

ECEN620: Network Theory Broadband Circuit Design Fall 2024

Lecture 13: Transimpedance Amplifiers (TIAs)



Sam Palermo
Analog & Mixed-Signal Center
Texas A&M University

Announcements

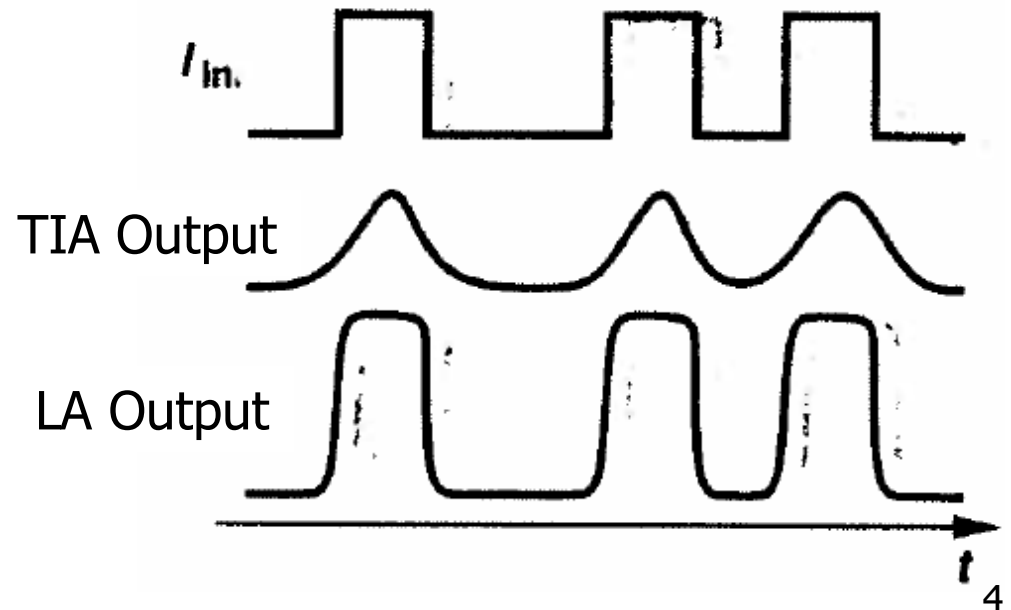
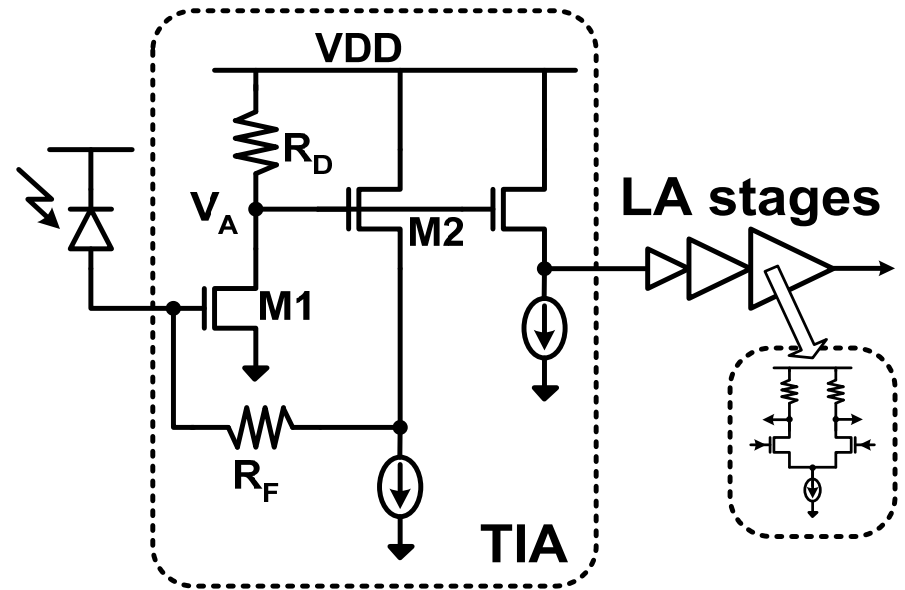
- Project Preliminary Report (HW5) due Nov 14

Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

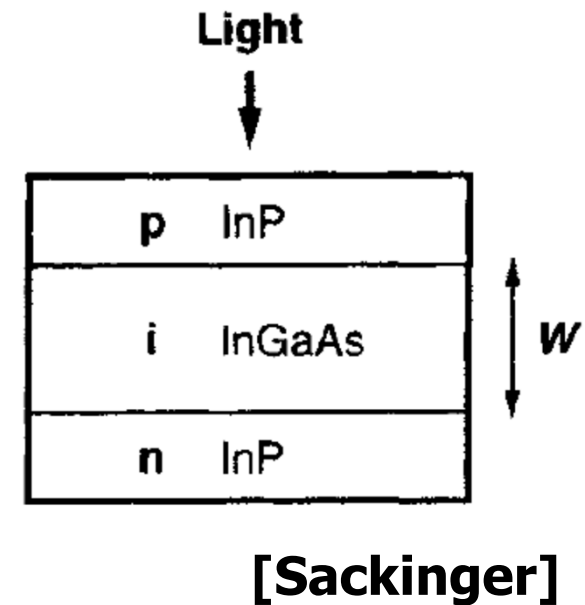
Optical Receiver Technology

- Photodetectors convert optical power into current
 - p-i-n photodiodes
 - Waveguide Ge photodetectors
- Electrical amplifiers then convert the photocurrent into a voltage signal
 - Transimpedance amplifiers
 - Limiting amplifiers
 - Integrating optical receiver



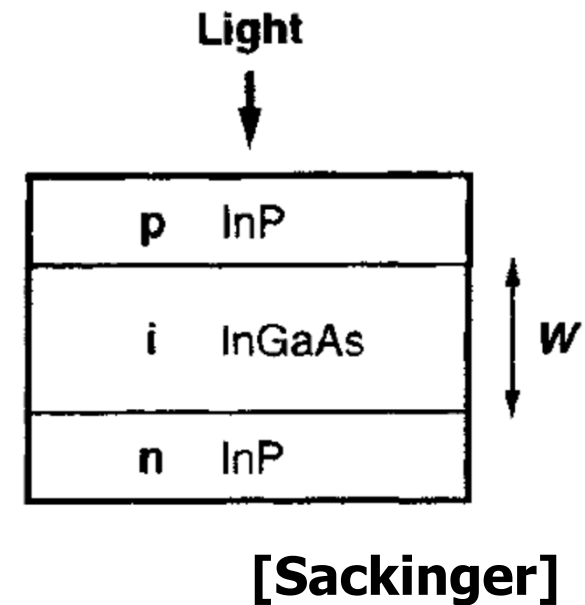
p-i-n Photodetector

- A p-i-n photodetector has an intrinsic layer (undoped or lightly doped) sandwiched between p- and n-doped material
- This p-n junction operates in reverse bias to create a strong electric field in the intrinsic region
- Normally incident photons create electron-hole pairs in the intrinsic region which are separated by the electric drift field, and photocurrent appears at the terminals



p-i-n Photodetector Tradeoffs

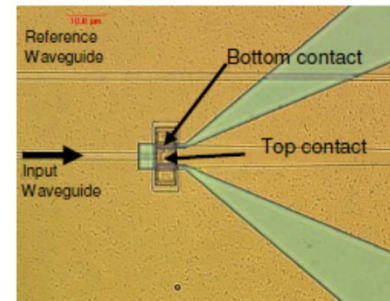
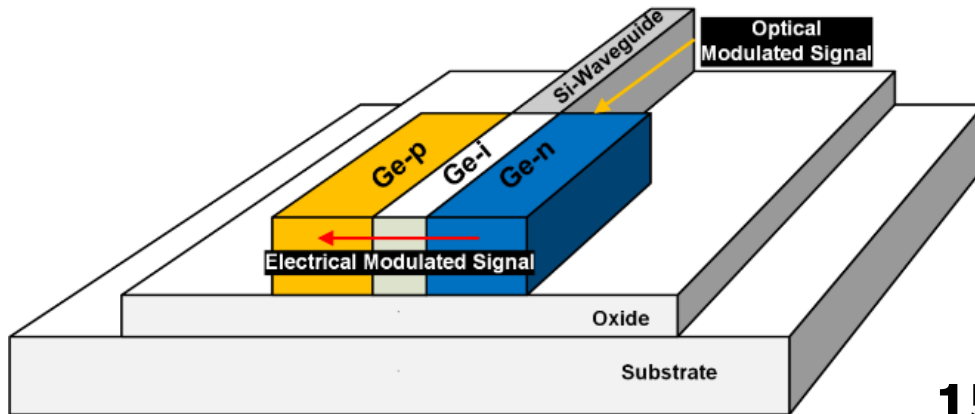
- There is a tradeoff between efficiency and speed set by the intrinsic layer width W
- The quantum efficiency η is the fraction of photons that create electron-hole pairs and is determined by W and detector's absorption coefficient α
- A wider W allows for higher η , but also results in longer carrier transit times which reduces the detector bandwidth



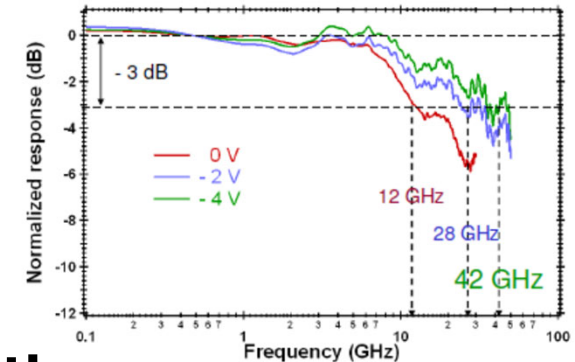
$$\eta \approx 1 - e^{-\alpha W}$$

Waveguide p-i-n Photodetector

[Vivien OptExp 2009]



15 μm detector length



- A waveguide p-i-n photodetector structure allows this efficiency-speed trade-off to be broken
- The light travels horizontally down the intrinsic region and the electric field is formed orthogonal
- Allows for both a thin i-region for short transit times and a sufficiently long i-region for high quantum efficiency

p-i-n Photodetector Responsivity

- The efficiency at which a photodetector converts optical power to electrical current is called **responsivity** R

- As each photon has energy of hc/λ ,

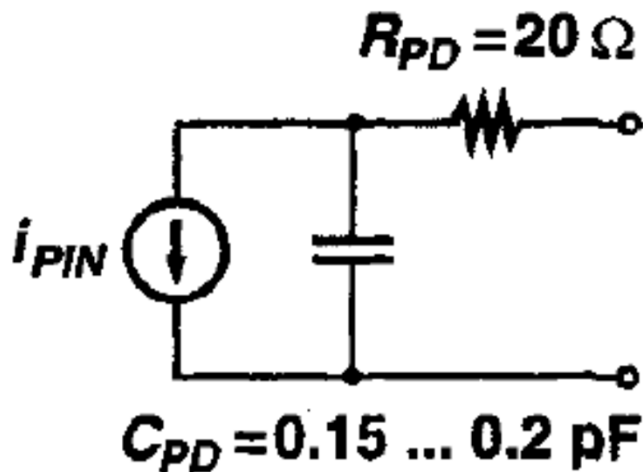
$$I_{PIN} = \eta \cdot \frac{\lambda q}{hc} \cdot P = R \cdot P$$

$$\text{where } R = \eta \cdot \frac{\lambda q}{hc} = (8 \times 10^5) \eta \lambda \text{ (A/W)}$$

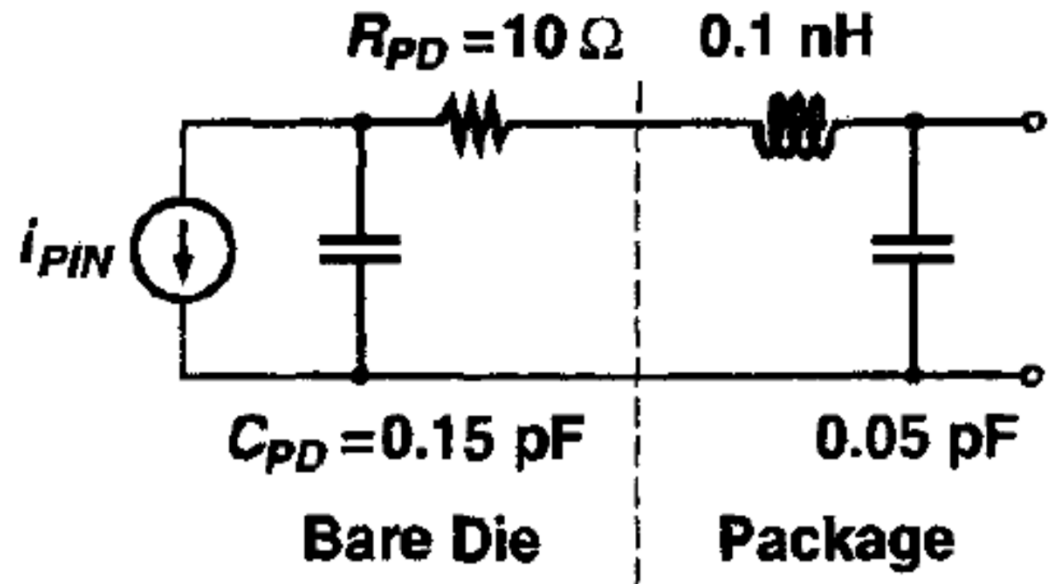
- Example: A photodetector with $\eta=0.6$ operating at 1310nm has $R=0.63$ A/W
- There is the potential for $R>1$, with $\eta=1$ and 1550nm operation $\Rightarrow R=1.24$

p-i-n Photodetector Equivalent AC Circuits

Bare Photodetector

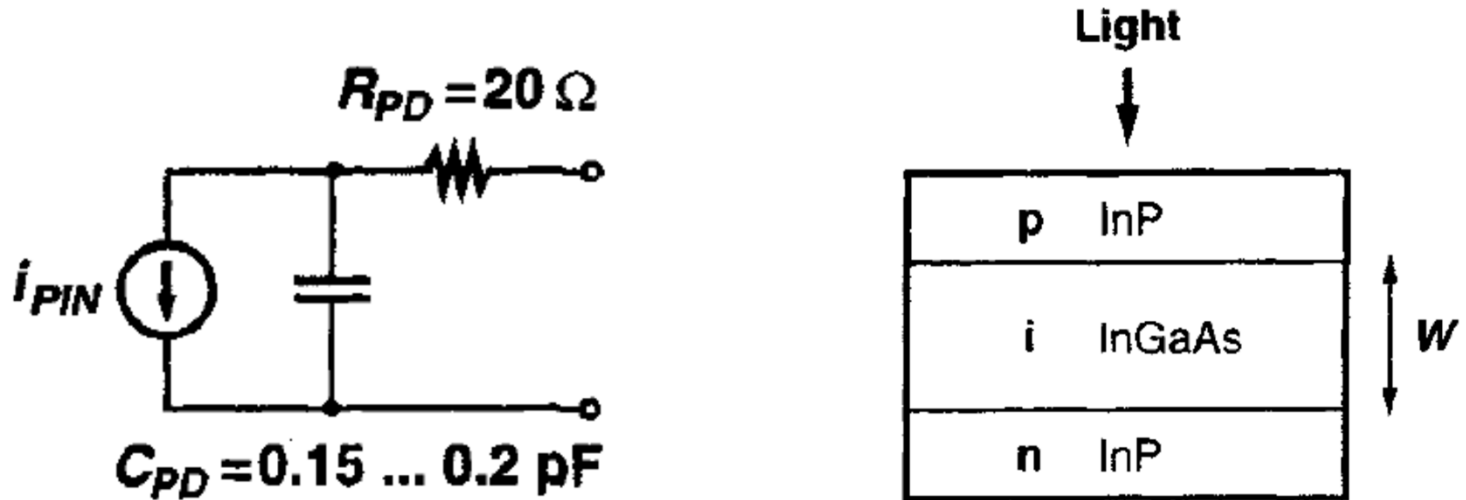


Photodetector + Packaging Parasitics



- The photodetectors main parasitics are the junction capacitance C_{PD} and contact/spreading resistance R_{PD}
- Additional LC parasitics are present in packaged devices due to wirebonds, etc...

