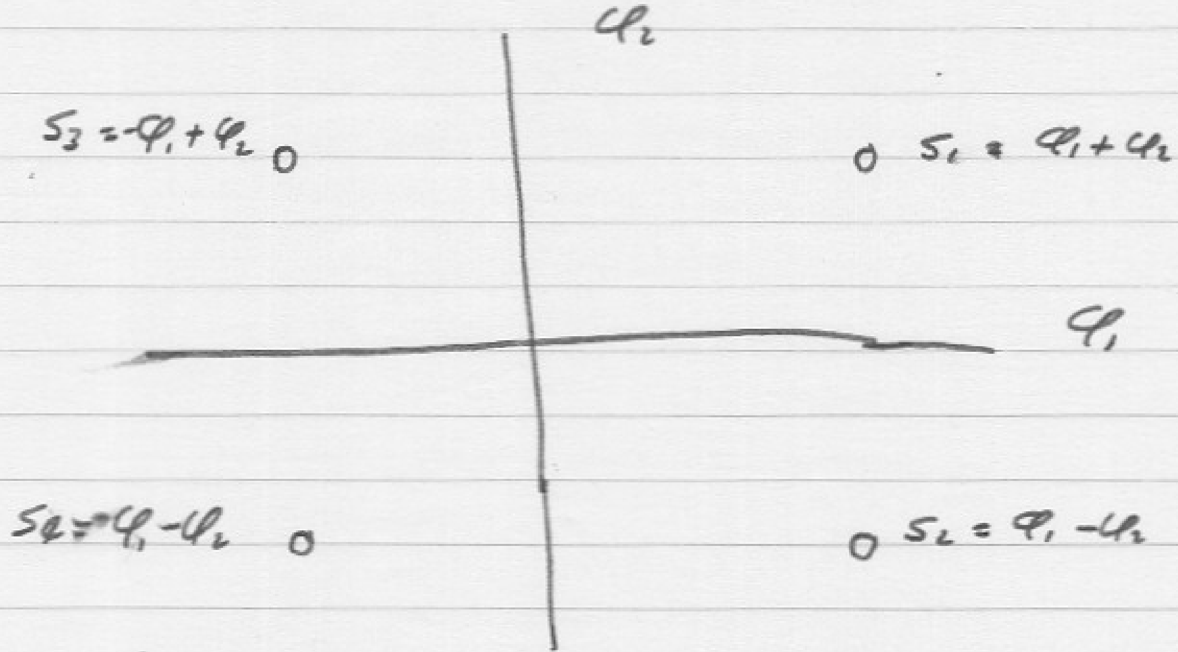
Next bit



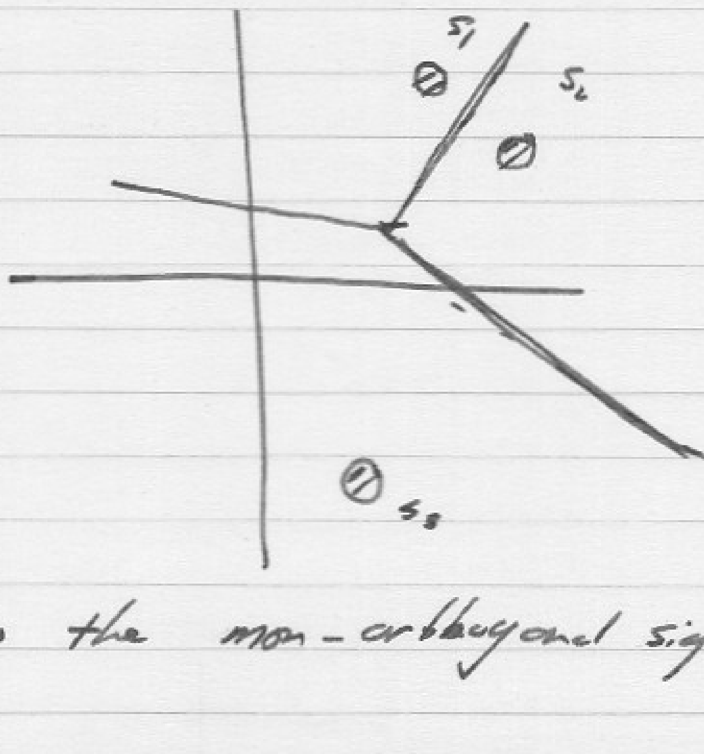
etc:

Very strong intersymbol interference.

again, Intersymbol Interference is ok:



no intersymbol interference allows us to decide on bit-by-bit basis (φ_1, φ_2, φ_3 ...)



these pictures pertain to the non-orthogonal signal cases

... but with 25ps-duration bits spread in time to ± 160ps, the decision algorithm must consider $\sim 2 \cdot (160/25) = 13$ successive outputs of the correlator / matched filter in order to determine a given bits' value.

→ Ideal hardware becomes straight unrealizable

⇓

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

- or -

Sacrifice substantial performance for extreme hardware simplification → Nyquist filter method.

* I will not discuss

decision feedback here.

* Nyquist Method:

* Extremely simple hardware.

* Replace whitening & matched filter / correllator

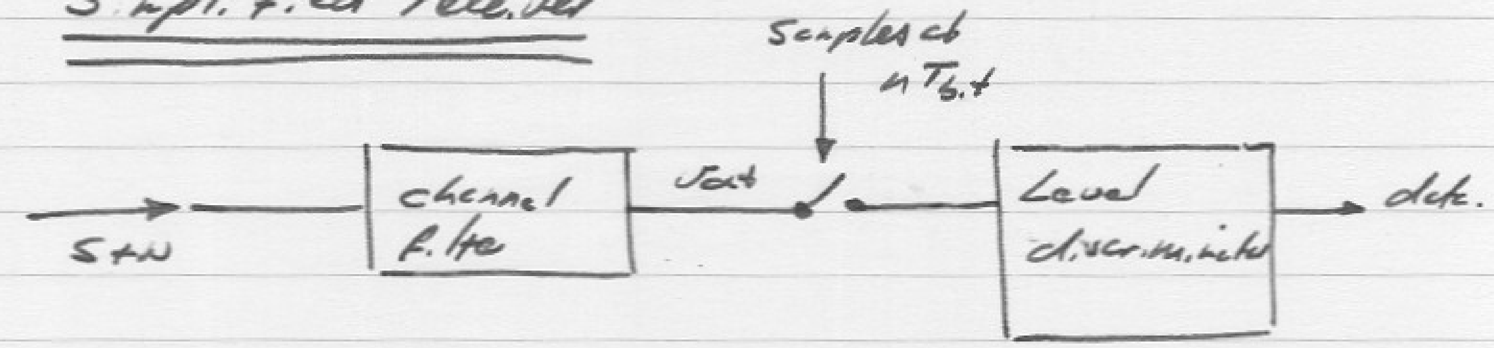
with simple noise-bandlimiting filter

* subject to constraint, zero

intersymbol interference: E.G.

Orthogonal bit slots

Simplified receiver



(1 - or - 0)

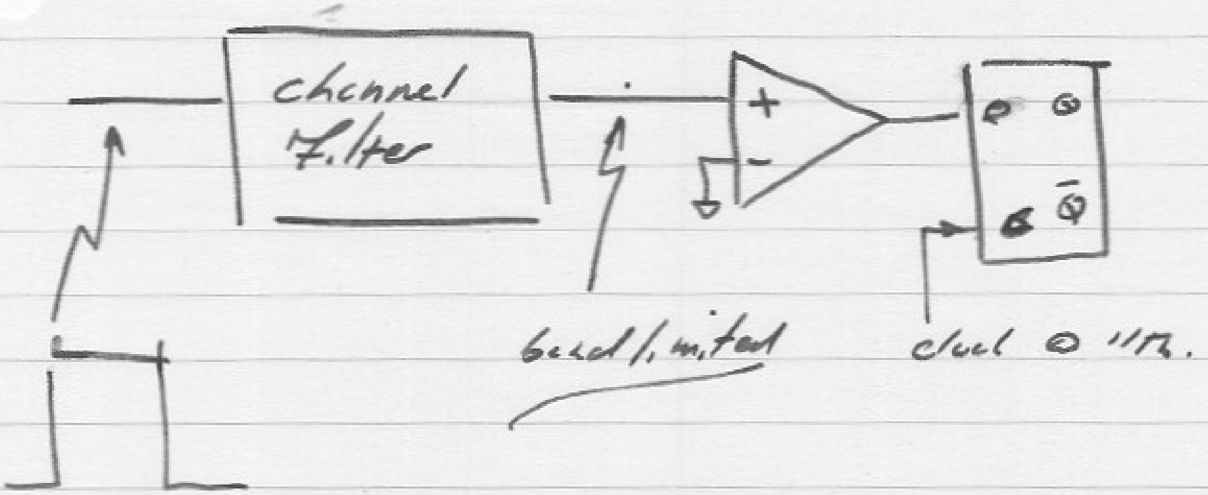
★ Channel filter constraint:

$\equiv$ $V_{out}(nT_{bit})$ depends only on value of n^{th} bit

- zero ISI - intersymbol interference.

$\equiv$ try to minimize noise...

★ order of $\left\{ \text{sampler} \right\}$ and discriminator can & often is reversed $\rightarrow$ becomes comparator & clocked latch.

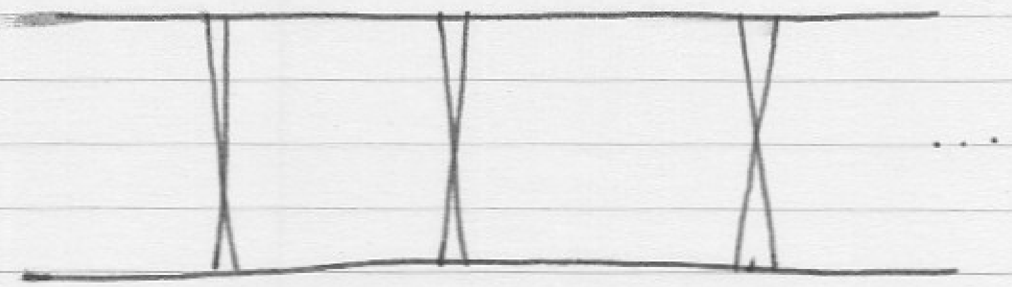

channel
Filter
+
−
D
Q
$\bar{Q}$
bandlimited
clock @ 1Hz.