p-i-n Photodetector Bandwidth



- Two time constants set the bare photodetector bandwidth

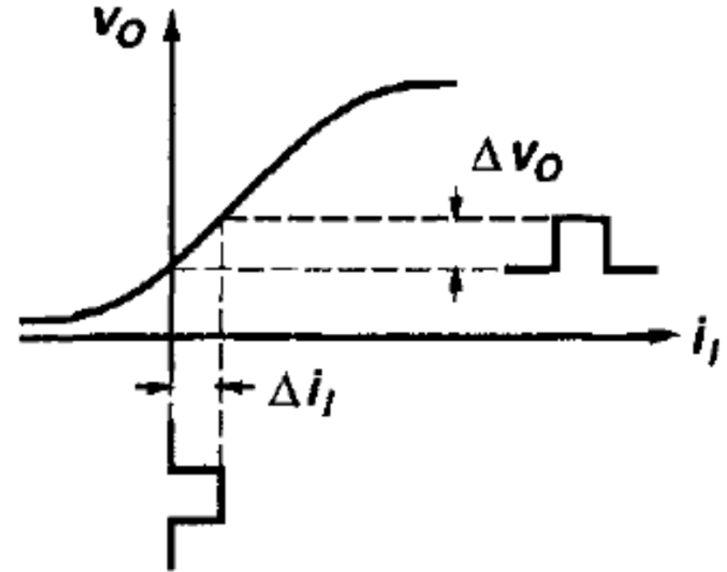
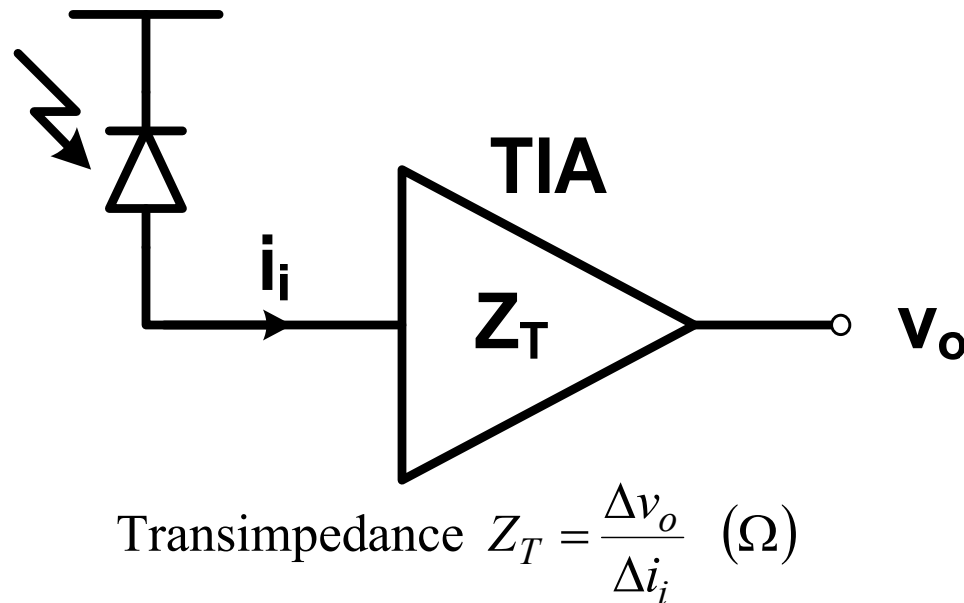
Transit Time: $\tau_{TR} = \frac{W}{v_n}$ where v_n is the carrier velocity

R - C Time Constant: $\tau_{RC} = R_{PD}C_{PD}$

The bandwidth in Hz can be approximated by

$$BW = \frac{1}{2\pi} \cdot \frac{1}{W/v_n + R_{PD}C_{PD}}$$

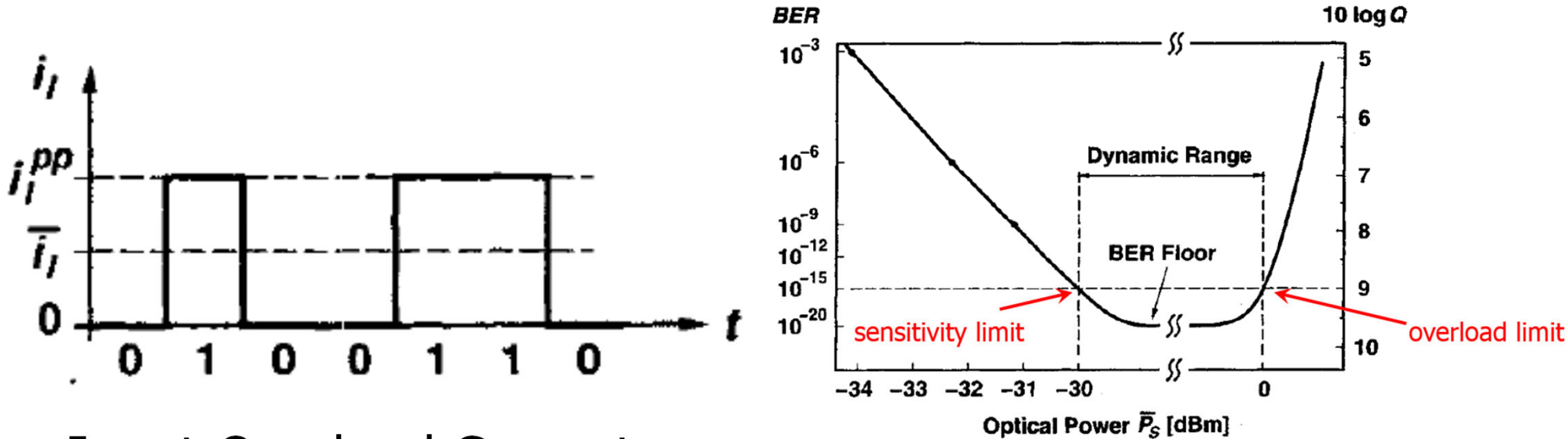
Transimpedance Amplifier (TIA)



Also expressed in units of $\text{dB}\Omega$ by $20\log(|Z_T|)$

- Key design objectives
 - High transimpedance gain
 - Low input resistance for high bandwidth and efficient gain
- For large input currents, the TIA gain can compress and pulse-width distortion/jitter can result

Maximum Currents

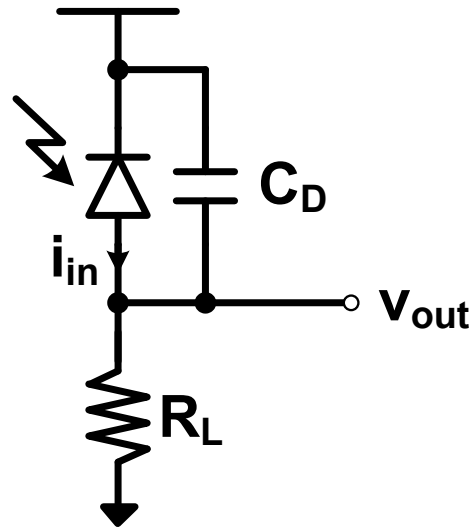


- Input Overload Current
 - The maximum peak-to-peak input current for which we can achieve the desired BER
 - Assuming high extinction ratio $i_{ovl}^{pp} = 2R\bar{P}_{ovl}$
- Maximum Input Current for Linear Operation
 - Often quantified by the current level for a certain gain compression (1dB)

$$i_{lin}^{pp} < i_{ovl}^{pp}$$

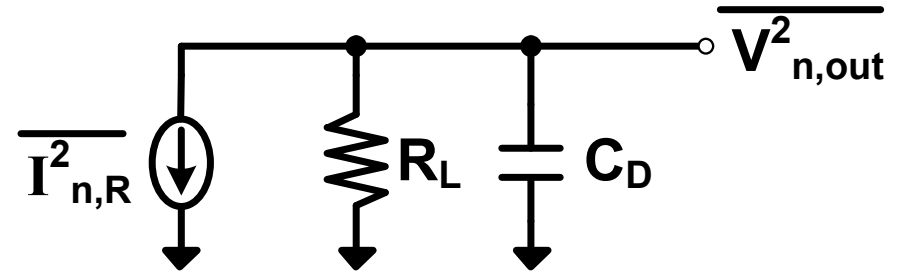
Resistive Front-End

[Razavi]



$$R_T = R_{in} = R_L$$

$$BW_{3dB} = \omega_p = \frac{1}{R_{in}C_D} = \frac{1}{R_LC_D}$$



$$\overline{V_{n,out}^2} = \int_0^\infty \overline{I_n^2} Z_T^2 df = \int_0^\infty \frac{4kT}{R_L} \left(\frac{R}{1 + j2\pi fRC} \right)^2 df = \frac{kT}{C_D}$$

$$\overline{I_{n,in}^2} = \frac{\overline{V_{n,out}^2}}{R_L^2} = \frac{kT}{R_L^2 C_D}$$

$$I_{n,in,rms} = \frac{\sqrt{KT/C_D}}{R_L}$$

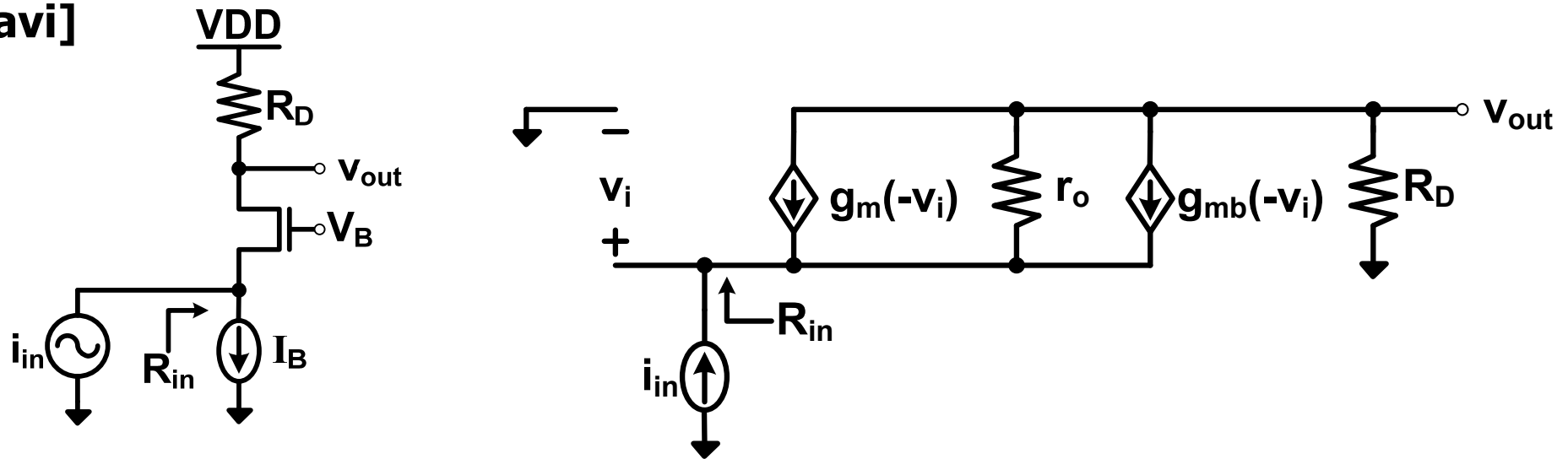
- Direct trade-offs between transimpedance, bandwidth, and noise performance

Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

Common-Gate TIA

[Razavi]



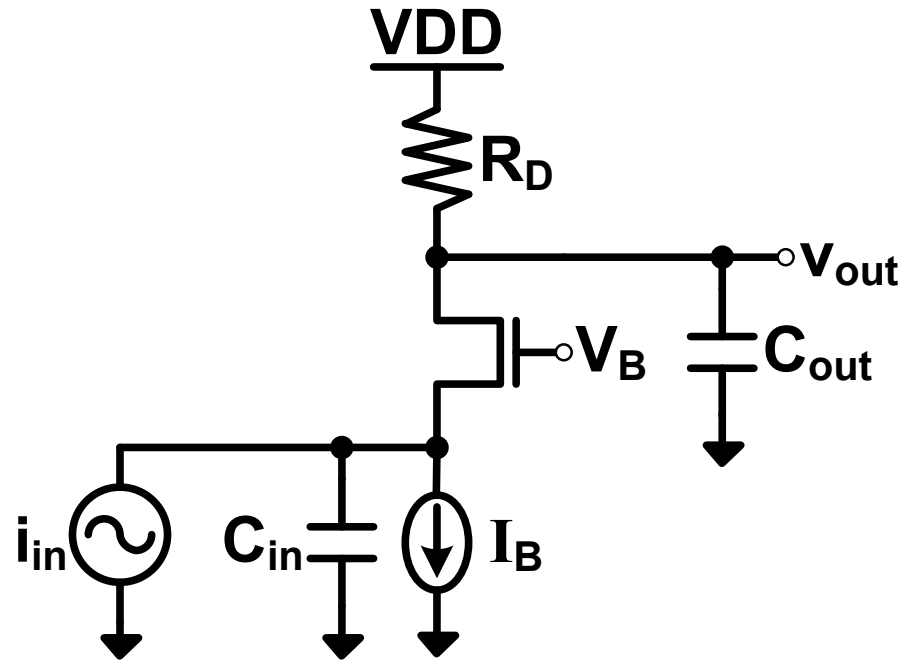
$$R_T = R_D$$

$$R_{in} = \frac{r_o + R_D}{1 + (g_m + g_{mb})r_o} \approx \frac{1}{g_m}$$

- Input resistance (input bandwidth) and transimpedance are decoupled

Common-Gate TIA Frequency Response

[Razavi]



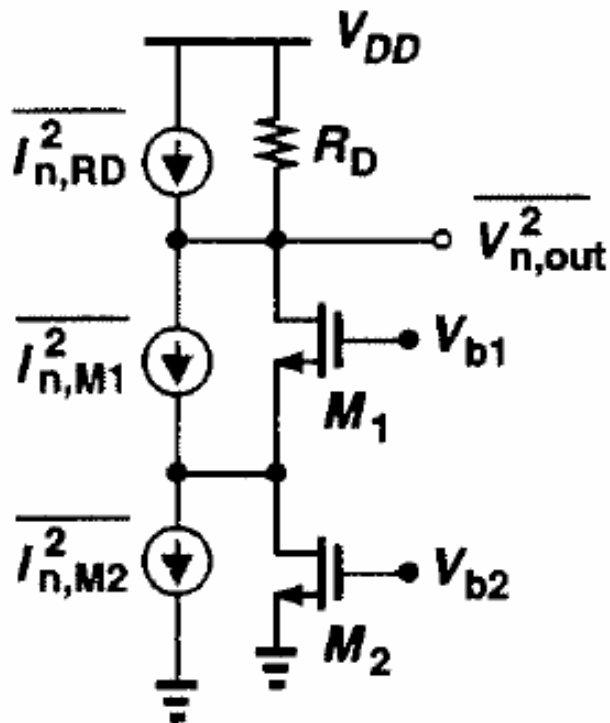
Neglecting transistor r_o :

$$\frac{v_{out}}{i_{in}} = \frac{R_D}{\left(1 + s \frac{C_{in}}{g_{m1} + g_{mb1}}\right)(1 + sR_D C_{out})}$$

- Often the input pole may dominate due to large photodiode capacitance (100 – 500fF)

Common-Gate TIA Noise

[Razavi]



Neglecting transistor r_o :

$$\overline{V_{n,out}^2} = \left(\overline{I_{n,M2}^2} + \overline{I_{n,RD}^2} \right) R_D^2 = 4kT \left(\frac{2}{3} g_{m2} + \frac{1}{R_D} \right) R_D^2 \quad \left(\frac{\text{V}^2}{\text{Hz}} \right)$$

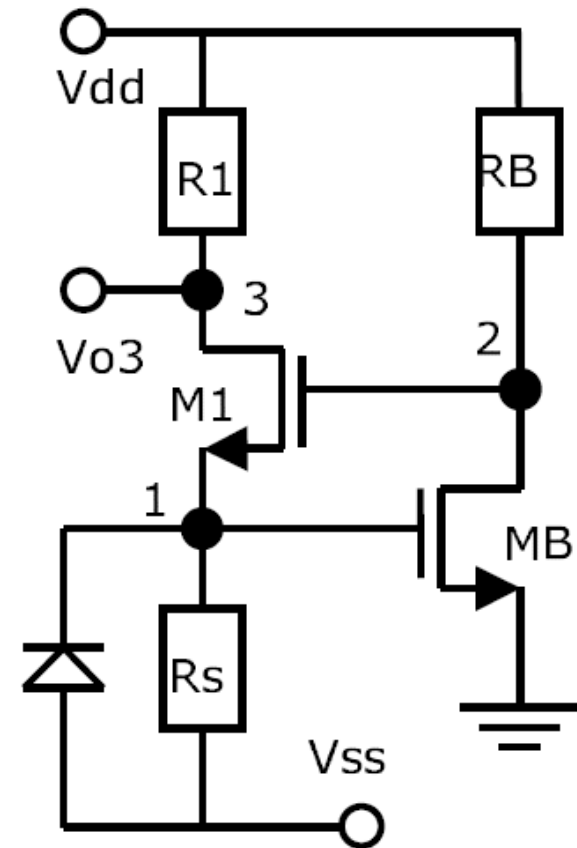
$$\overline{I_{n,in}^2} = 4kT \left(\frac{2}{3} g_{m2} + \frac{1}{R_D} \right) \quad \left(\frac{\text{A}^2}{\text{Hz}} \right)$$

- Both the bias current source and R_D contribute to the input noise current
- R_D can be increased to reduce noise, but voltage headroom can limit this
- Common-gate TIAs are generally not for low-noise applications
- However, they are relatively simple to design with high stability

Regulated Cascode (RGC) TIA

- Input transistor g_m is boosted by common-source amplifier gain, resulting in reduced input resistance
- Requires additional voltage headroom
- Increased input-referred noise from the common-source stage

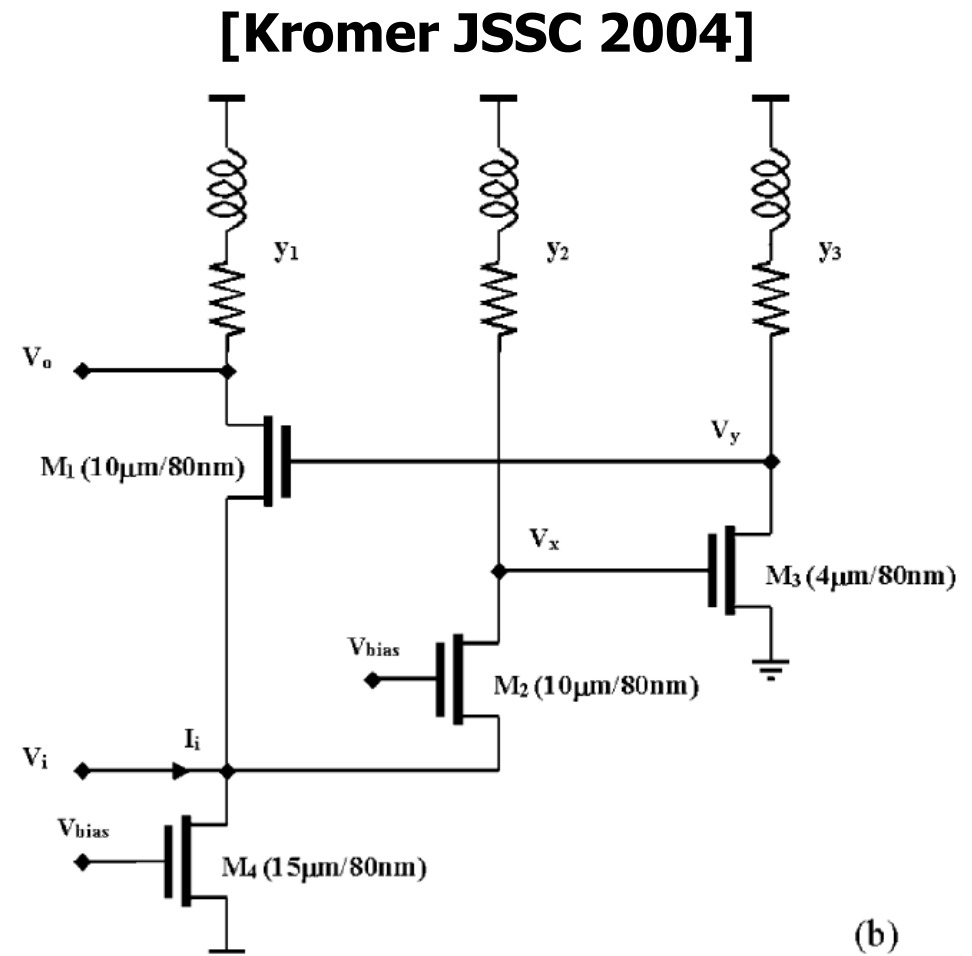
[Park ESSCIRC 2000]



$$Z_{in}(0) \cong \frac{1}{g_{m1}(1 + g_{mB}R_B)}$$

CMOS 20GHz TIA

- An additional common-gate stage in the feedback provides further gm-boosting and even lower input resistance
- Shunt-peaking inductors provide bandwidth extension at zero power cost, but very large area cost



$$Z_i \approx \frac{1}{g_{m1} (1 + |A_2 A_3|) + j\omega C_{i,\text{tot}}}$$

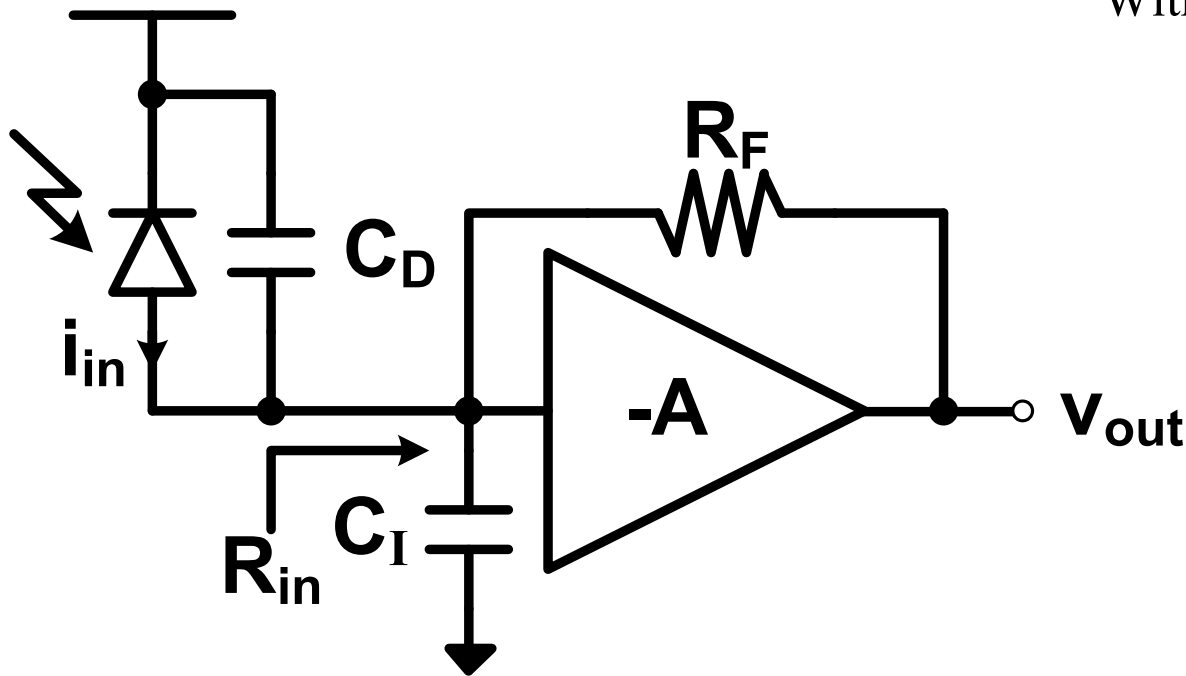
$$A_2 = g_{m2} R_2 \quad A_3 = -g_{m3} R_3$$

Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

Feedback TIA w/ Ideal Amplifier

With Infinite Bandwidth Amplifier:



$$Z_T(s) = -R_T \left(\frac{1}{1 + s/\omega_p} \right)$$

$$R_{in} = \frac{R_F}{A+1}$$

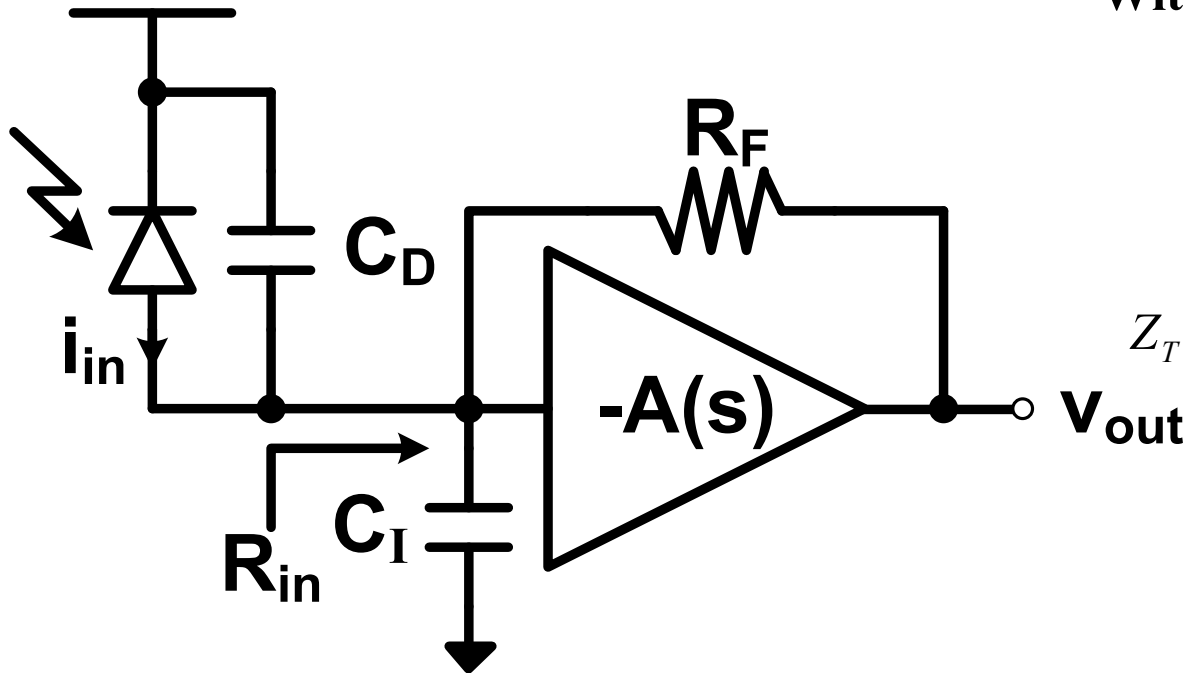
$$R_T = \frac{A}{A+1} R_F$$

$$\omega_p = \frac{1}{R_{in} C_T} = \frac{A+1}{R_F (C_D + C_I)}$$

- Input bandwidth is extended by the factor $A+1$
- Transimpedance is approximately R_F
- Can make R_F large without worrying about voltage headroom considerations

Feedback TIA w/ Finite Bandwidth Amplifier

With Finite Bandwidth Amplifier :



$$A(s) = \frac{A}{1 + \frac{s}{\omega_A}} = \frac{A}{1 + sT_A}$$

$$Z_T(s) = -R_T \left(\frac{1}{1 + s/(\omega_o Q) + s^2/\omega_o^2} \right)$$

$$R_T = \frac{A}{A+1} R_F$$

$$\omega_o = \sqrt{\frac{A+1}{R_F C_T T_A}}$$

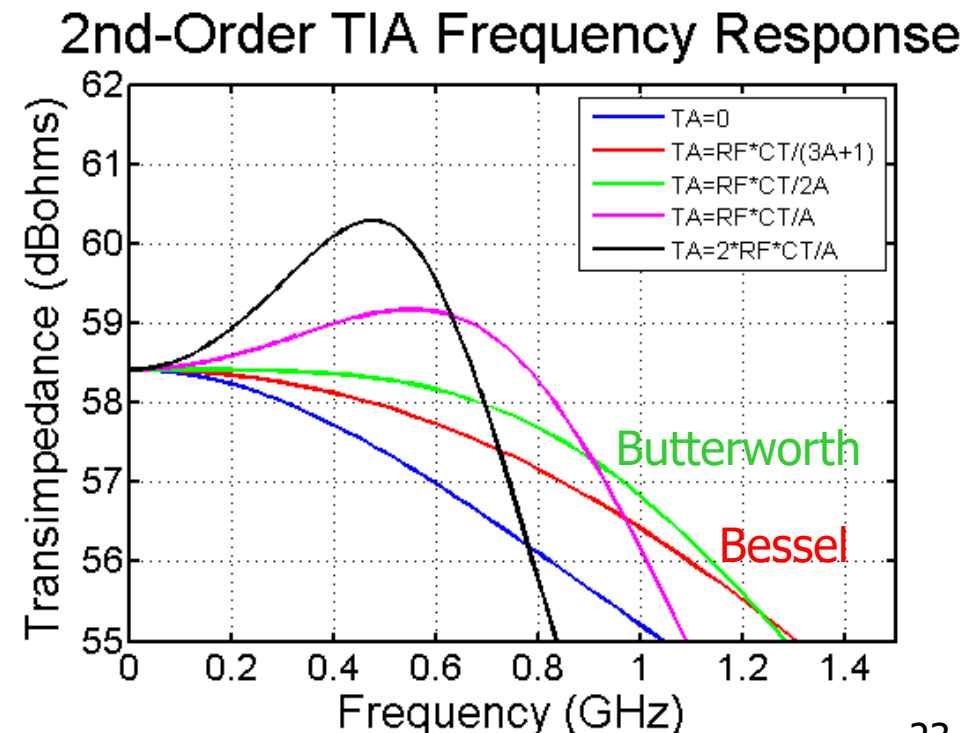
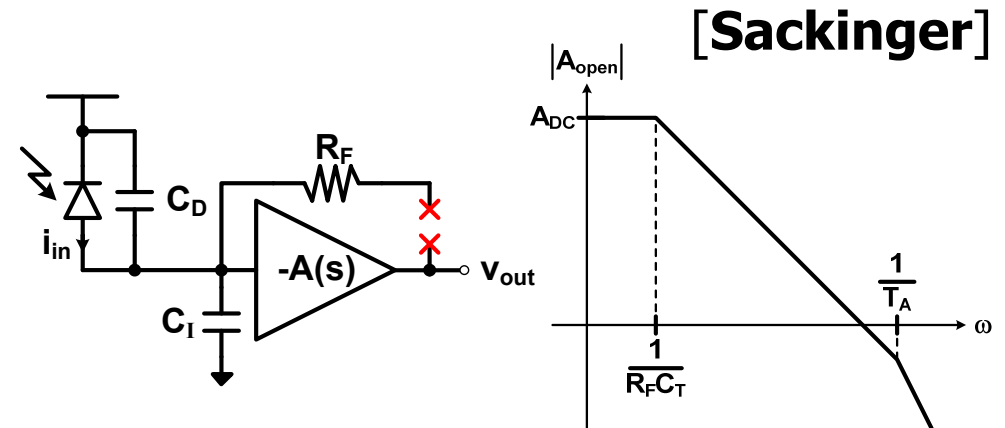
$$Q = \frac{\sqrt{(A+1)R_F C_T T_A}}{R_F C_T + T_A}$$

$$R_{in} = \frac{R_F}{A+1}$$

- Finite bandwidth amplifier modifies the transimpedance transfer function to a second-order low-pass function