Eye diagram:

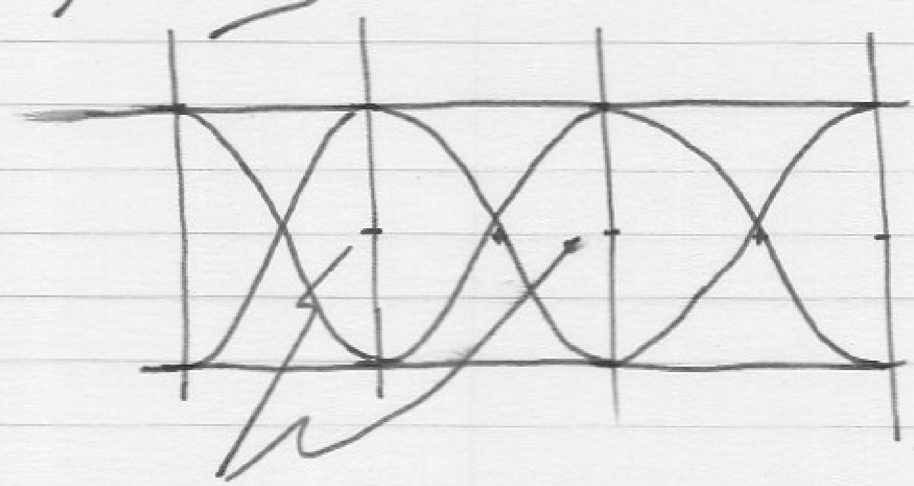
Born on an oscilloscope:

- scope triggered on clock
- scope displays data stream
- all possible trajectories superimposed.

Input Eye

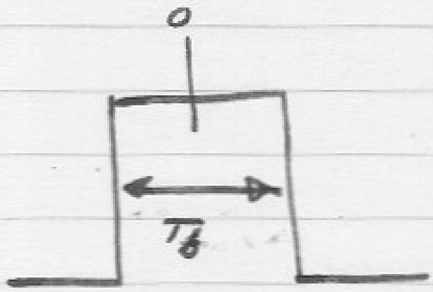


Output Eye (a good one)

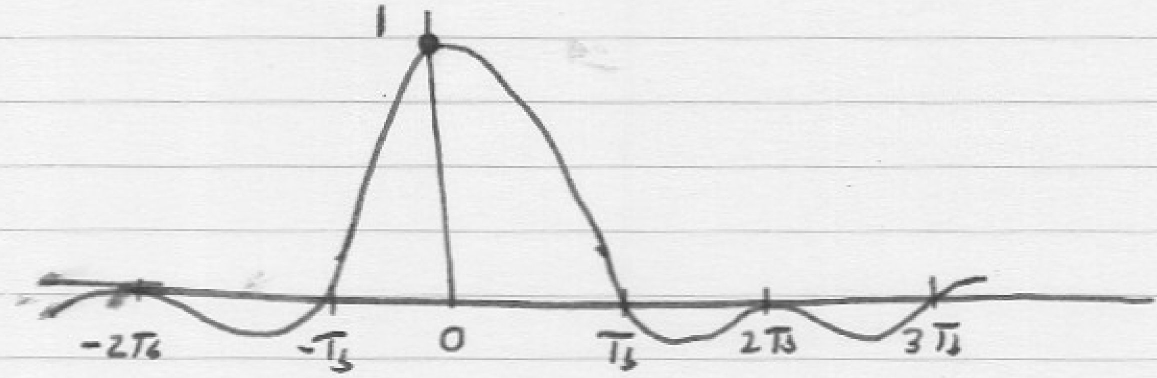


decision circuit looks at voltage at these points in time...

This demands that a bit input:



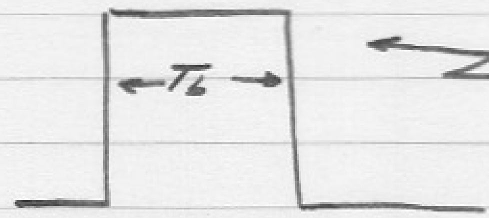
Produce (modulo a time delay) an output:



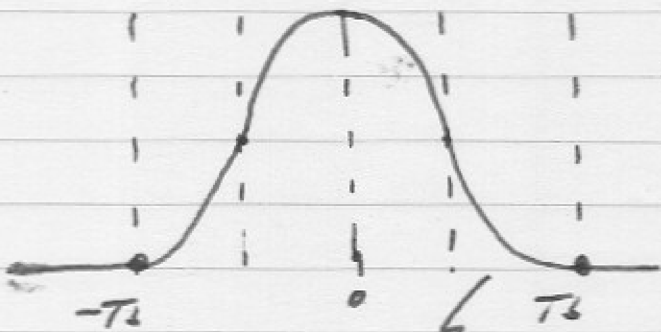
... which is <u>zero</u> at the center of all other bits.

Finding filters that do this is a general problem solved by Nyquist they are called Nyquist filters, although not pertaining to <u>sampling</u>

Smith & Pergomik considered a far more restrictive set of (idealized, not real) filters



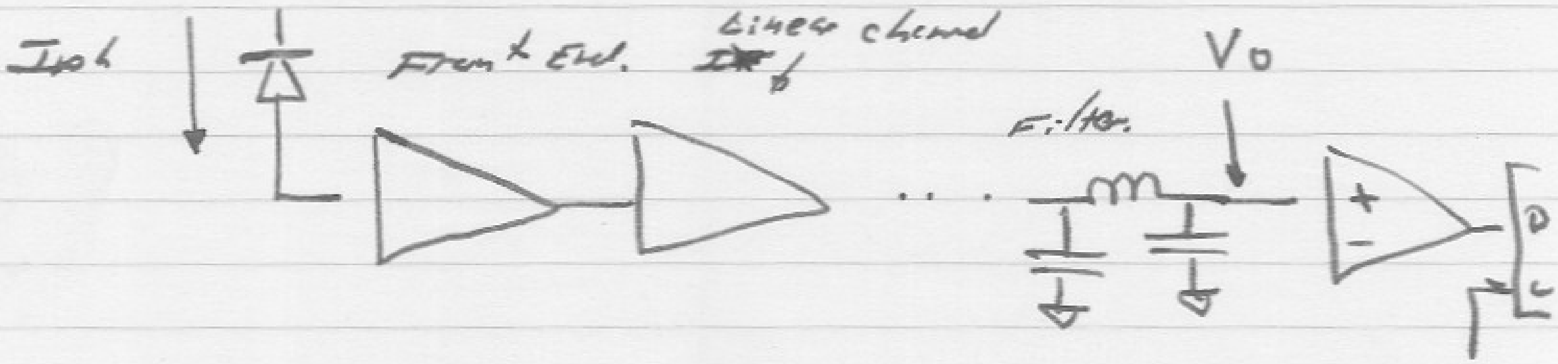
input



raised cosine:

$$\frac{\sin(\pi t/T_{bit})\cos(\pi t/T_{bit})}{(\pi t/T_{bit})\,(1-(2t/T_{bit})^2)}$$

How *they* do the analysis...



$$I_{ph} = P_{optical} \cdot \frac{q}{h\nu} \cdot \eta$$

1 electron per photon

detector quantum efficiency.

$$V_o = K \, Z_T(\omega) \, I_{ph}$$

Pick K so that $Z_T(\omega) \rightarrow 1$ at DC.

* We analyze input network to find <u>total</u> input-referred noise current:

$$\frac{d}{df}\langle i_n i_n^* \rangle = a + bf^2$$

more on this in a minute...

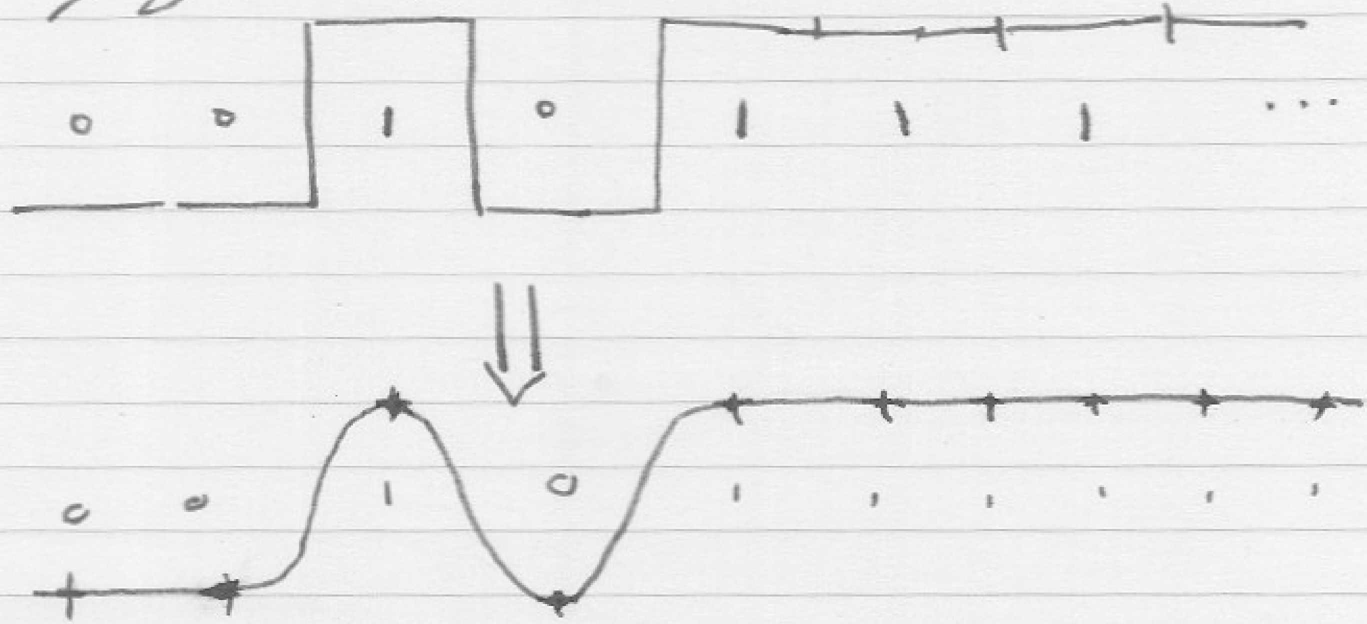
* The output noise voltage has variance

$$\sigma_{V_n}^2 = K^2 \int_0^{+\infty} \| Z_T(\omega) \|^2 \cdot (a + bf^2) \cdot df$$

* The output signal has magnitude:

$$V_{out} = I_{in} \cdot K \, Z_T(\omega = 0)$$

★ This has the key assumption of a Nyquist zero-ISI channel filter:



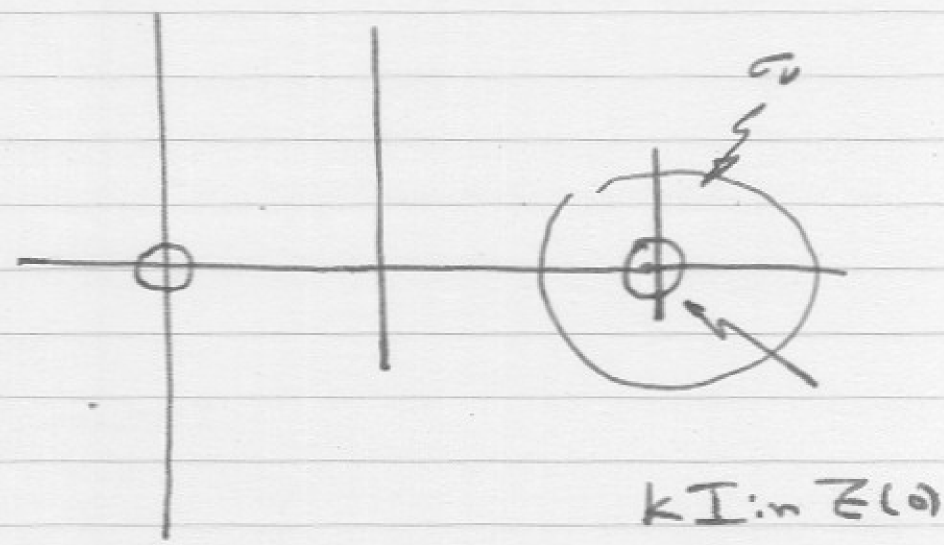
... so that the amplitude of an isolated bit is indeed $I_{in} \cdot k Z_T(\omega = 0)$

Then $\sigma_{v_n} = K \sqrt{\int_0^\infty \|Z_T(f)\|^2 (c + b f^2)\, df}$

and $V_{out} = k\, I_{in}\, Z_T(0)$

OOK

boundary $k I_{in} Z(0)/2$.



write $\sigma_{v_n} = K X$ for a moment.

$$P(\text{Error}) = Q\left[(k\, I_{in} Z(0)/2)/kX\right]$$

$$= Q\left[I_{in} Z(0)/2X\right]$$

as usual; $P(\text{Error}) = \text{e.g.} = 10^{-9}$

$$\longrightarrow I_{in} Z(0)/2X = \text{"}Q\text{"} = 6$$

$$I_{in} = \frac{2X \cdot Q}{Z(0)} = 2Q \sqrt{\int_0^\infty \left\| \frac{Z_T(f)}{Z_T(0)} \right\|^2 (a + bf^2)\, df}$$

where we have assumed $S_I(f) = a + bf^2$

so ~~of~~ in general:

$$I_{in} = 2Q \cdot \sqrt{\int_0^\infty \left\| \frac{Z_T(f)}{Z_T(0)} \right\|^2 S_I(f)\, df}$$

Receiver Sensitivity:

1 sent $\quad P_{opt} = P_1 \quad \Rightarrow \quad I_{in} = \frac{\eta q}{h\nu} P_1$

0 sent $\quad P_{opt} = 0 \quad \Rightarrow \quad I_{in} = 0$

$\bar{P}$ = average power = $P_1/2$

$$= I_{in} \frac{h\nu}{\eta q} \cdot \frac{1}{2}$$

$$\boxed{\bar{P}_{min} = \frac{h\nu}{\eta q} \cdot Q \sqrt{\int_0^\infty \left\| \frac{Z_T(f)}{Z_T(0)} \right\|^2 \cdot S_I(f)\, df}}$$

now usually: $S_I(f) = a + b f^2$

$$\langle I_n^2 \rangle \triangleq \int_0^\infty \left\| \frac{Z_T(f)}{Z_T(0)} \right\|^2 S_I(f)\, df$$

$$= \int_0^\infty \left\| \frac{Z_T(f)}{Z_T(0)} \right\|^2 \cdot a \cdot df$$

$$+ \int_0^\infty \left\| \frac{Z_T(f)}{Z_T(0)} \right\|^2 \cdot b f^2\, df$$

Lets normalize with respect to

Bit rate: B bits/second.

B bits/second $\rightarrow$ filter $Z_T(f)$

$2B$ bits/second $\rightarrow$ filter $Z_T(f/2)$

* We want to work with normalized filter frequency responses; e.g. normalized to the bit rate...

$$\tilde{Z}_T(f/B) = Z_T(f)$$

$$\langle I_n^2 \rangle = \int_0^\infty \left\| \frac{\tilde{Z}_T(f/B)}{\tilde{Z}_T(0)} \right\|^2 \cdot a \cdot df$$

$$+ \int_0^\infty \left\| \frac{\tilde{Z}_T(f/B)}{\tilde{Z}_T(0)} \right\|^2 \cdot b f^2 \, df$$

write $\tilde{f} = f/B$

$$= aB \int_0^\infty \left\| \frac{\tilde{Z}_T(\tilde{f})}{\tilde{Z}_T(0)} \right\|^2 d\tilde{f}$$

$$+ bB^3 \int_0^\infty \left\| \frac{\tilde{Z}_T(\tilde{f})}{\tilde{Z}_T(0)} \right\|^2 \tilde{f}^2 \, d\tilde{f}$$

$$I_2 = \int_0^{\infty} \left\| \frac{\tilde{Z}_T(\tilde{f})}{\tilde{Z}_T(0)} \right\|^2 d\tilde{f}$$

$$I_3 = \int_0^{\infty} \left\| \frac{\tilde{Z}_T(\tilde{f})}{\tilde{Z}_T(0)} \right\|^2 \tilde{f}^2 \, d\tilde{f}$$

.. are called Personik Coefficients.

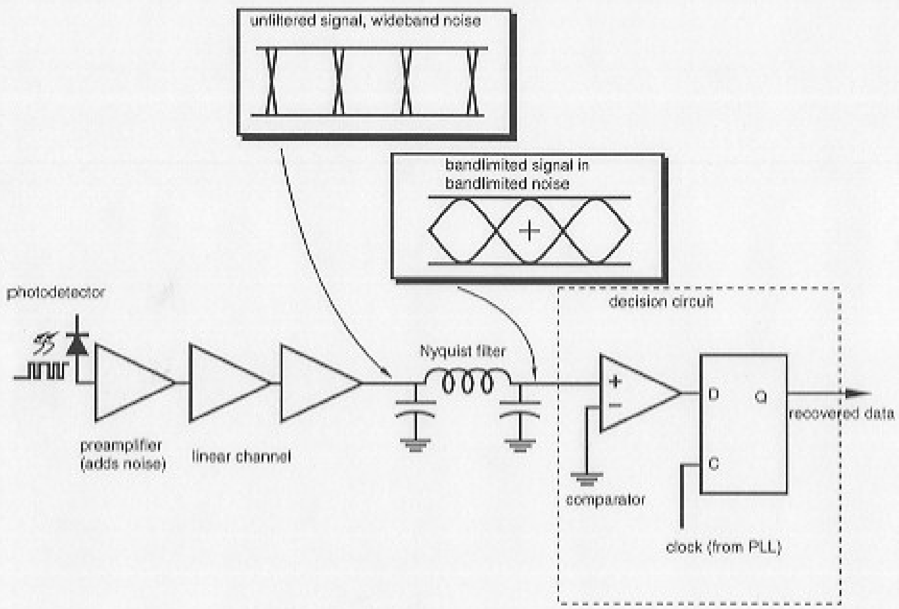
They are the integrals of the filter's transfer functions in normalized frequency (normalized to the bit rate) and therefore depend only on the shape of the filtered pulse.

$$\boxed{\begin{aligned} P_{min} &= \frac{h\nu}{q} \cdot Q \cdot \frac{1}{\eta} \cdot \sqrt{\langle i_n^2 \rangle} \\ \text{where} \\ \langle i_n^2 \rangle &= B \cdot I_2 \cdot a + B^3 I_3 \cdot b \\ \text{if } S_I(f) &= a + bf^2 \end{aligned}}$$

$I_2 \cong 0.68$
$I_3 \cong 0.12$ } parameters for a "real world" 5-pole L-C filter which approximates a raised cosine.

Raised cosine: I_2 about the same
$I_3 \sim 0.06$

Optical Receiver Basics



Photodetector converts light into current, drives preamplifier

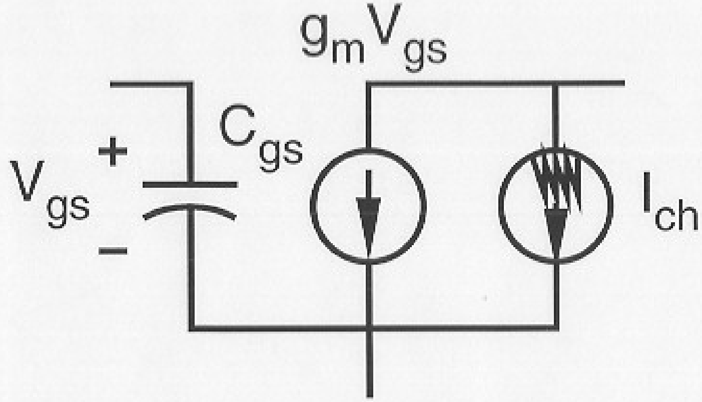
Preamplifier & photodetector set noise performance, hence attainable sensitivity

Nyquist filter bandlimits noise to ≈B/2.