Feedback TIA w/ Finite Bandwidth Amplifier

- Non-zero amplifier time constant can actually increase TIA bandwidth!!
- However, can result in peaking in frequency domain and overshoot/ringing in time domain
- Often either a Butterworth ($Q=1/\sqrt{2}$) or Bessel response ($Q=1/\sqrt{3}$) is used
 - Butterworth gives maximally flat frequency response
 - Bessel gives maximally flat group-delay



Feedback TIA Transimpedance Limit

If we assume a Butterworth response for maximally flat frequency response :

$$Q = \frac{1}{\sqrt{2}} \quad \Rightarrow \quad \omega_A = \frac{1}{T_A} = \frac{2A}{R_F C_T}$$

For a Butterworth response :

$$\omega_{3dB} = \omega_0 = \sqrt{\frac{(A+1)\omega_A}{R_F C_T}} = \frac{\sqrt{(A+1)2A}}{R_F C_T} \approx \sqrt{2} \text{ times larger than } T_A = 0 \text{ case of } \frac{A+1}{R_F C_T}$$

Plugging $R_T = \frac{A}{A+1} R_F$ into above expression yields the maximum possible R_T for a given bandwidth

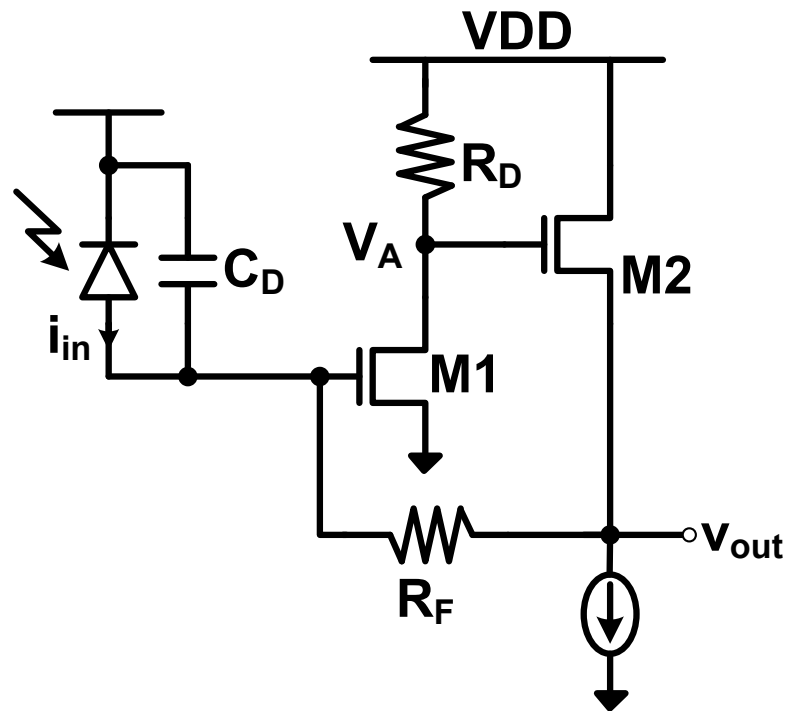
$$\sqrt{\frac{(A+1)\omega_A}{\left(\frac{A+1}{A}\right)R_T C_T}} \geq \omega_{3dB}$$

$$\boxed{\text{Maximum } R_T \leq \frac{A \omega_A}{C_T \omega_{3dB}^2}}$$

[Mohan JSSC 2000]

- Maximum R_T proportional to amp gain-bandwidth product
- If amp GBW is limited by technology f_T , then in order to increase bandwidth, R_T must decrease quadratically!

Feedback TIA



Assuming that the source follower has an ideal gain of 1

$$A = g_{m1} R_D$$

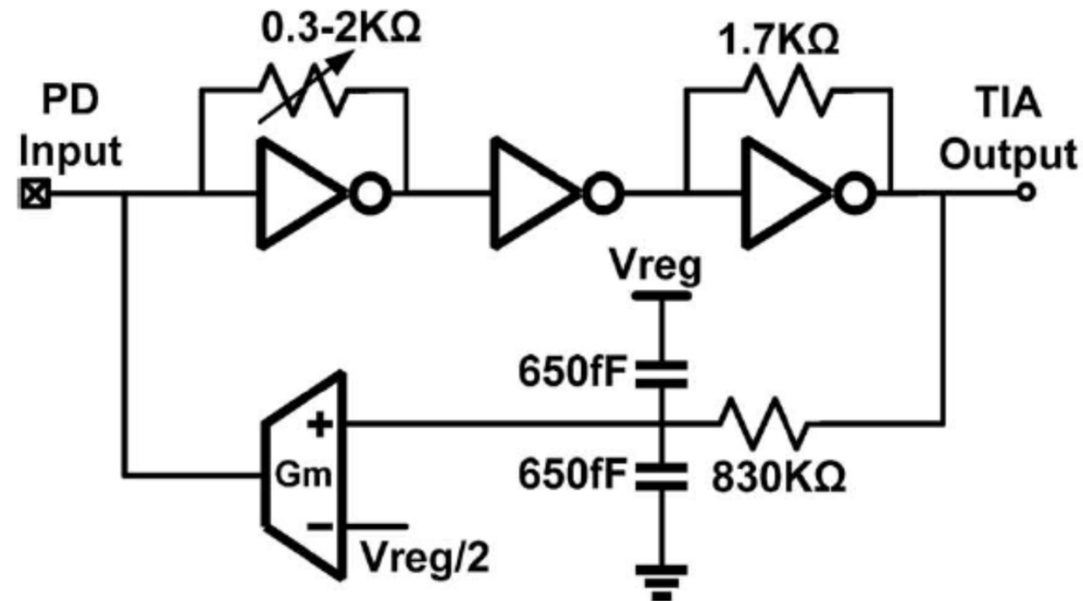
$$R_T = \frac{g_{m1} R_D}{1 + g_{m1} R_D} R_F$$

$$R_{in} = \frac{R_F}{1 + g_{m1} R_D}$$

$$R_{out} = \frac{1}{g_{m2} (1 + g_{m1} R_D)}$$

- As power supply voltages drop, there is not much headroom left for R_D and the amplifier gain degrades

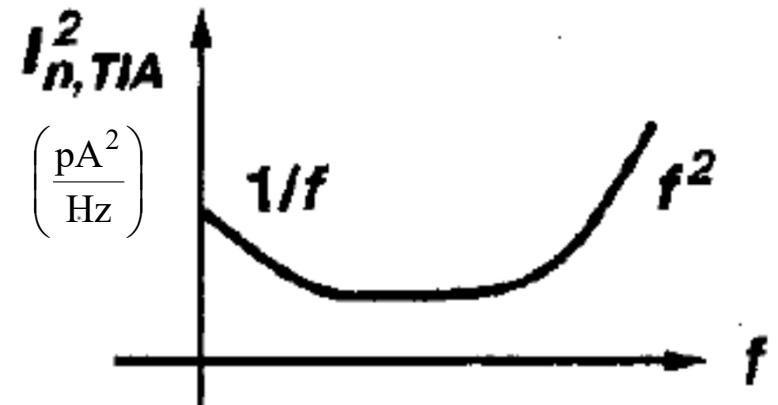
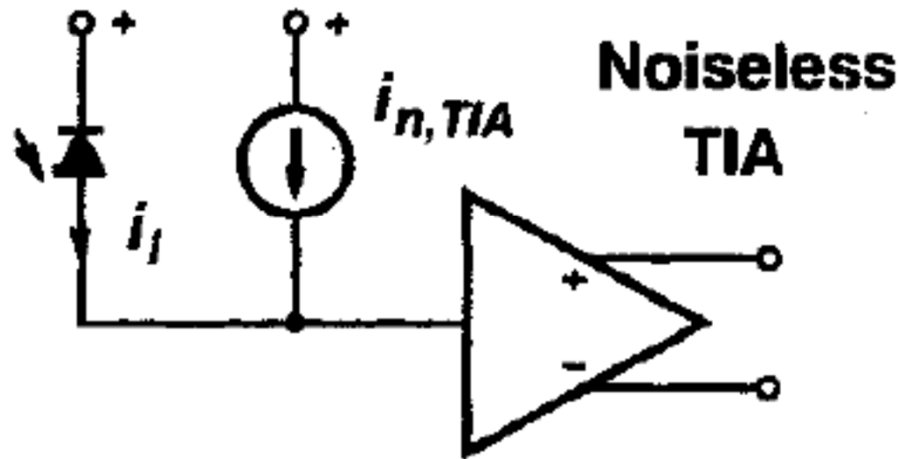
CMOS Inverter-Based Feedback TIA



[Li JSSC 2014]

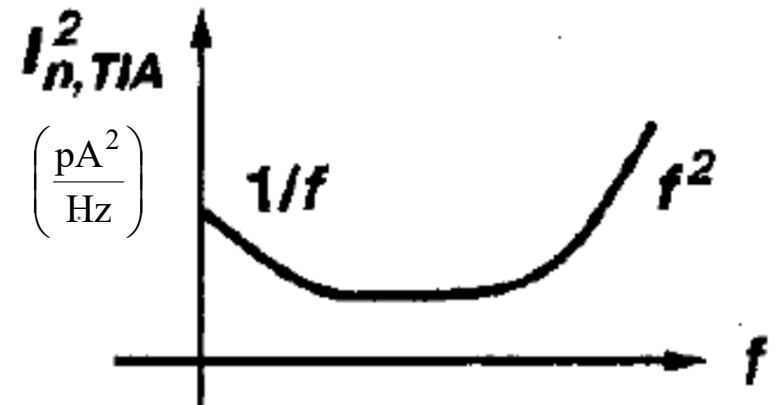
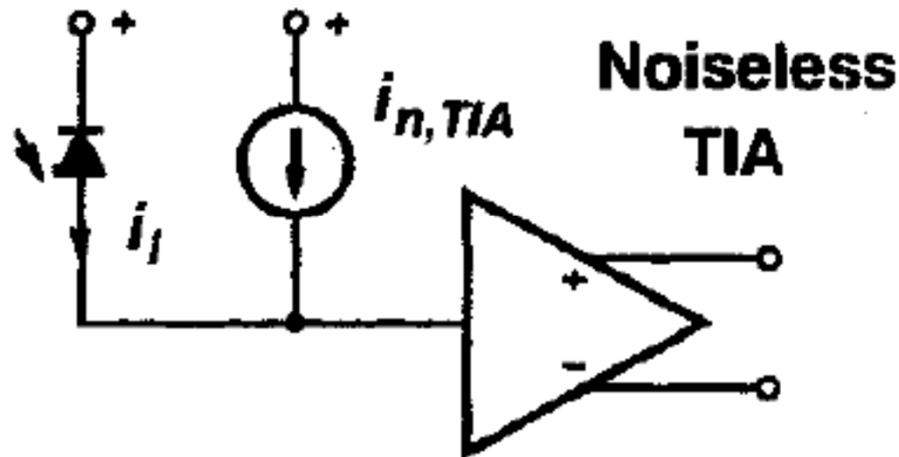
- CMOS inverter-based TIAs allow for reduced voltage headroom operation
- Cascaded inverter-gm + TIA stage provide additional voltage gain
- Low-bandwidth feedback loop sets the amplifier output common-mode level

Input-Referred Noise Current



- TIA noise is modeled with an input-referred noise current source that reproduces the output TIA output noise when passed through an ideal noiseless TIA
- This noise source will depend on the source impedance, which is determined mostly by the photodetector capacitance

Input-Referred Noise Current Spectrum



- Input-referred noise current spectrum typically consists of uniform, high-frequency f^2 , & low-frequency $1/f$ components
- To compare TIAs, we need to see this noise graph out to $\sim 2X$ the TIA bandwidth
 - Refer to noise bandwidth tables

Input-Referred RMS Noise Current

- The input-referred rms noise current can be calculated by dividing the rms output noise voltage by the TIA's midband transimpedance value

$$i_{n,TIA}^{rms} = \frac{1}{R_T} \sqrt{\int_0^{>2BW} |Z_T(f)|^2 I_{n,TIA}^2(f) df}$$

- If we integrate the output noise, the upper bound isn't too critical. Often this is infinity for derivations, or 2X the TIA bandwidth in simulation
- This rms current sets the TIA's electrical sensitivity

$$i_{sens}^{pp} = 2Q i_{n,TIA}^{rms}$$

- To determine the total optical receiver sensitivity, we need to consider the detector noise and responsivity

Averaged Input-Referred Noise Current Density

- TIA noise performance can also be quantified by the averaged input-referred noise current density

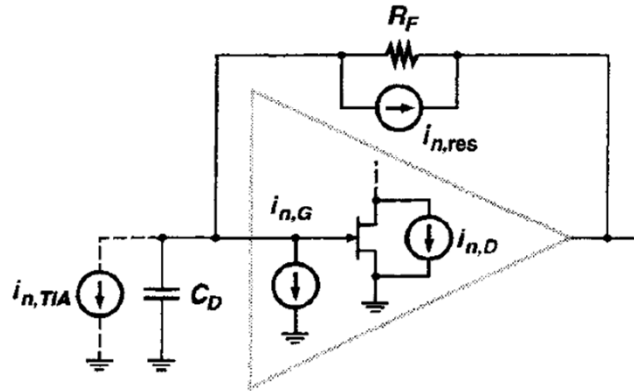
$$i_{n,TIA}^{avg} = \frac{i_{n,TIA}^{rms}}{\sqrt{BW_{3dB}}}$$

This quantity has units of $\left(\frac{\text{pA}}{\sqrt{\text{Hz}}} \right)$.

Note, this is different than averaging the input - referred noise spectrum,

$I_{n,TIA}^2(f)$ over the TIA bandwidth.

FET Feedback TIA Input-Referred Noise Current Spectrum



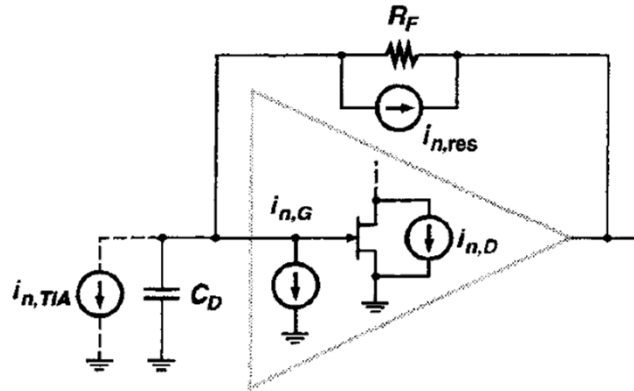
- The feedback resistor and amplifier front-end noise components determine the input-referred noise current spectrum

$$I_{n,TIA}^2(f) = I_{n,res}^2(f) + I_{n,front}^2(f)$$

- The feedback resistor component is uniform with frequency

$$I_{n,res}^2(f) = \frac{4kT}{R_F}$$

FET Feedback TIA Input-Referred Noise Current Spectrum



- Gate current-induced shot noise

$$I_{n,G}^2 = 2qI_G$$

This is typically small for CMOS designs

- FET channel noise

$$I_{n,D}^2 = 4kT\Gamma g_m$$

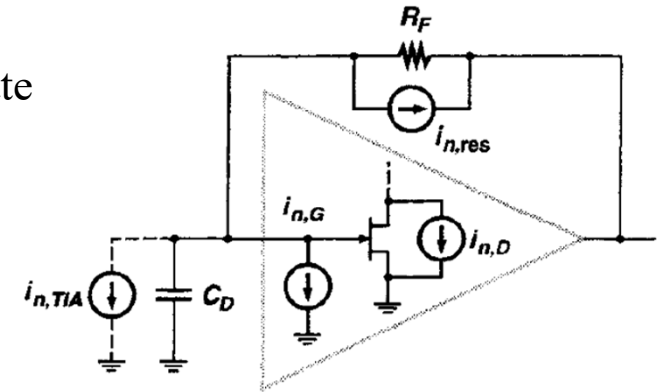
Γ is the channel noise factor, typically 0.7 - 3 depending on the process.

Input-Referring the FET Channel Noise

To do this, we could calculate $\frac{i_{n,TIA}}{i_{n,D}} = \frac{\left(\frac{v_{out}}{Z_T}\right)}{Z_T}$

But it is easier (and equivalent) to ground the output and calculate

$$\left(\frac{i_{n,D}}{i_{n,TIA}}\right)^{-1} = \frac{g_m i_{n,TIA}}{sC_T + \frac{1}{R_F}} = \frac{g_m R_F}{1 + sR_F C_T} i_{n,TIA}$$



where $C_T = C_D + C_I$, the summation of the detector and amplifier input capacitance.

$$\left(\frac{i_{n,D}}{i_{n,TIA}}\right)^{-1} = \frac{1 + sR_F C_T}{g_m R_F}$$

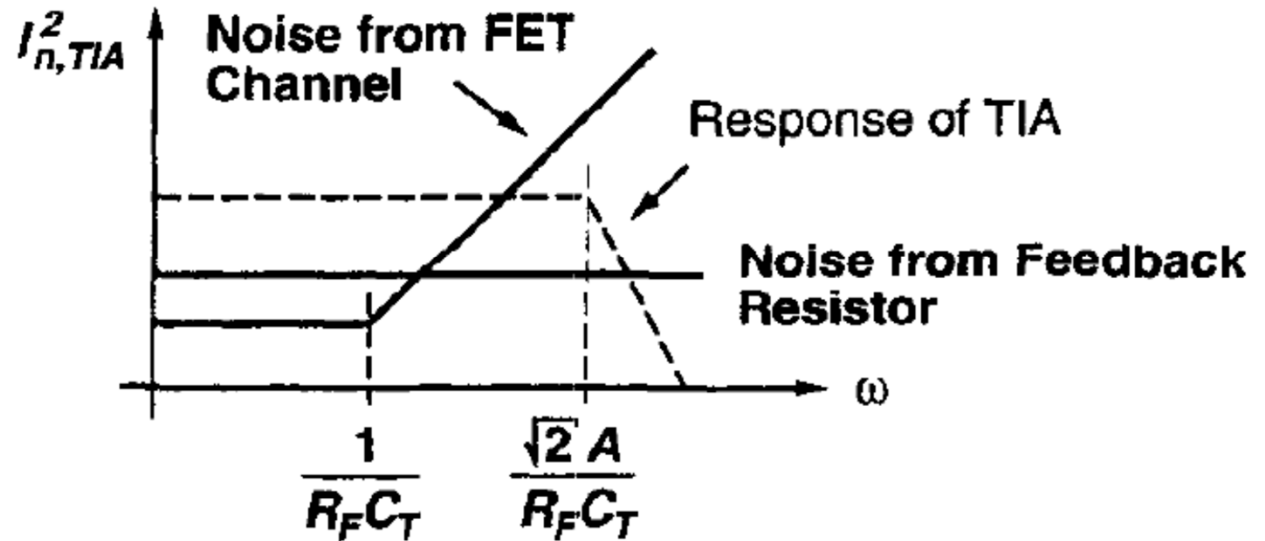
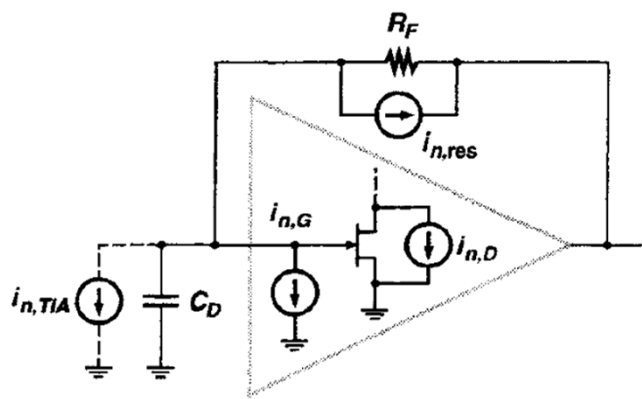
Using this high - pass transfer function, the input - referred FET channel noise is

$$I_{n,front,D}^2(f) = \frac{1 + (2\pi f R_F C_T)^2}{(g_m R_F)^2} \cdot 4kT\Gamma g_m$$

$$= 4kT\Gamma \left(\frac{1}{g_m R_F^2} \right) + 4kT\Gamma \left(\frac{(2\pi C_T)^2}{g_m} \right) f^2$$

← Uniform and f^2 component!

Total Input-Referred FET Feedback TIA Noise



$$I_{n,TIA}^2(f) = \frac{4kT}{R_F} + 2qI_G + 4kT\Gamma\left(\frac{1}{g_m R_F^2}\right) + 4kT\Gamma\left(\frac{(2\pi C_T)^2}{g_m}\right)f^2$$

Feedback Resistor

Gate Shot Noise

FET Channel Noise

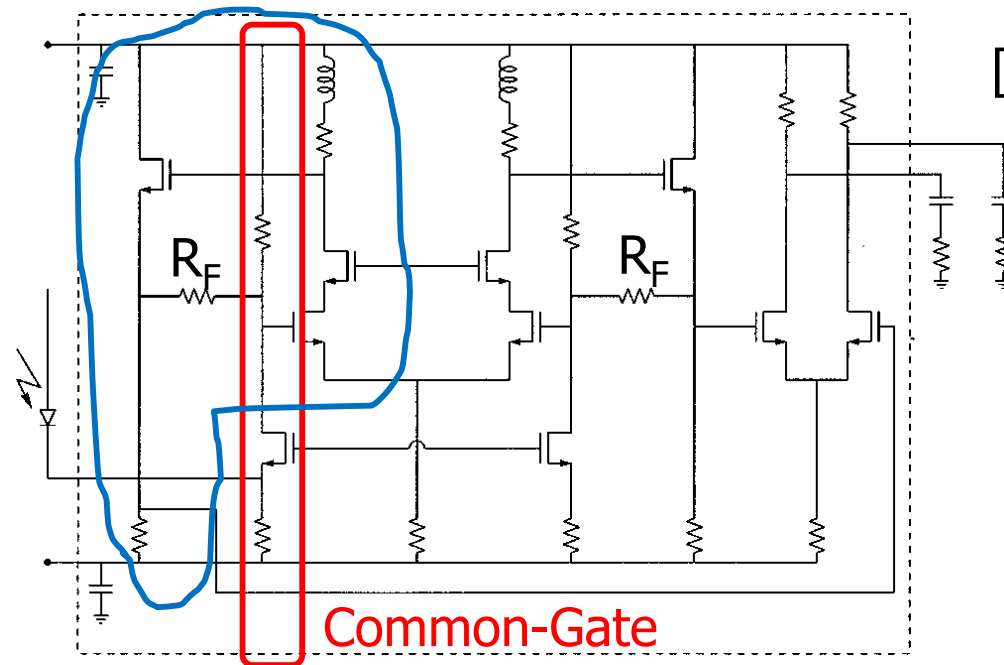
- Note that the TIA input-referred noise current spectrum begins to rise at a frequency lower than the TIA bandwidth

Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

Common-Gate & Feedback TIA

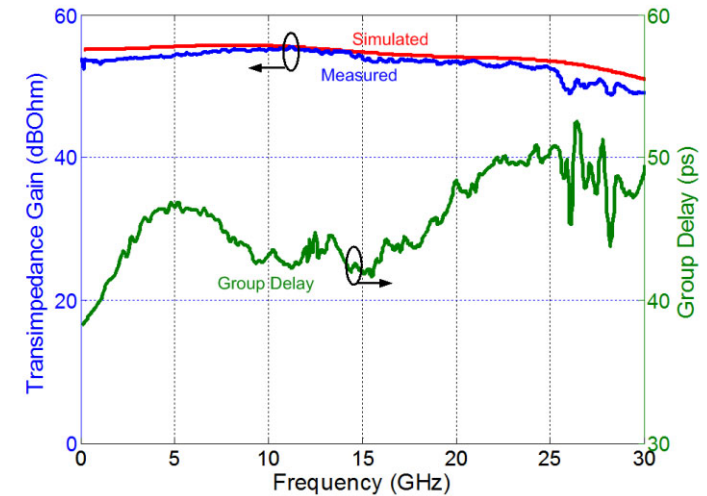
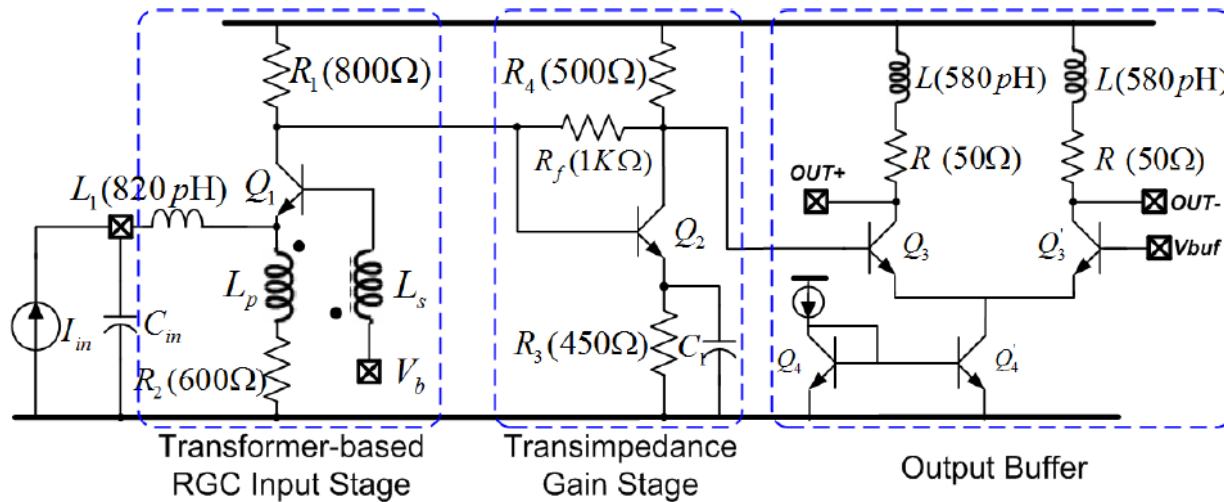
Feedback TIA



[Mohan JSSC 2000]

- Recall that the feedback TIA stability depends on the ratio of the input pole (set by C_D) and the amplifier pole
 - Large variation in C_D can degrade amplifier stability
- Common-gate input stage isolates C_D from input amplifier capacitance, allowing for a stable response with a variety of different photodetectors
- Transimpedance is still approximately $R_F A / (1 + A)$

BJT Common-Base & Feedback TIA



- Transformer-based negative feedback boosts g_m with low power and noise overhead
- Input series peaking inductor isolates the photodetector capacitance from the TIA input capacitance
- High frequency techniques allow for 26GHz bandwidth with group delay variation less than 19ps

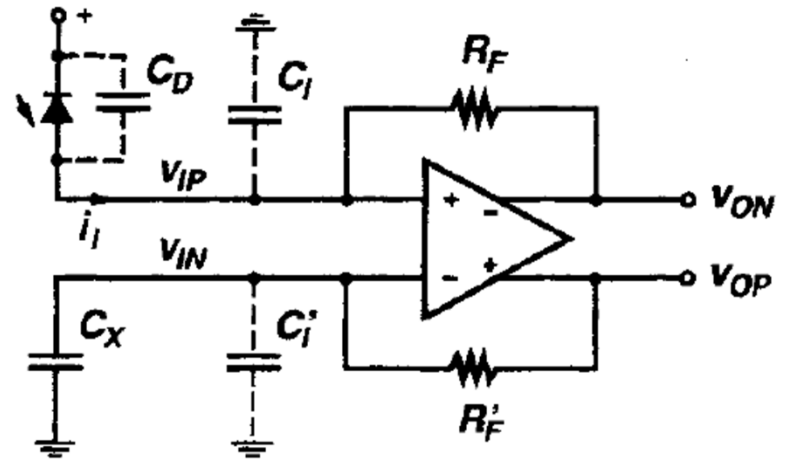
[Li JSSC 2013]

Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

Differential TIAs

- Differential circuits have superior immunity to power supply/substrate noise
- A differential TIA output allows easy use of common differential main/limiting amplifiers
 - This comes at the cost of higher noise and power
- How to get a differential output with a single-ended photocurrent input?
 - Two common approaches, based on the amount of capacitance applied at the negative input

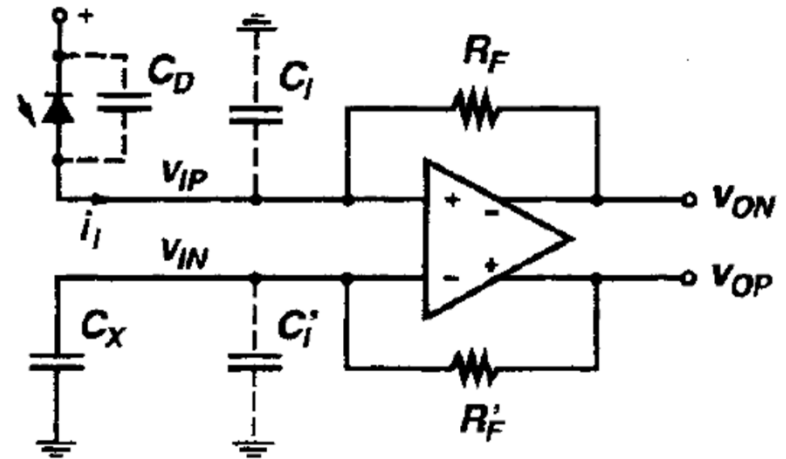


Balanced TIA

- A balanced TIA design attempts to match the capacitance of the two differential inputs

$$C_X \approx C_D$$

- This provides the best power supply/substrate noise immunity, as the noise transfer functions are similar
- Due to double the circuitry, the input-referred rms noise current is increased by $\sqrt{2}$



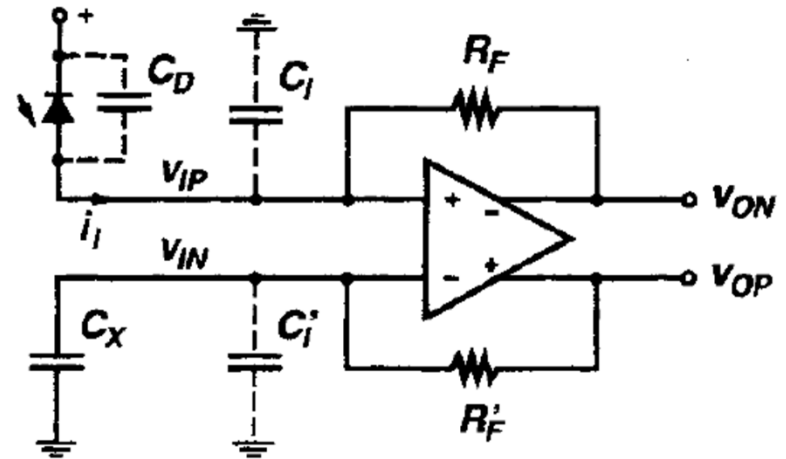
Assuming an high BW amplifier
and $C_T = C_D + C_I$

$$Z_T(s) = \frac{v_{OP} - v_{ON}}{i_i} = \frac{\left(\frac{A}{A+1}\right)R_F}{1 + \frac{sC_T R_F}{A+1}}$$

Same transfer function as the single - ended design

Pseudo-Differential TIA

- A pseudo-differential TIA design uses a very large capacitor at the negative input, such that it can be approximated as an AC ground $C_X \rightarrow \infty$
- While not good to reject power supply/substrate noise, it does provide significant filtering of the R_F' noise
- The differential transimpedance is approximately doubled relative to the single-ended case