Decision circuit recovers binary data

Key Component Models

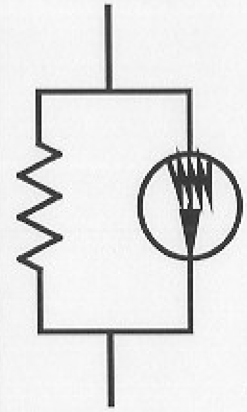
Field-Effect Transistors



$$\frac{d}{df}\left\langle I_n I_n^* \right\rangle = 4kT\Gamma g_m$$

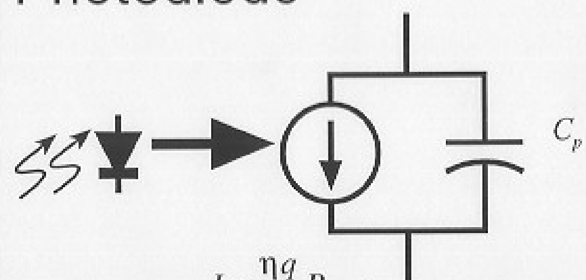
Other FET parasitics at kT
1/f noise not relevant at rates above 10 GB/s

Resistors



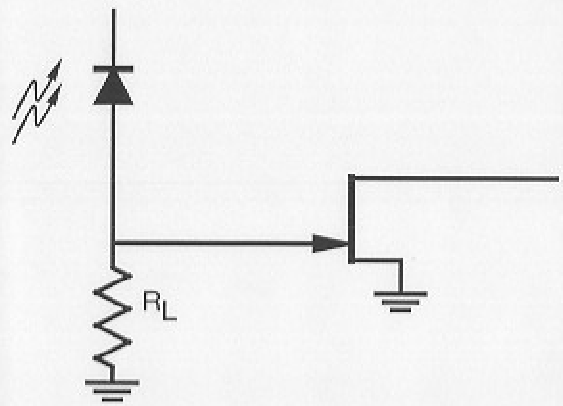
$$\frac{d}{df}\left\langle I_n I_n^* \right\rangle = 4kT / R$$

Photodiode



$$I = \frac{\eta q}{h\nu} P_{optical}$$

receiver noise analysis



Approximate input referred noise

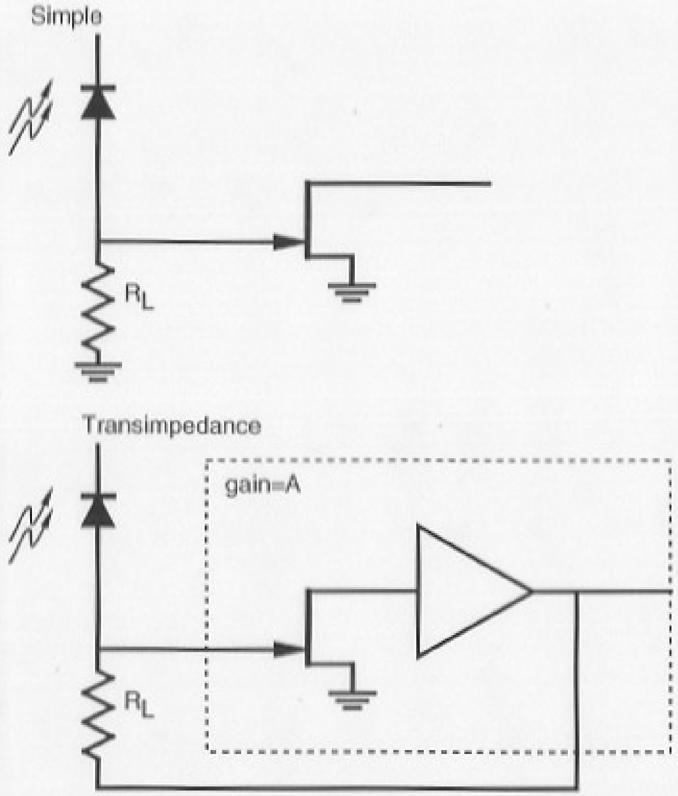
$$\frac{d\langle I_n I_n^* \rangle}{df} = \frac{4kT}{R_L} + \frac{4kT\Gamma}{g_m}(2\pi f)^2 \left(C_{gs} + C_p\right)^2$$

but $C_{gs} = g_m / 2\pi f_\tau$, so choose $C_{gs} = C_p$

hence:

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

Front-End Noise and Bandwidth



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

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

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

This assumes intelligent choice of FET size

•Bandwidth is

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

(set A=1 for simple stage)

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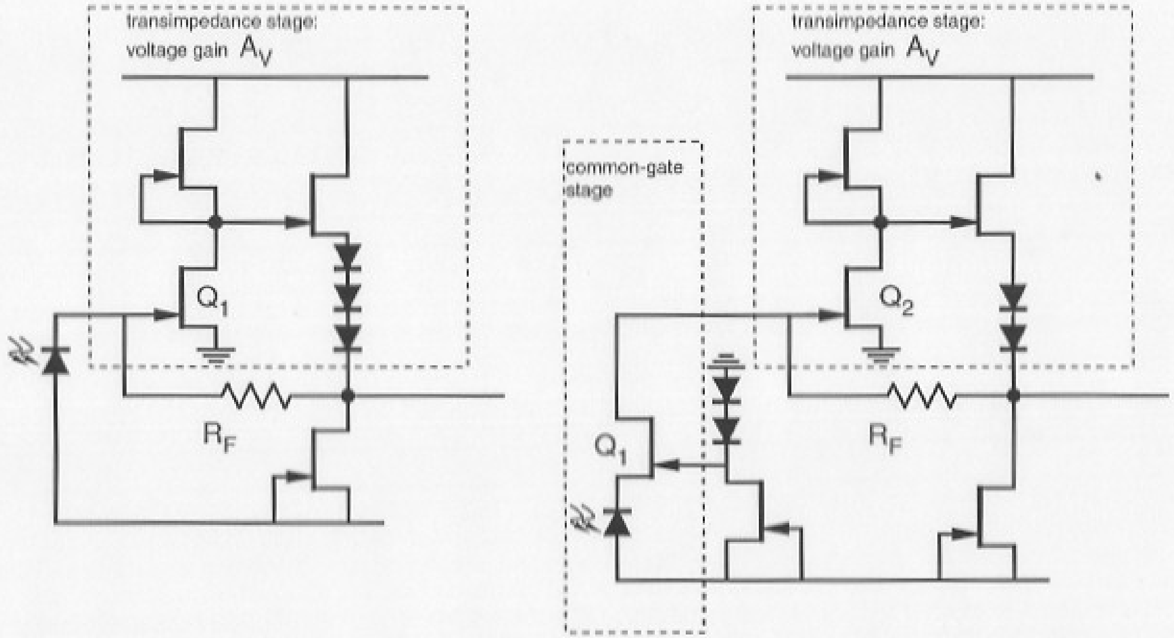
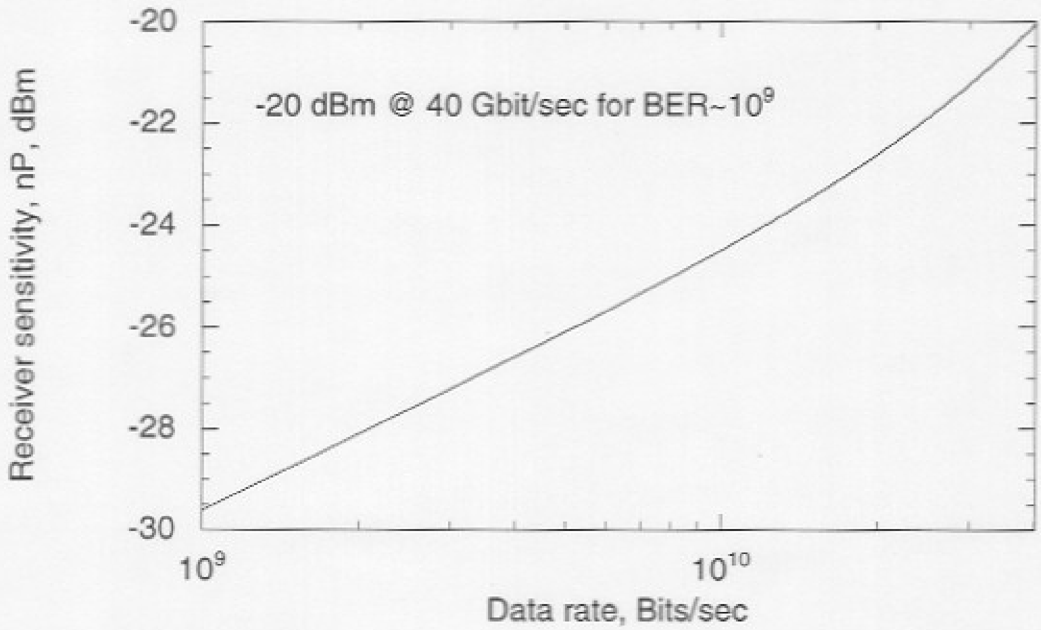
Common-Gate / Transimpedance Front-Ends

<table>
<tr><td>same noise voltage and current in CS and CG modes.

CG stage gives input pole at $f_\tau / 2$.

need high f_τ, small $C_{photodiode}$.</td><td colspan="2">
</td></tr>
<tr><td>input noise, given $C_{gs} = C_{photodiode}$:</td><td colspan="2">$$\frac{d\langle I_n I_n^* \rangle}{df} \cong 32\pi kT\Gamma f^2 C_p / f_\tau + 4kT / R_f$$</td></tr>
<tr><td>bandwidth:</td><td>$$f_{3dB} \approx \frac{1+A}{R_f(C_{diode} + C_{gs})}$$</td><td>$$f_{3dB} \approx f_\tau / 2$$</td></tr>
</table>

Optical Receiver Sensitivity: Design Target



Key observations regarding noise:

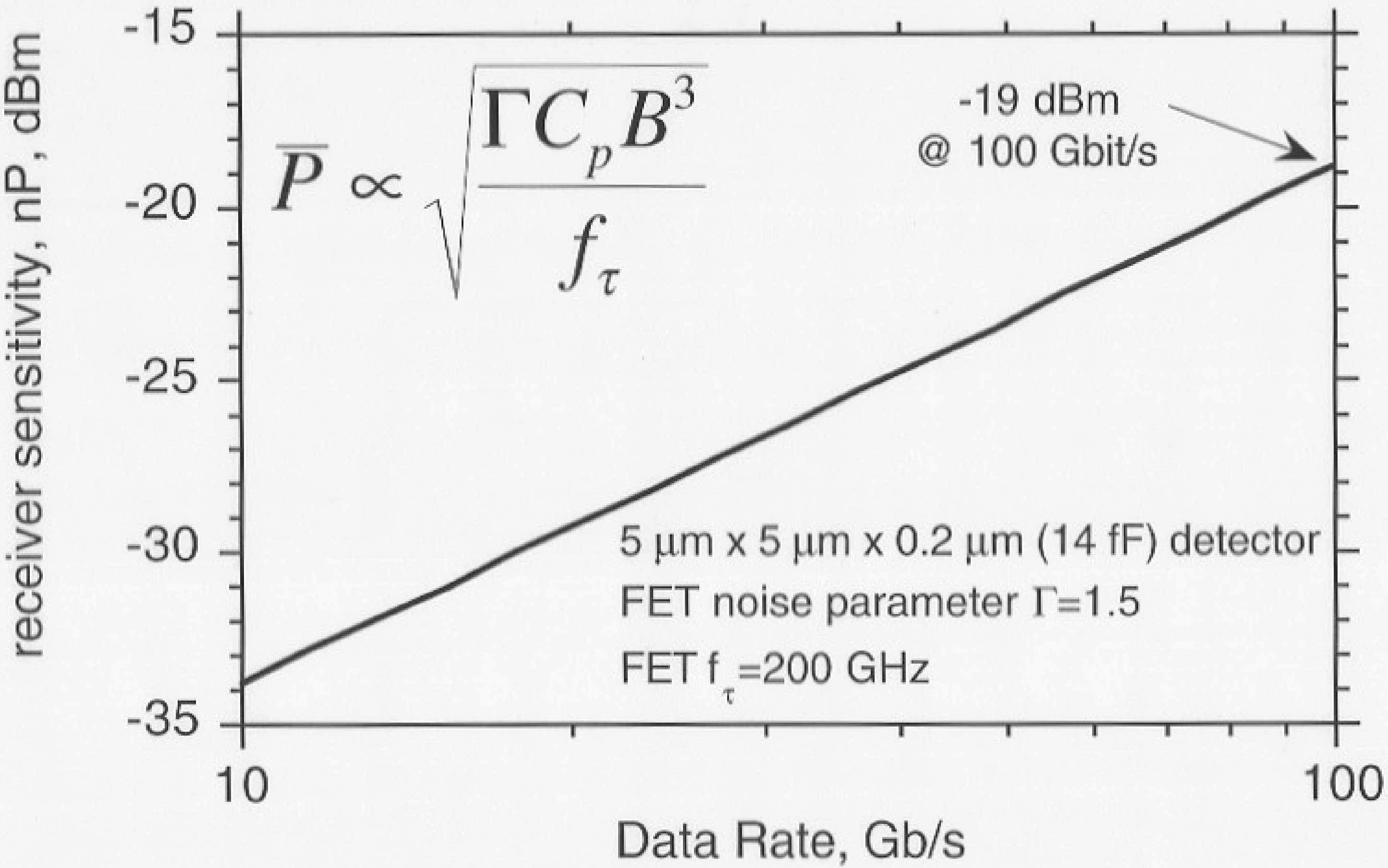
Maximize HEMT f_t

Minimize detector and layout capacitance

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

50Ω load would incur large noise penalty

Good transistors will make more sensitive receivers



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

$$\overline{P} \cong 6\frac{h\nu}{\eta q} \bullet \sqrt{\frac{32\pi kT\Gamma C_p B^3 I_3}{f_\tau} + \frac{4kTBI_2}{R_f}}$$

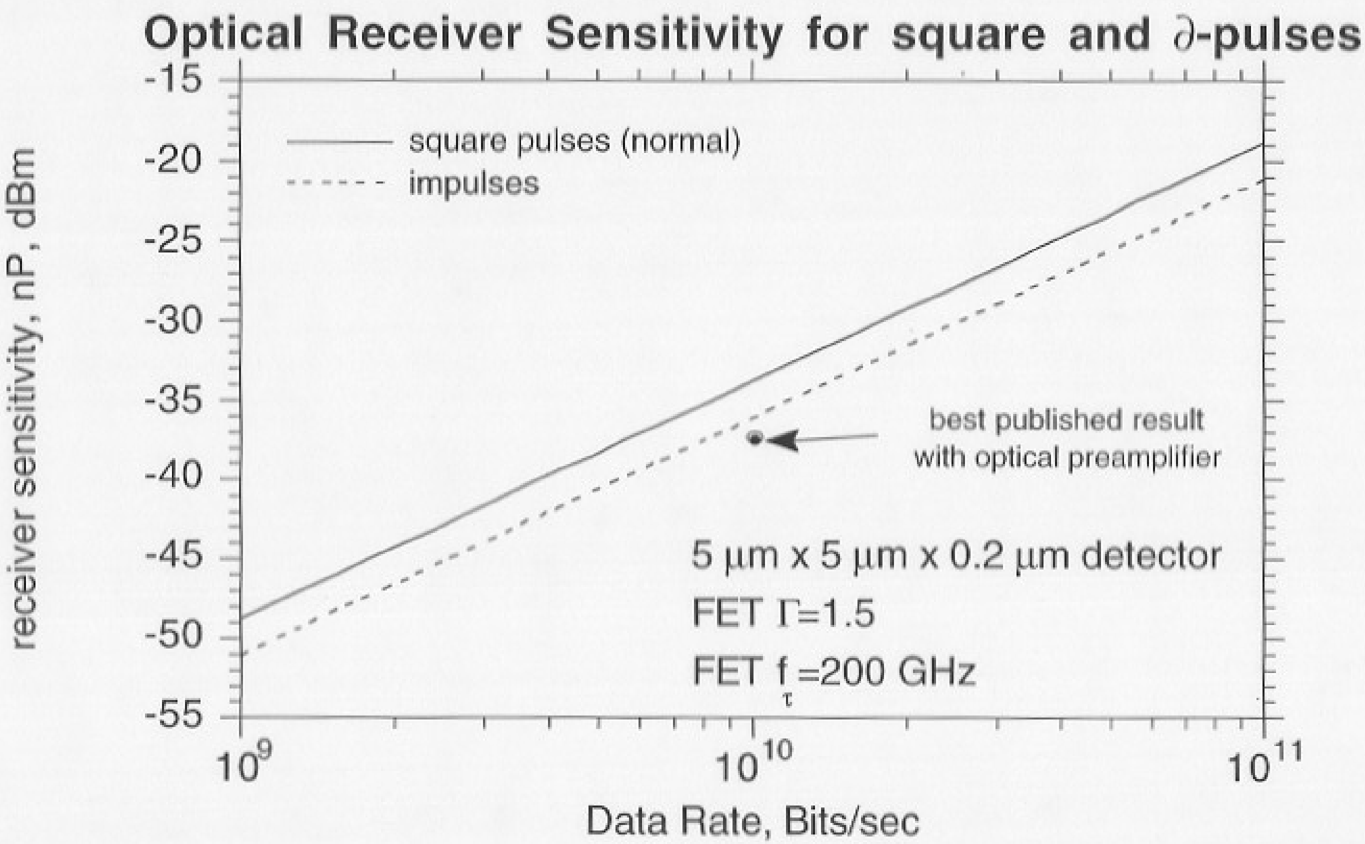
B is the bit rate. For a square optical pulse and a raised-cosine eye, I_2=0.68 and I_3=0.12. Sending impulses (solitons) will make I_3 somewhat smaller.

$$I_2 = [1/BH(0)] \cdot \int H(f)df$$

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

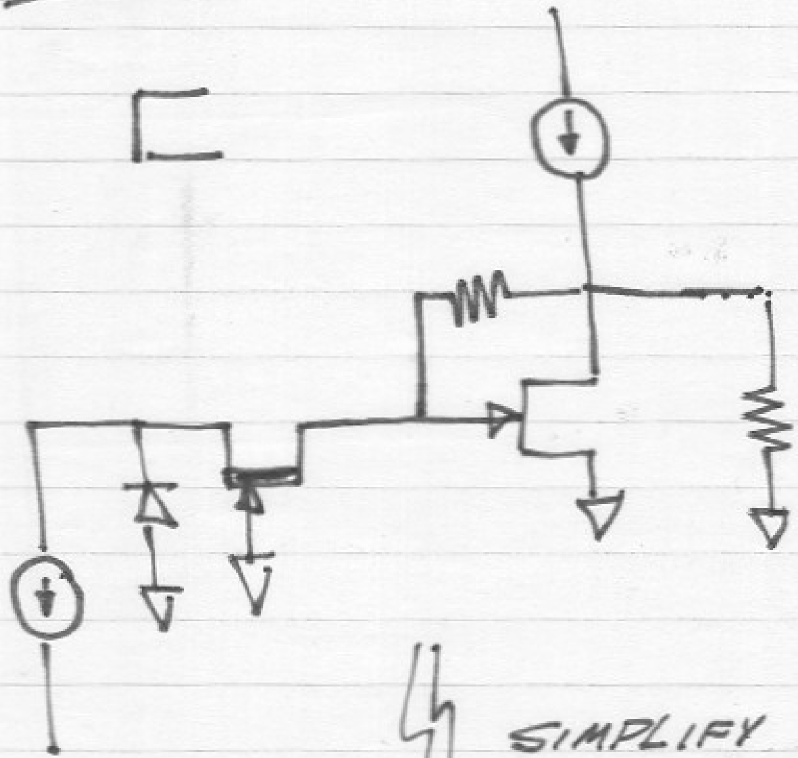
Note at 100 GB/sec, a 5 μmx5μmx0.2μm detector, f_t=200 GHz, Γ=1.5, that R_L=209Ω results in equal FET and resistor noise.

At 40 GB/sec, the point of equal contribution is R_L=1.3 kΩ.

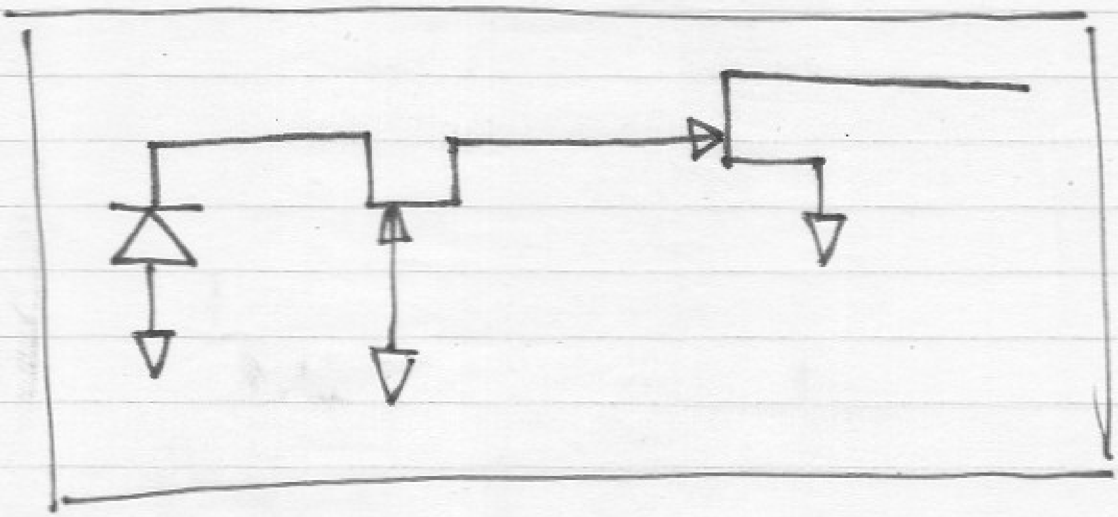


Optical Receiver Sensitivity for square and ∂-pulses
receiver sensitivity, nP, dBm
-15
-20
-25
-30
-35
-40
-45
-50
-55
square pulses (normal)
impulses
best published result
with optical preamplifier
5 µm x 5 µm x 0.2 µm detector
FET Γ=1.5
FET f_τ=200 GHz
10^9
10^{10}
10^{11}
Data Rate, Bits/sec

Noise Analysis



SIMPLIFY for now
to focus on key
issues:



More Notes on C.G. opt. rcvr:

FET 2 Noise, input-referred:* ($R_f \to \infty$)

$$\frac{4kT\Gamma}{g_{m2}}(2\pi f)^2(C_{gs2}+C_{gd1})^2 \cdot \left(1+\left(\frac{2f}{f_T}\right)^2\right)^* = \left.\frac{d\langle II^*\rangle}{df}\right|_{Fet2} = A$$

← ignore for now

Fet 1 Noise, input-referred.

$$\frac{4kT\Gamma}{g_{m1}}(2\pi f)^2(C_{gs1}+C_p)^2 = \left.\frac{d\langle II^*\rangle}{df}\right|_{Fet1} = B$$

$$\left[\frac{4kT\Gamma}{g_{m1}}(2\pi f)^2 C_{gs1}^2 \cdot\right] 4 =$$

because $C_{gs1}=C_p$ minimizes B

Examine the ratio of second to first-stage noise; assuming $C_{gs2} \gg C_{gd1}$

$$\frac{A}{B} = \left(\frac{C_{gs2}}{C_{gs1}}\right)^2 \frac{1}{4} \cdot \frac{g_{m1}}{g_{m2}} = \frac{1}{4}\left(\frac{W_{g2}}{W_{g1}}\right)$$

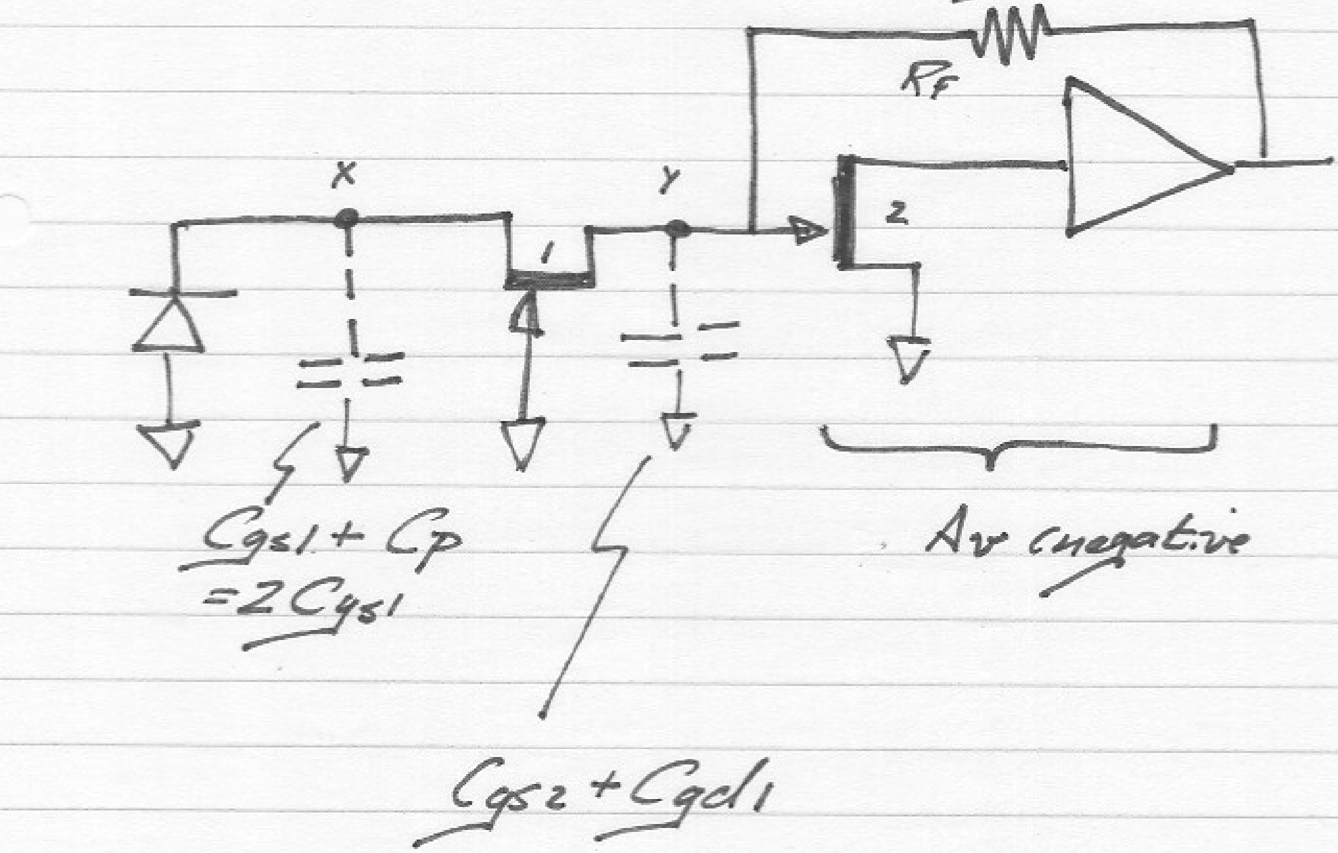
Note that $W_{g2} \leq W_{g1}$ will ensure

negligible FET2 noise contribution.

*(drop for now - but check the effect later)

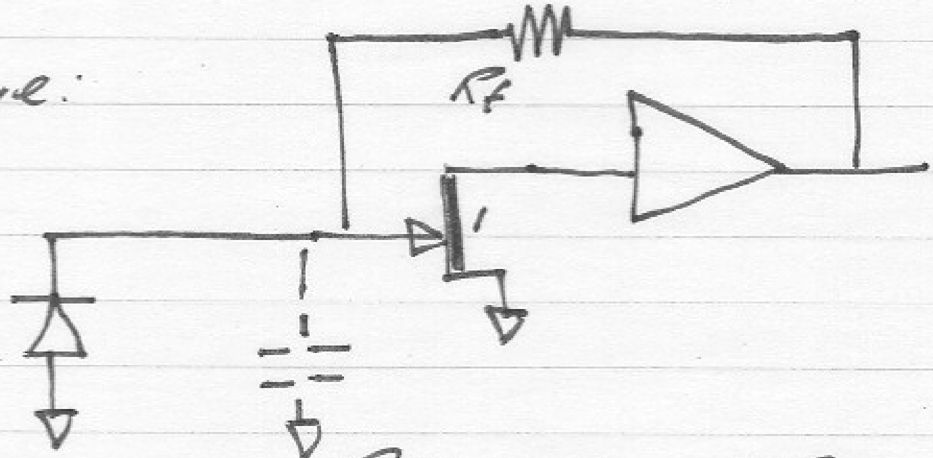
So, is that the end of the design effort?

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



the pole at "x" is at $f_T/2$; while at "y" it is at $f_{p2} = (1-A_v)/2\pi R_f(C_{gs2}+C_{gd1})$

Compare now to the simple transimpedance stage:



$$C_{gs1} + C_p = 2C_{gs1} = 2C_p$$

For equal values of R_f, thus equal noise penalties from R_f, the trans-Z has larger capacitance, (hence lower b.w.) by the ratio

if $C_{gd1} \ll C_{gs2}$

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

If we make $W_{g2}/W_{g1} = 1/2$,

we get four times the bandwidth with the cg stage added.

So, it would seem that a reasonable workable solution would be to make Fet 2 have 1/2 the gate width of Fet #1. We will find that the noise penalty of Fet 2 is only 1/8 (i.e. $\langle I_n^2 \rangle$, hence ~ 1/16 in $\sqrt{\langle I_n^2 \rangle}$!) of the noise of Fet 1, and that for equal transimpedance-stage ~~#~~ bandwidths, that the Resistor Rf ~~can be~~ is four times larger.

Why not simply make Q2 much smaller than Q1, given that an "exact" analysis gives minimum Q2 noise, and bigger stage-2 bandwidth, if Q2 is very small - in fact equal to $W_{g1} \times \frac{C_{gd1}}{C_{gs1}}$?

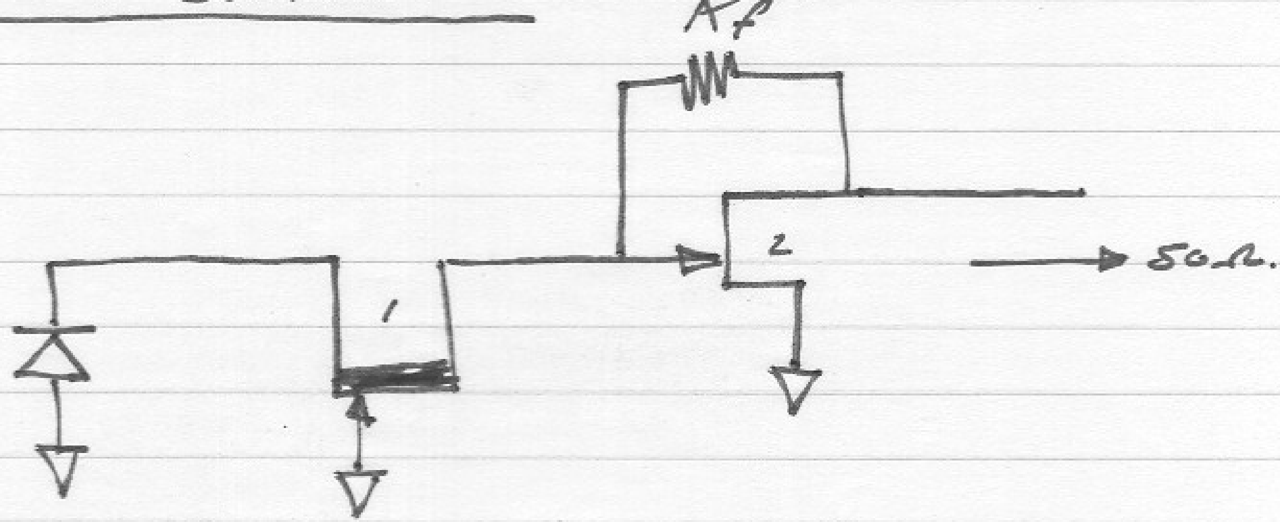
Because:

1) We can't make a Fet that small
   $(W_{g2} = W_{g1} \times C_{gd1} / C_{gs1})$

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: $C_{gs1} = C_p$ W_{g1} Q2: $W_{g2} = W_{g1}/2$

Pick R_f so that

$$\frac{4kT}{R_f}\,\mathbb{B}\,I_2 = \left(\frac{1}{2}\right) \times \frac{32\pi kT\Gamma C_p}{f_T}\,\mathbb{B}^3 I_3$$

the factor of 1/2 being negotiable ...

Note that $I_2 = 0.68$ $I_3 = 0.12$

Now a $5\mu m \times 5\mu m \times 0.2\mu m$ detector has

$C_p = \underline{14\,fF}$, but we'll use $25\,fF$ to allow for

stray C.