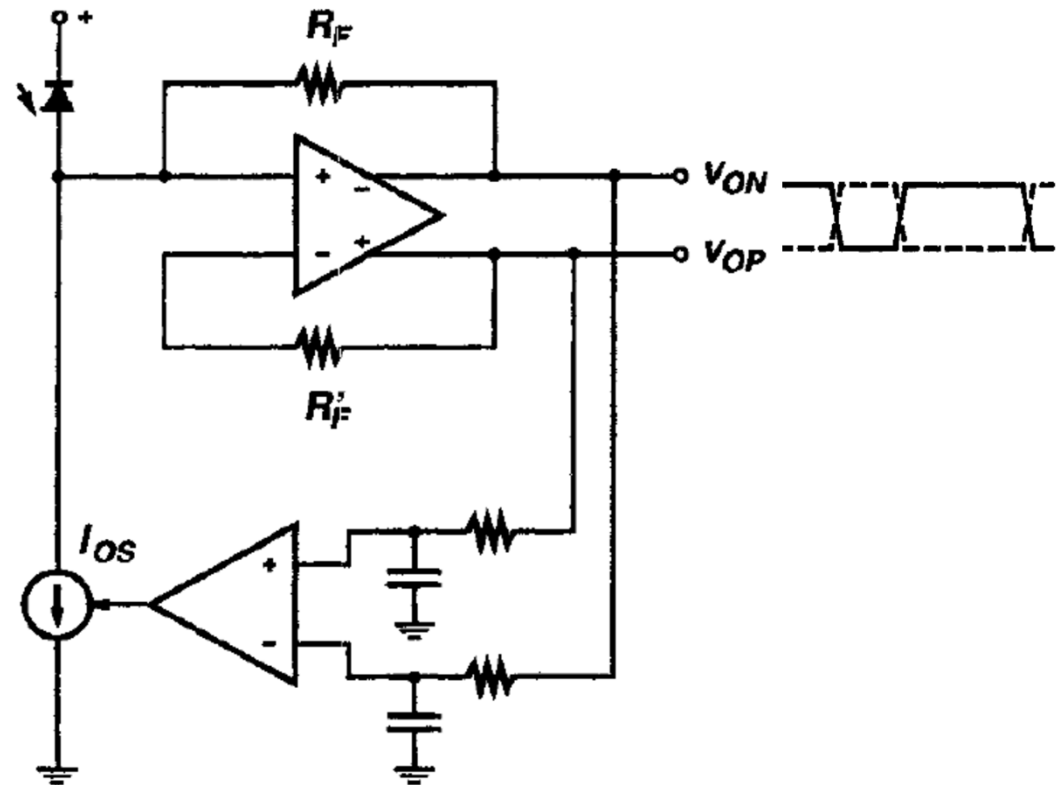
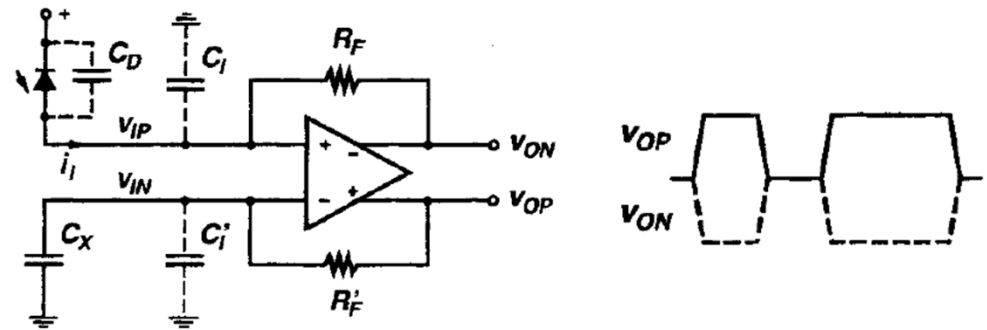


Assuming an high BW amplifier
and $C_T = C_D + C_I$

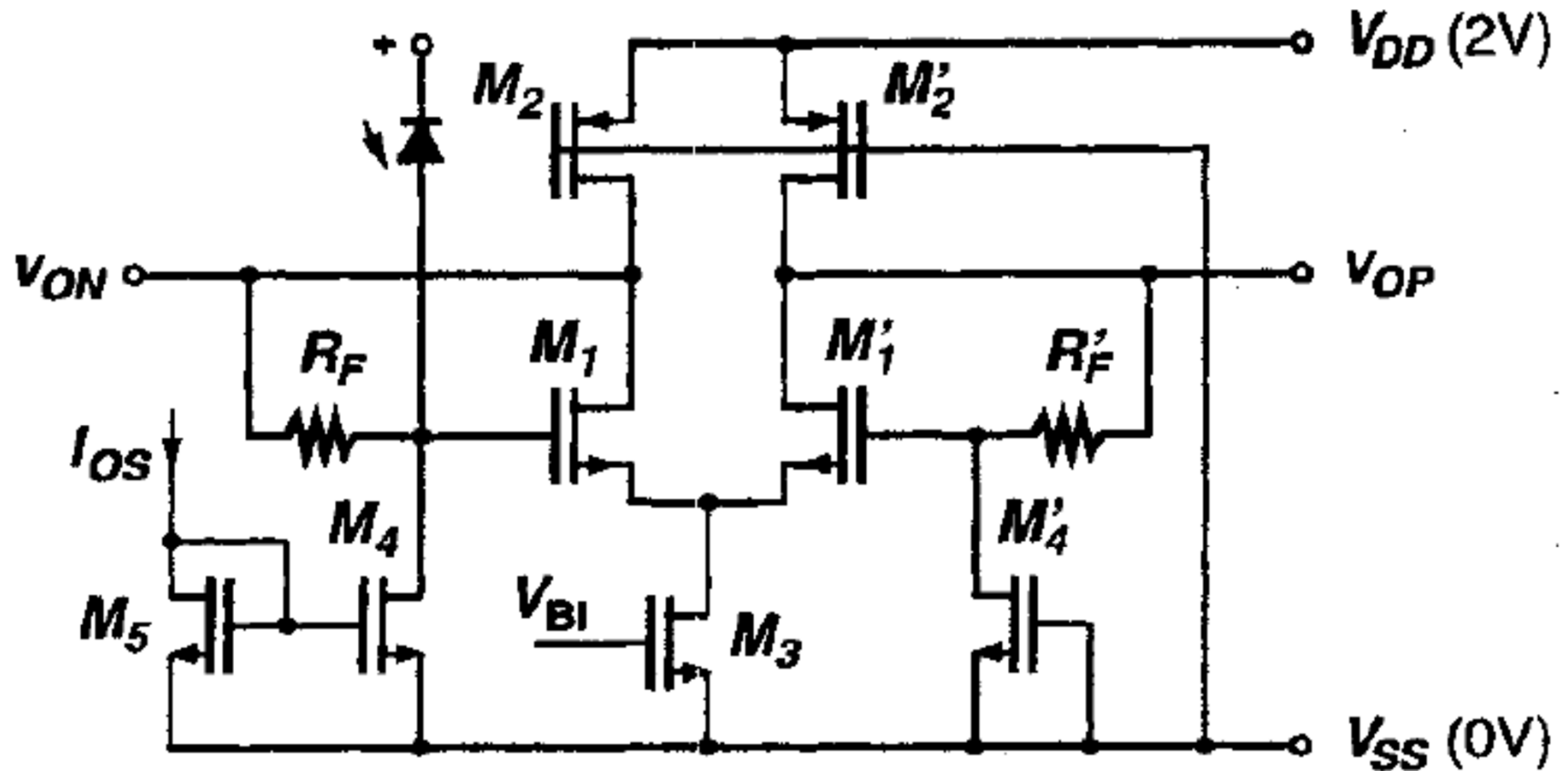
$$Z_T(s) = \frac{v_{OP} - v_{ON}}{i_i} = \frac{\left(\frac{2A}{A+2}\right)R_F}{1 + \frac{sC_T R_F}{\frac{A}{2} + 1}}$$

Offset Control

- Due to the single-ended photodetector signal, the differential output signal swings from 0 to V_{ppd} , which can limit the dynamic range
- Adding offset control circuitry can allow for an output swing of $\pm V_{ppd}/2$



Differential Shunt Feedback TIA



Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

Optical RX Scaling Issues

- ☺ Traditionally, TIA has high R_T and low R_{in}

$$R_T = R_F \left(\frac{A}{1+A} \right)$$

$$\omega_{3dB} \approx \frac{1+A}{R_F C_{IN}}$$

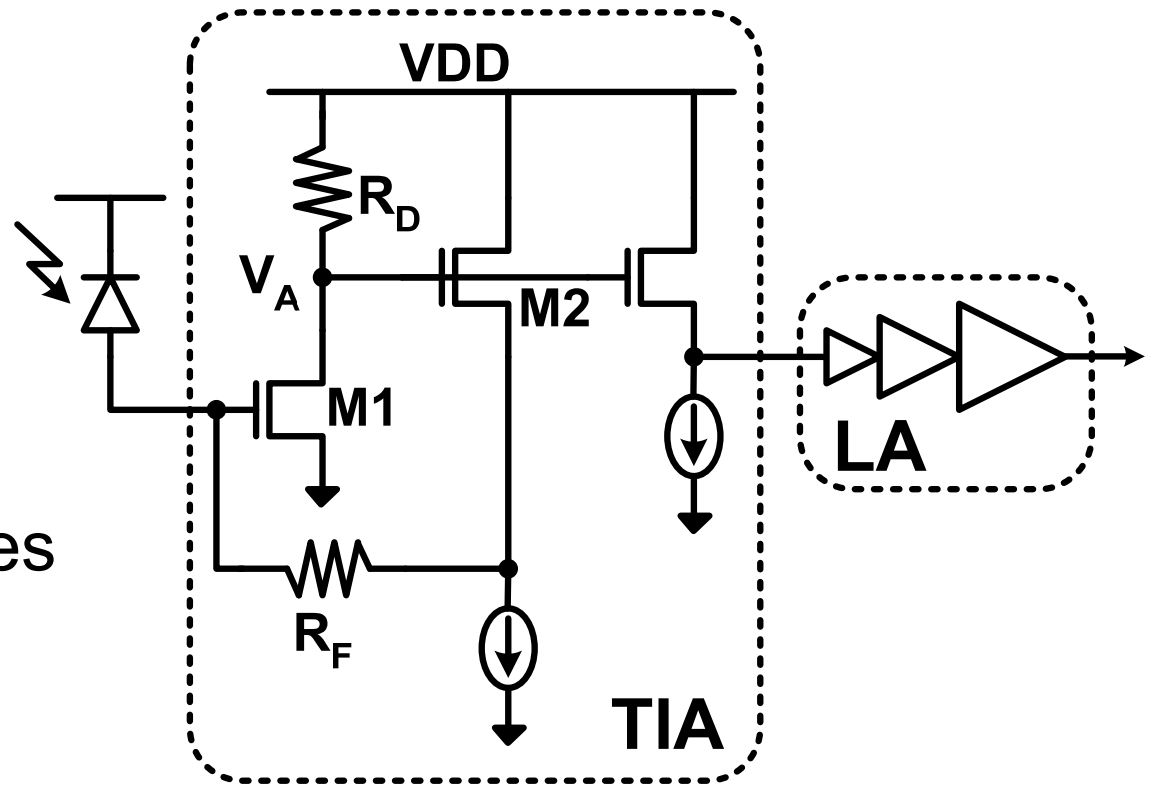
- ☹ Headroom/Gain issues in 1V CMOS

- $A \approx 2 - 3$

- ☹ Power/Area Costs

$$\text{TIA } I_D \propto (R_T C_{IN})^2 f_{3dB}^4$$

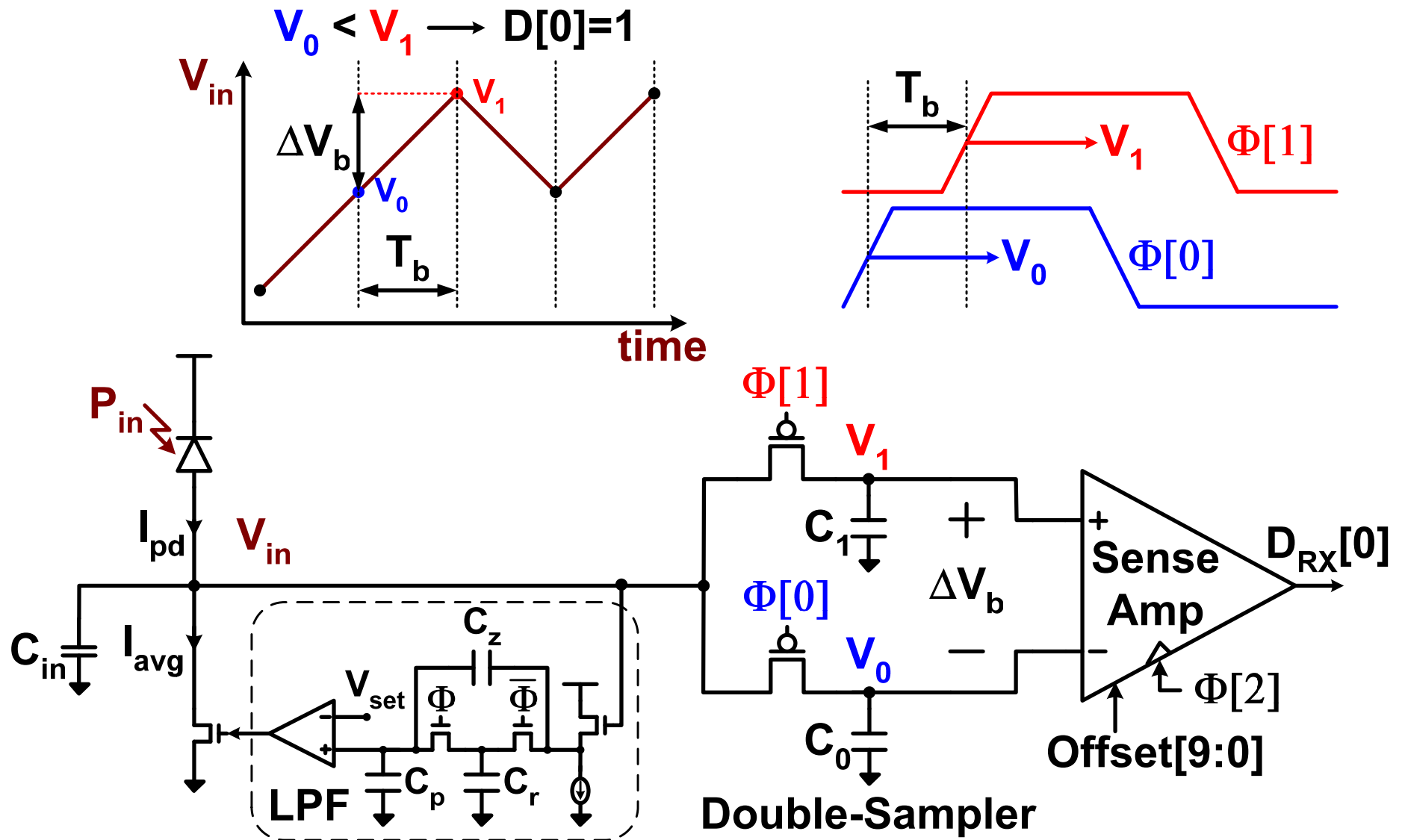
$$\text{LA } I_D \propto f_{3dB}^2$$



$$V_A = V_{GS1} + V_{GS2} \approx 0.8 * VDD$$

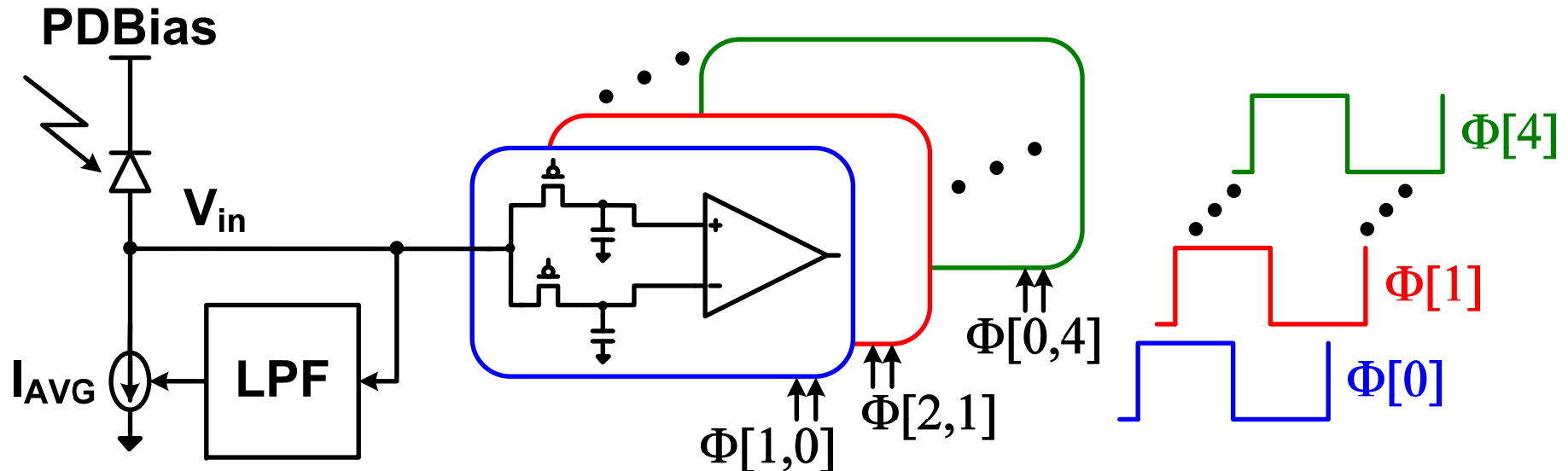
$$A \approx g_{m1} R_D = \frac{\alpha(VDD - V_A)}{VOD} \approx \frac{\alpha(0.2 * VDD)}{VOD}$$

Integrating Receiver Block Diagram



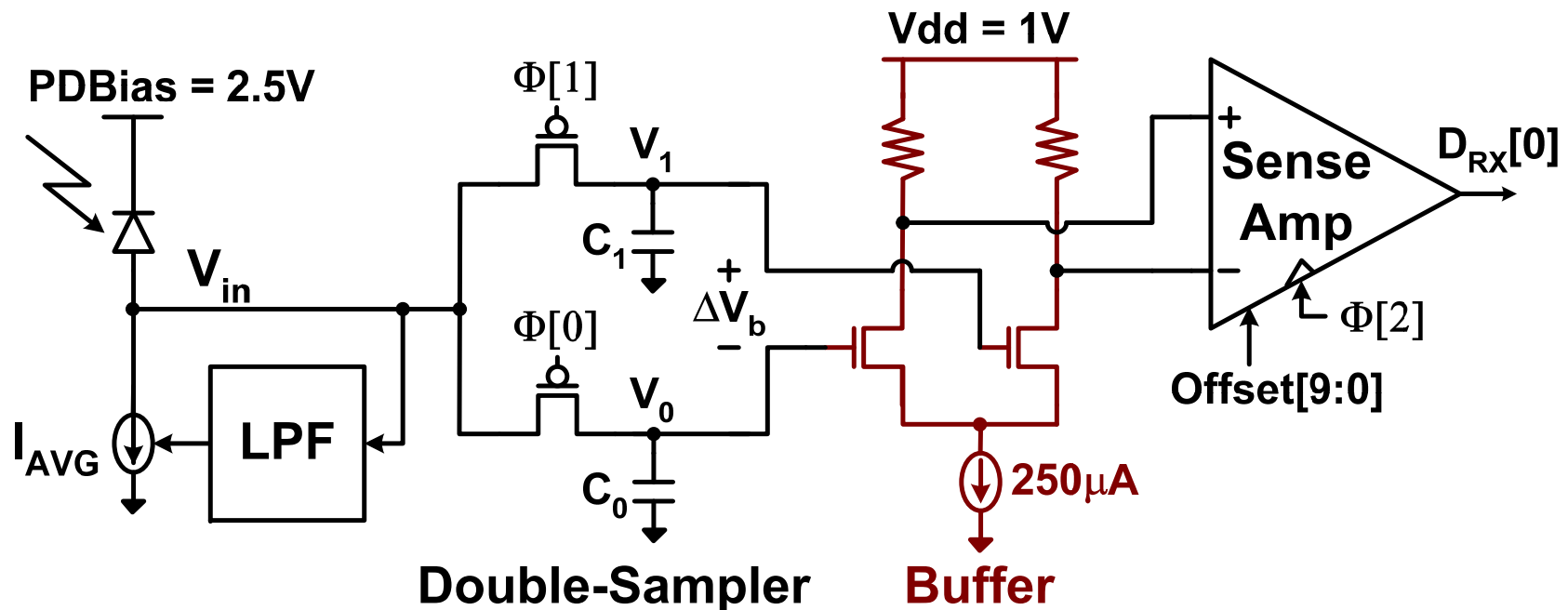
[Emami VLSI 2002]

Demultiplexing Receiver



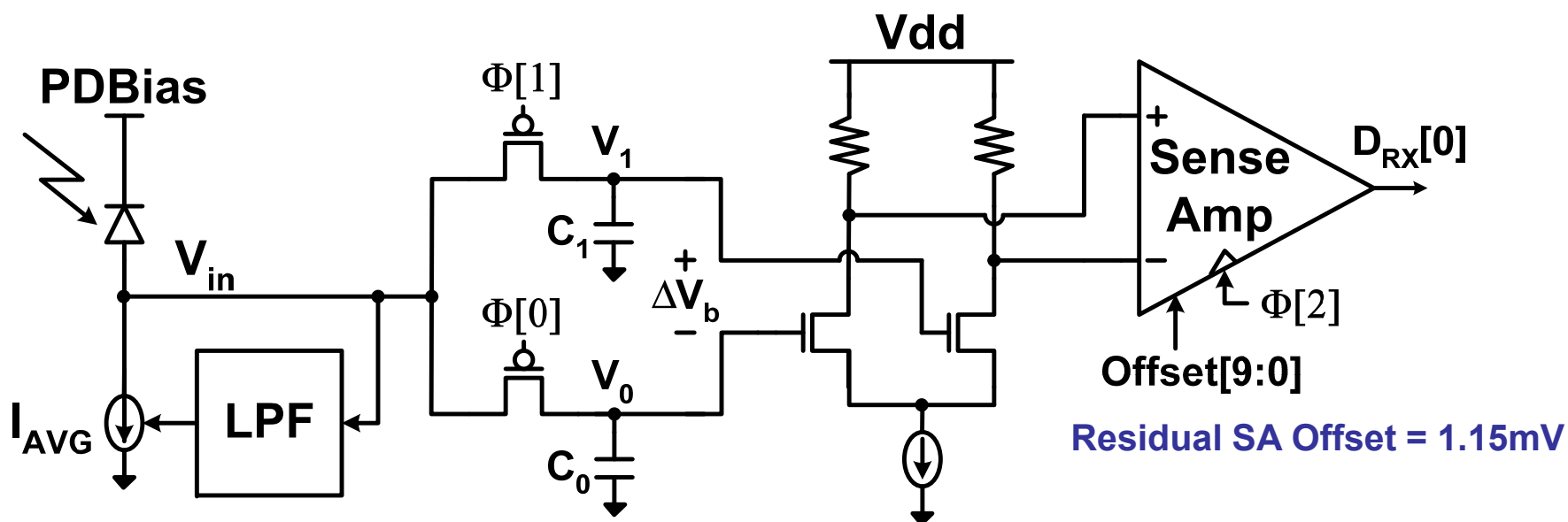
- Demultiplexing with multiple clock phases allows higher data rate
 - Data Rate = #Clock Phases x Clock Frequency
 - Gives sense-amp time to resolve data
 - Allows continuous data resolution

1V Modified Integrating Receiver



- Differential Buffer
 - ☺ Fixes sense-amp common-mode input for improved speed and offset performance
 - ☺ Reduces kickback charge
 - ☹ Cost of extra power and noise
- Input Range = 0.6 – 1.1V

Receiver Sensitivity Analysis



$$\text{Max } \Delta V_{in}(\Delta I_{AVG}) = 0.6mV$$

$$\sigma_{samp} = \sqrt{\frac{2kT}{C_{smp}}} = 0.92mV \quad \sigma_{buffer} = 1.03mV \quad \sigma_{SA} = 0.45mV$$

$$\text{Clock Jitter Noise } \sigma_{clk} = \left(\frac{\sigma_j}{T_b} \right) \Delta v_b \approx 0.65mV \text{ at } 16Gb/s$$

$$\text{Total Input Noise } \sigma_{tot} = \sqrt{\sigma_{smp}^2 + \sigma_{buffer}^2 + \sigma_{SA}^2 + \sigma_{clk}^2} = 1.59mV$$

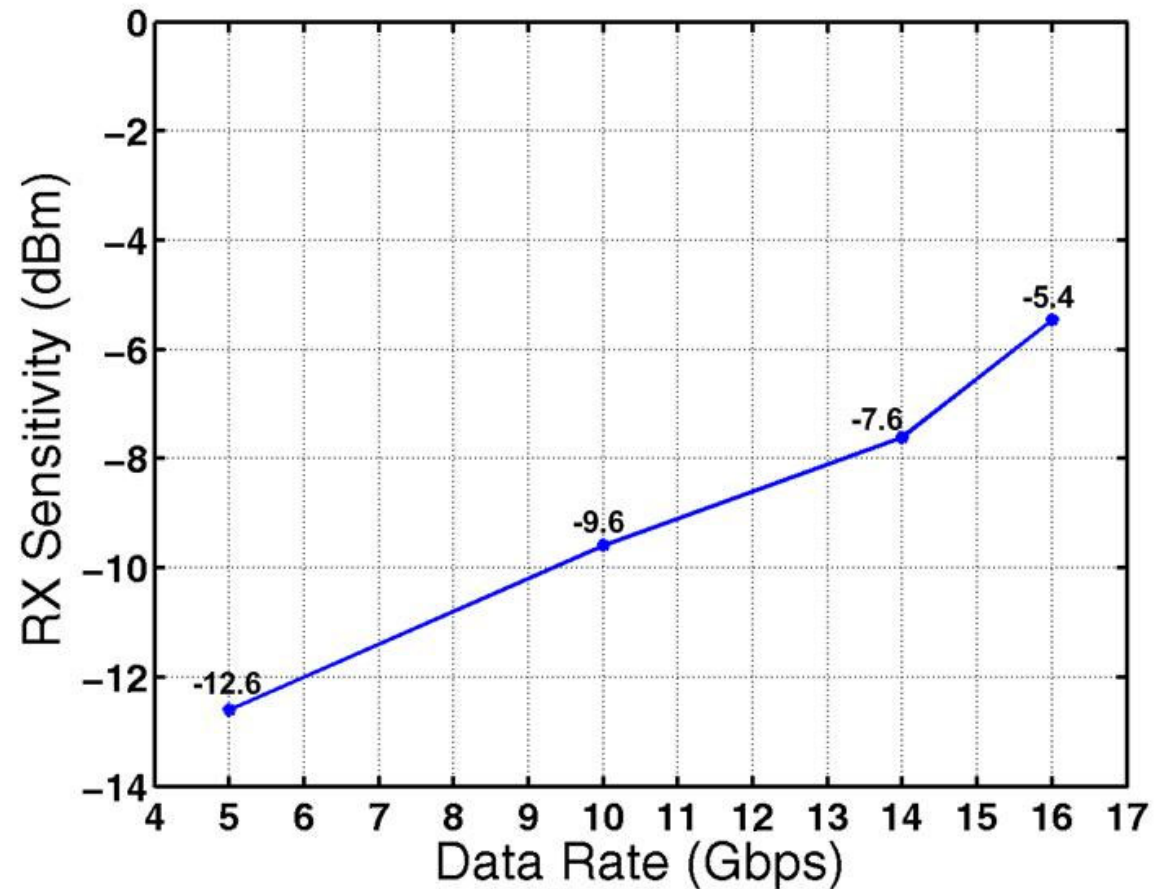
$$\Delta V_b \text{ for BER} = 10^{-10} = 6.4\sigma_{tot} + \text{Offset} = 11.9mV$$

$$P_{avg} = \frac{\Delta V_b (C_{pd} + C_{in})}{\rho T_b}$$

Gb/s	P _{avg} (dBm)
10	-9.8
16	-7.8

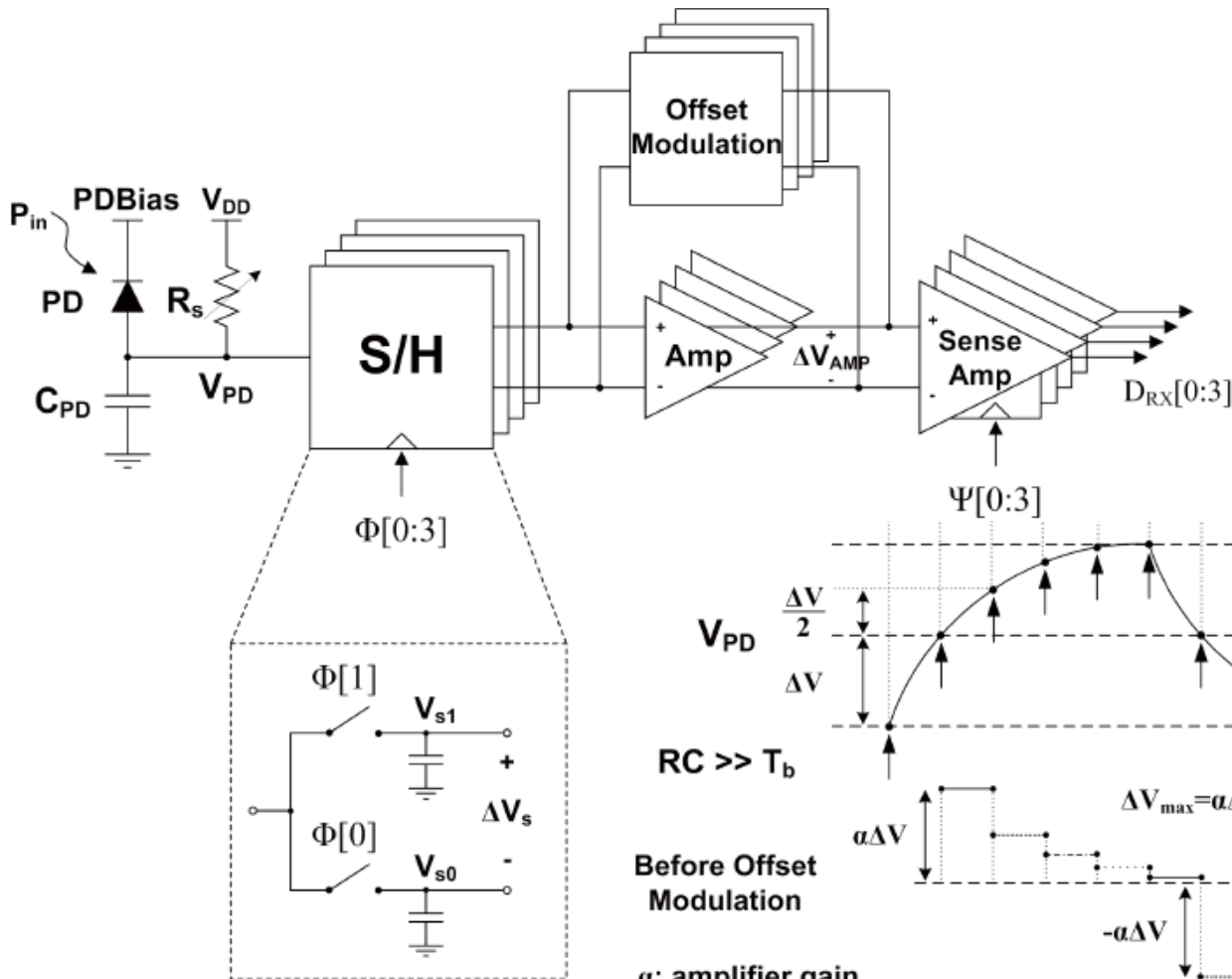
Integrating Receiver Sensitivity

- Test Conditions
 - 8B/10B data patterns (variance of 6 bits)
 - Long runlength data (variance of 10 bits)
- $\text{BER} < 10^{-10}$