$C_p = 25\,fF$ $\qquad f_T = 100\,GHz$

$C_{gs1} = 25\,fF \rightarrow \quad g_{m1} = 2\pi C_{gs1} f_T = 15.7\,ms = 1/64\,\Omega$

$C_{gs2} = 12.5\,fF, \quad g_{m2} = g_{m1}/2 = 1/128\,\Omega$

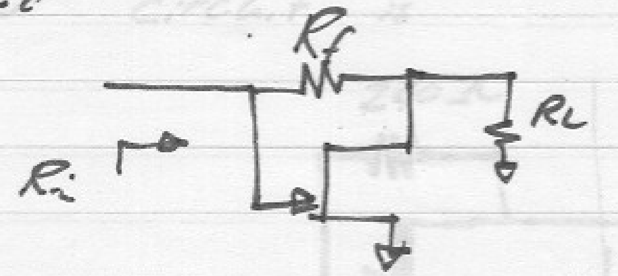
$$\frac{32\pi kT \Gamma C_p IB^3 I_3}{f_T} = 1.87(10^{-11})\,A^2 \quad @\ 100\,Gb/s.$$

→ allowing the noise contribution of R_f to be 50% of

this gives us $R_f = \begin{cases} 120\,\Omega & (B = 100\,Gb/s) \\ 750\,\Omega & (B = 40\,Gb/sec) \end{cases}$

These are minimum values.

Recall that



$$R_{in} = \frac{R_f + R_L}{1 + g_m R_L} \quad ; \quad -Z_T = R_f \frac{g_m R_L}{1 + g_m R_L} - R_L \frac{1}{1 + g_m R_L}$$

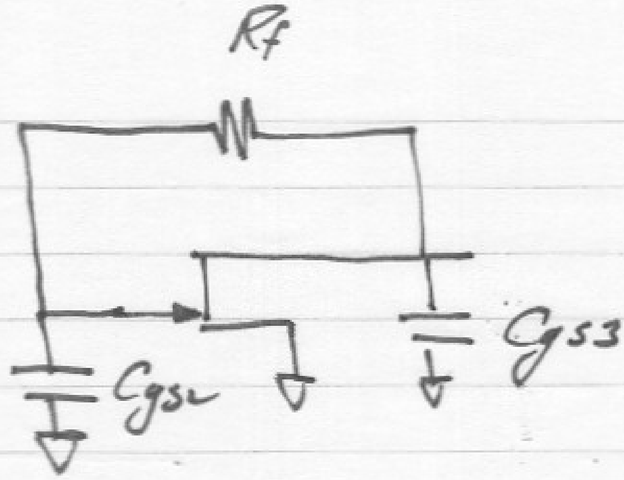
100 GB/sec. choose $R_L = \infty$!
& choose $R_f = 240\Omega$!

$$Z_T = R_f - \frac{1}{g_{m2}} = 240\Omega - 128\Omega = 112\Omega$$

$$R_{in} = \frac{1}{g_{m2}} = 128\Omega$$

second pole position: (neglecting C_{gd2})

$$f_{p2} = \frac{1}{2\pi(128\Omega)C_{gs2}} = f_T = 100\text{ GHz}!$$



$a_1 = (C_{gs3} + C_{gs2}) / g_{m2}$

$a_2 = C_{gs3} C_{gs2} R_f / g_{m2}$

$$\frac{a_2}{a_1} = \frac{C_{gs3} C_{gs2}}{C_{gs3} + C_{gs2}} R_f \sim C_{gs2} R_f$$

→ underdamped response will be observed

unless ($\approx$) $g_{m2} R_f < 1$

→ trouble.

Optimum Noise Design of Multi-Gigabit Common-Gate Optical Receivers

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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_v and R_d form an inverting amplifier with gain $A = -g_{m1} R_d A_v$. 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

$$\frac{d\langle I_n I_n^* \rangle}{df} = \frac{4kT\Gamma}{g_{m1}} (2\pi f)^2 (C_p + C_{gs1})^2 + 4kT / R_f \tag{1}$$

where C_p is the photodiode capacitance, f_τ is the HEMT current-gain cutoff frequency and Γ, its channel noise coefficient. Using the optimum HEMT size (that which gives $C_p = C_{gs}$), this becomes

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

At the gate of Q1, the node impedance is $R_f / (1-A)$ and the node capacitance is $C_p + C_{gs1} = 2C_p$. Hence, the preamplifier 3 dB bandwidth is given by

$$f_{3dB} = (1-A) / 4\pi C_p R_f \qquad (3)$$

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_{3dB}. Both, very high gain and wide bandwidth is difficult to attain, given finite transistor cut-off frequency f_τ. 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 $1/g_{m1}$ and the node capacitance is $C_p + C_{gs1} = 2C_{gs1}$. Hence the input pole frequency is $f_\tau / 2$. The dominant pole is at the gate of Q2, where the node impedance is $R_f / (1-A)$ and the node capacitance is $C_{gd1} + C_{gs2}$.

$$f_{3dB} = (1-A) / 2\pi (C_{gd1} + C_{gs2}) R_f \qquad (4)$$

As before, taking only the dominant noise terms, the input referred equivalent noise current spectral density is given by

$$\frac{d\langle I_n I_n^* \rangle}{df} = \frac{4kT\Gamma}{g_{m1}}(2\pi f)^2 (C_p + C_{gs1})^2$$
$$+\frac{4kT}{R_f}(1+(2\pi f)^2 (C_p + C_{gs1})^2 / g_{m1}^2)$$
$$+\frac{4kT\Gamma}{g_{m2}}(1/R_f^2 + (2\pi f)^2 (C_{gd1} + C_{gs2})^2) \tag{5}$$

The optimum gate widths of Q1 and Q2 are those which give $C_{gs1} = C_p$ and $C_{gd1} = C_{gs2}$ simplifying (5) to

$$\frac{d\langle I_n I_n^* \rangle}{df} \cong 32\pi kT\Gamma f^2 C_p / f_\tau + \frac{4kT}{R_f}(1+(2f/f_\tau)^2) + \frac{4kT\Gamma}{g_{m2} R_f^2} \tag{6}$$

where the approximation $C_{gs} >> C_{gd}$, 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_{gd1}) 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_{gd1}. 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/InP 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 Ω 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

[2] Y. Akahori, M. Ikeda, A. Kohzen and Y. Akatsu, "11 GHz ultrawide-bandwidth monolithic photoreceiver using InGaAs *pin* PD and InAlAs/InGaAs HEMTs," Elect. Lett., vol. 30, no. 3, p. 267, 1994

[3] Moon S. Park and Robert A. Minasian, "Ultralow Noise 10 Gb/s p-i-n-HEMT Optical Receiver," IEEE Phot. Tech. Lett., vol.5, no. 2, p. 161, 1993

[4] N. Ohkawa, "20 GHz bandwidth low-noise HEMT preamplifier for optical receivers," Elect. Lett., vol. 24, no. 17,p. 1061, 1988

[5] James L. Gimlett, "Ultrawide Bandwidth Optical Receivers," J. of Lightwave Tech., vol. 7, no. 10, p.1432, 1989

[6] I. Z. Darwazeh, P. M. R. S. Moreira, P. M. Lane and J. J. O'Reilly, "A Low Input Impedance Multi-Gbit GaAs Optical Receiver Tolerant to Photodiode Parasitics," presented at IEEE Intl' symposium on MMICs for comm. systems," London, Sept. 1992

[7] R. G. Smith and S. D. Personick, "Receiver Design for Optical Fiber Communication Ststems,"

[8] Michael Schlechtweg, Werner Reinert, Paul J. Tasker, Ronald Bosch, Jurgen Braunstein, Axel Hulsmann and Klaus Kohler, "Design and Characterisation of High Performance 60 GHz Pseudomorphic MODFET LNAs in CPW-Technology Based on

Accurate S-Parameter and Noise Models," IEEE Trans. on Microwave Theory and Techniques, vol. 40, no. 12, p. 2445, 1992

[9] Yih-Guei Wey, Kirk S. Giboney, John E. Bowers and Mark J. W. Rodwell, "100 GHz Double Heterostructure GaInAs/InP p-i-n Photodiode," OSA Proceedings on Ultrafast Elect. and Optoelect., vol. 14, 1993

Figure Captions

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

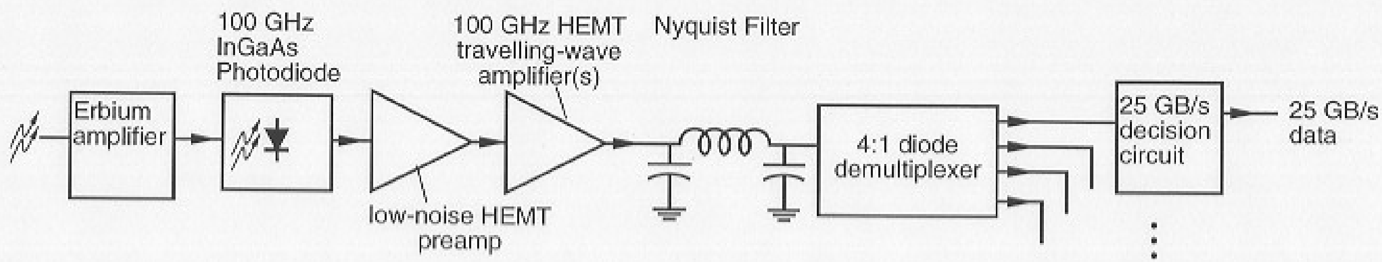
Fig. 2 Schematic diagram of common-gate/transimpedance photoreceiver

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

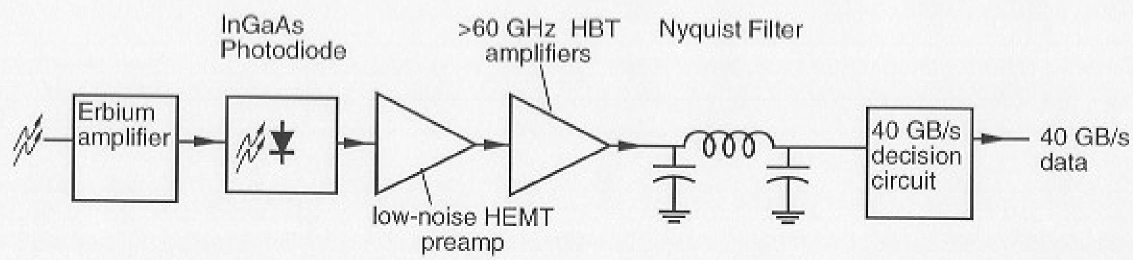
Fig. 4 Predicted sensitivity of common-gate/transimpedance preamplifier and integrating front-end using Personick coefficients I_2=0.68 and I_3=0.12

Receiver Implementation: In Summary

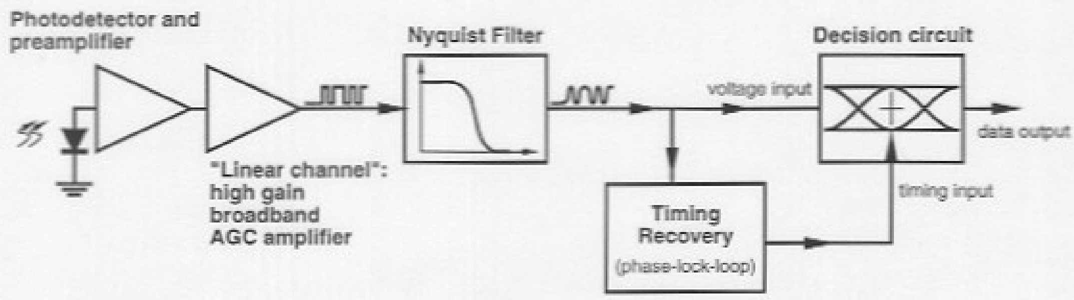
100 GB/sec Optical Receiver



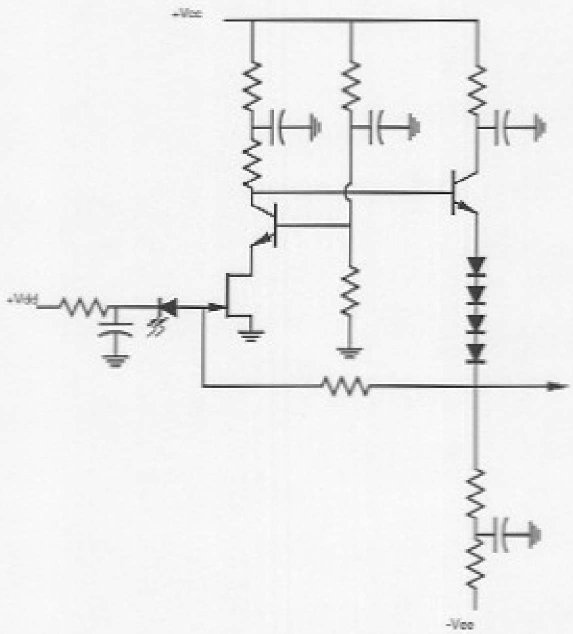
40 GB/sec Optical Receiver



Block diagm.

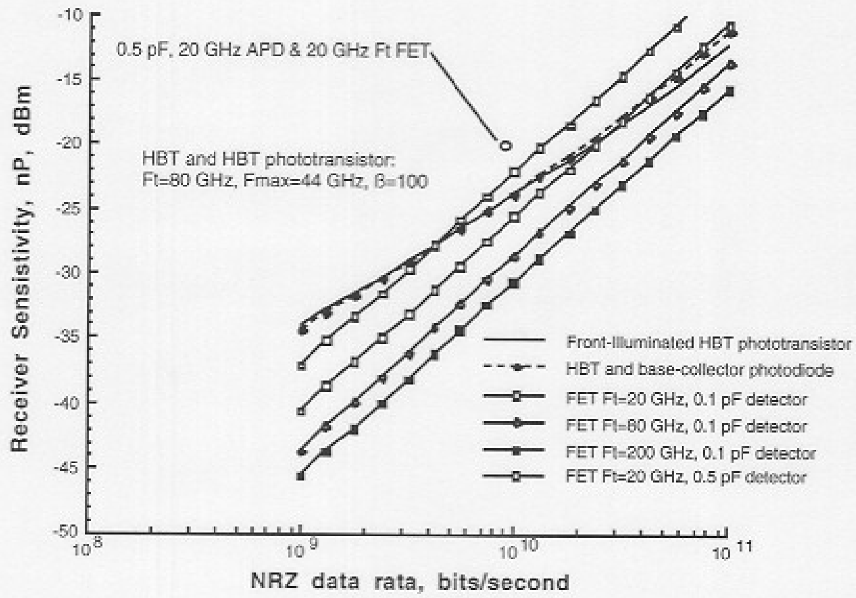


Photodetector and preamplifier
"Linear channel": high gain broadband AGC amplifier
Nyquist Filter
voltage input
Decision circuit
data output
timing input
Timing Recovery
(phase-lock-loop)



Front End.

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



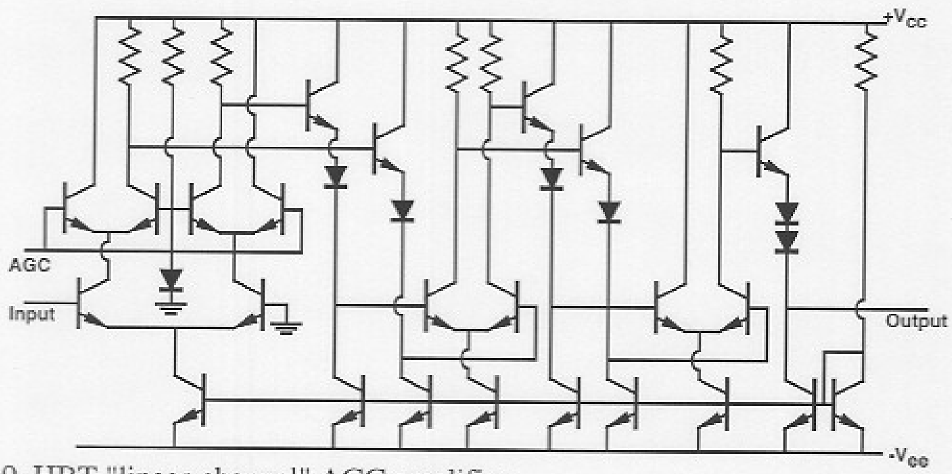
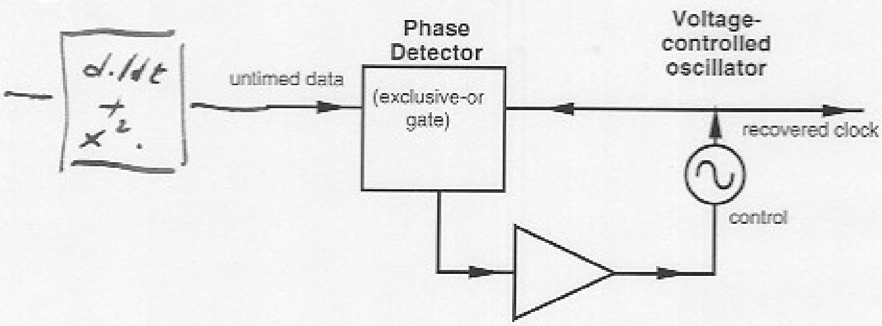
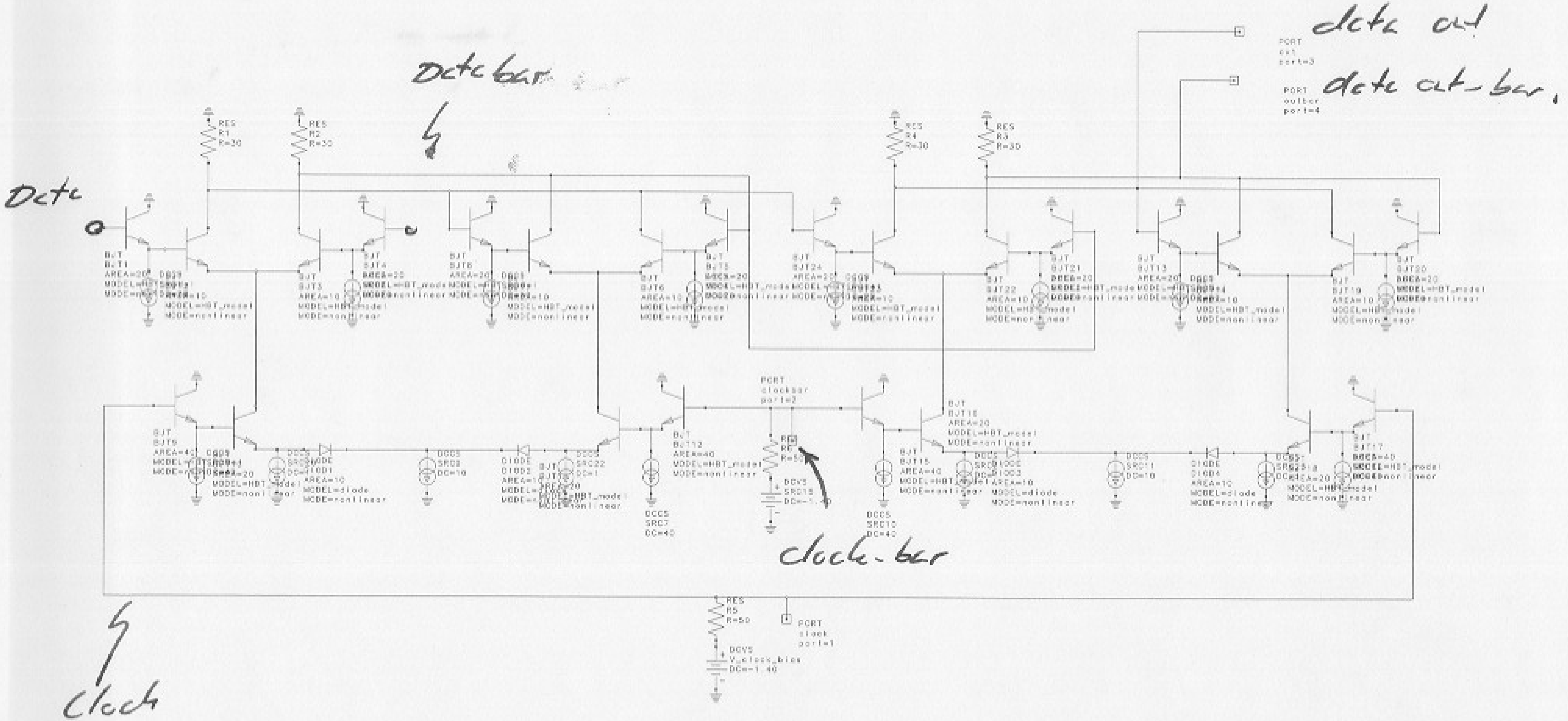


Figure 9. HBT "linear-channel" AGC amplifier

Phaselock loop.



decision circuit



AGC Amplifier

gain control.

outputs.

acg detector.

In