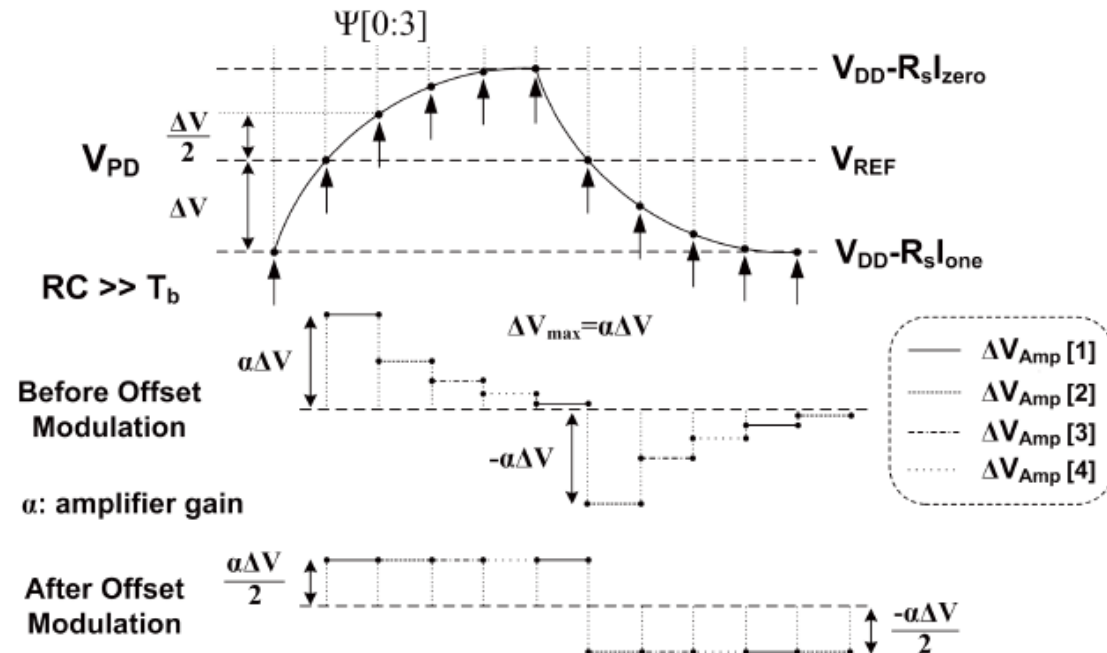
[Palermo JSSC 2008]

Integrating RX with Dynamic Threshold

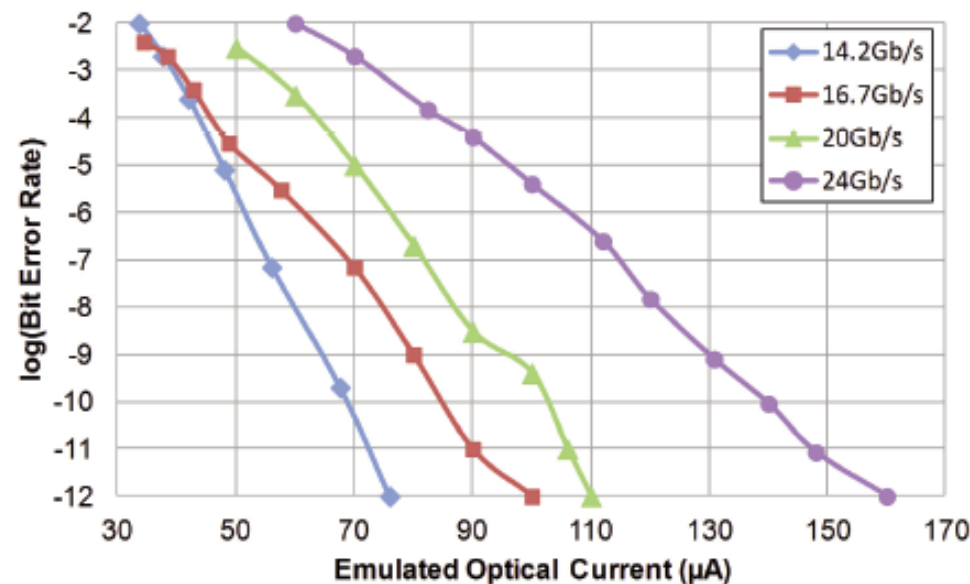
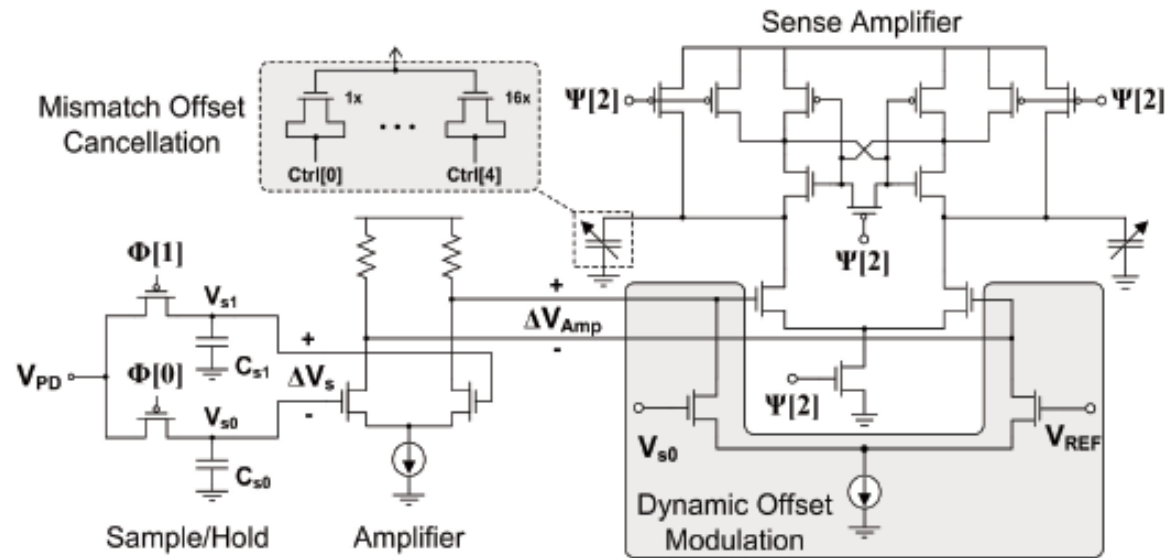


- Dynamic threshold adjustment allows for un-coded data

[Nazari ISSCC 2012]



Integrating RX with Dynamic Threshold



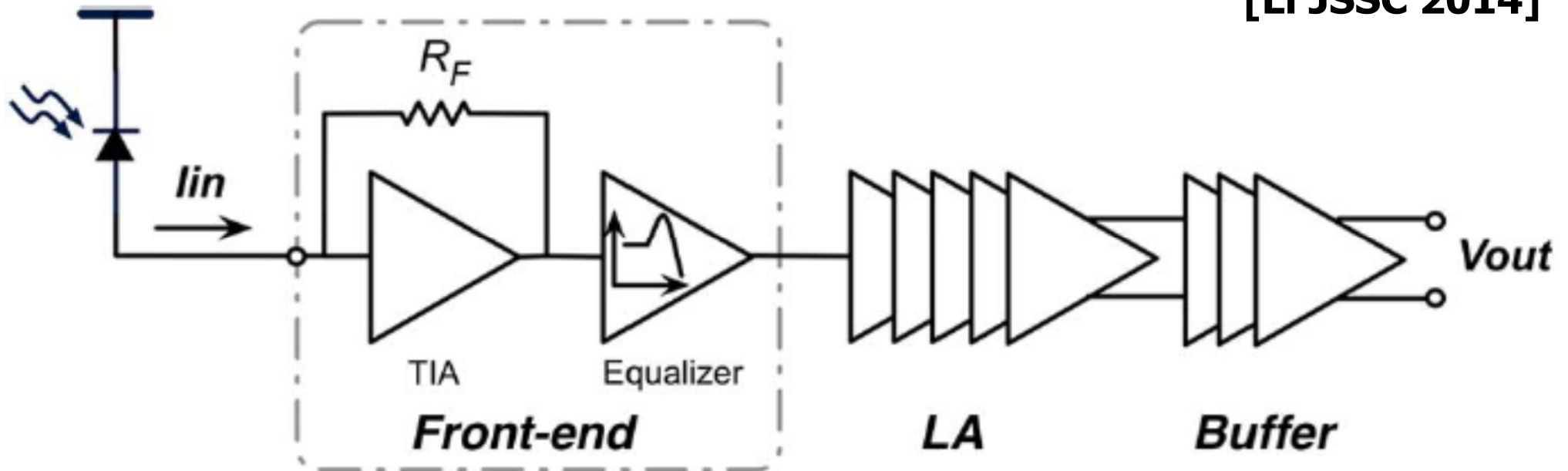
[Nazari ISSCC 2012]

Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
 - Common-Gate TIAs
 - Feedback TIAs
 - Common-Gate & Feedback TIA Combinations
 - Differential TIAs
- Integrating Optical Receivers
- Equalization in Optical Front-Ends

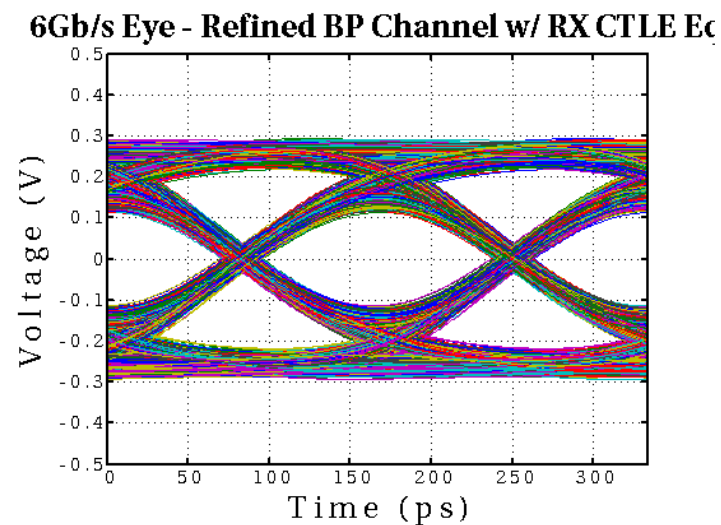
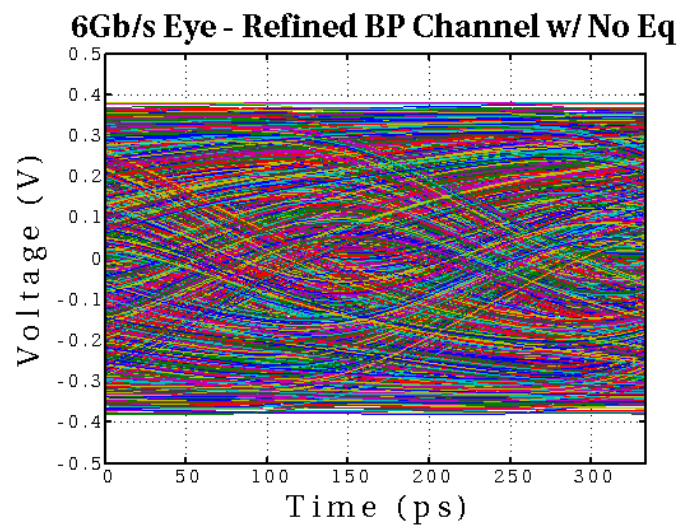
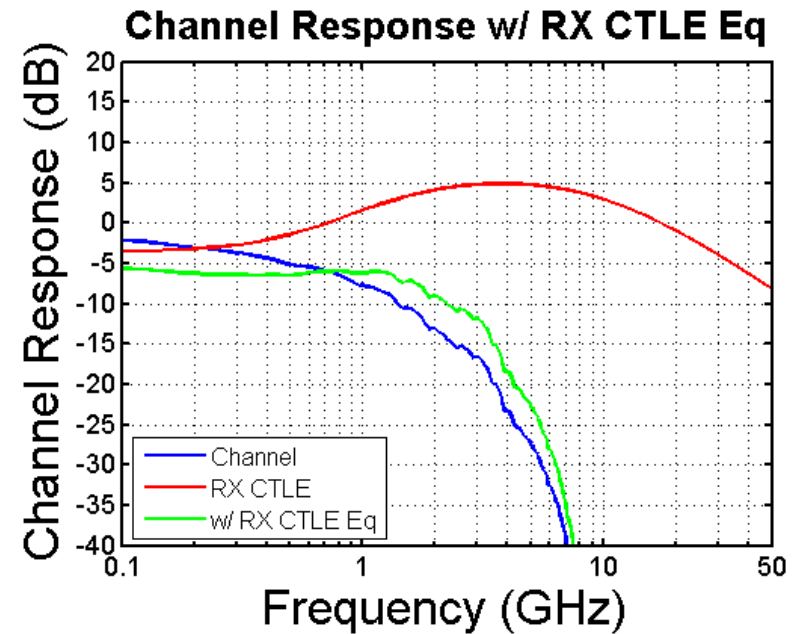
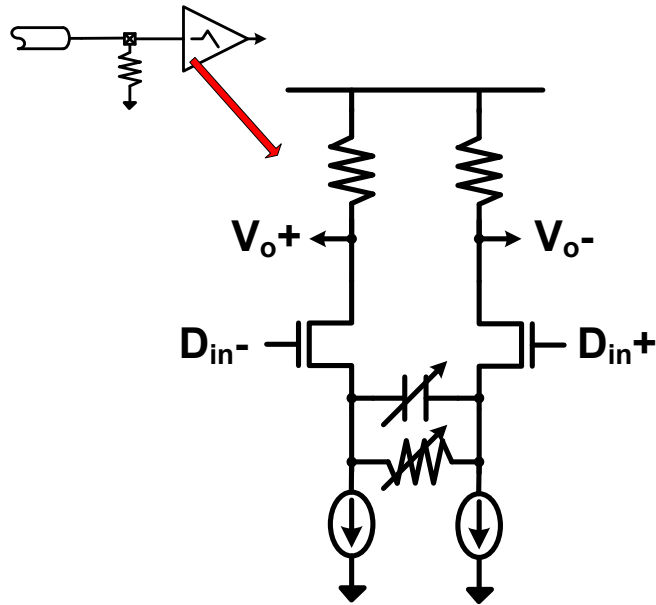
Low-BW TIA & CTLE Front-End

[Li JSSC 2014]

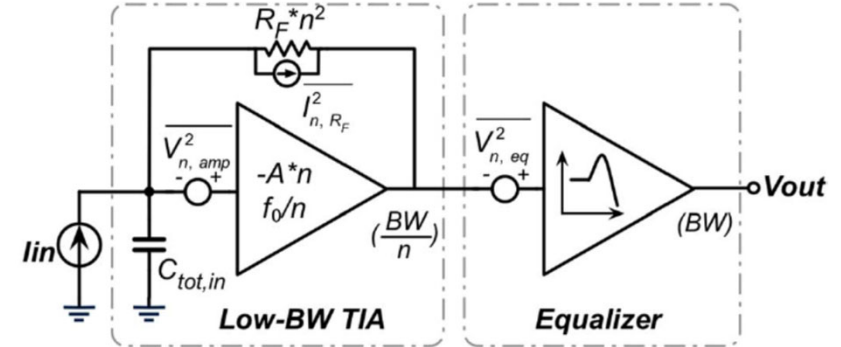
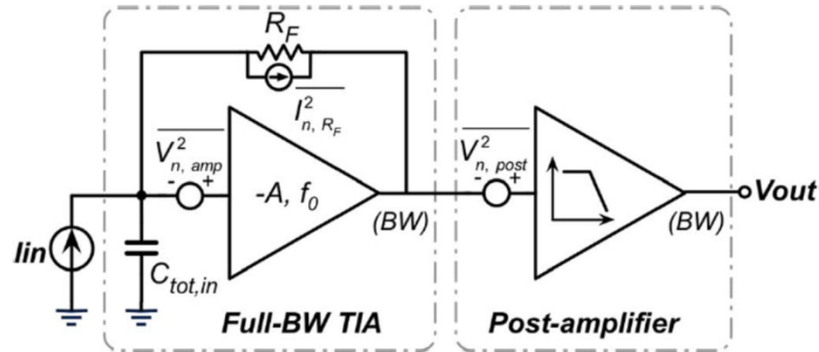


- Improved sensitivity is possible by increasing the first stage feedback resistor, resulting in a high-gain low-bandwidth TIA
- The resultant ISI is cancelled by a subsequent CTLE

Active CTLE Example



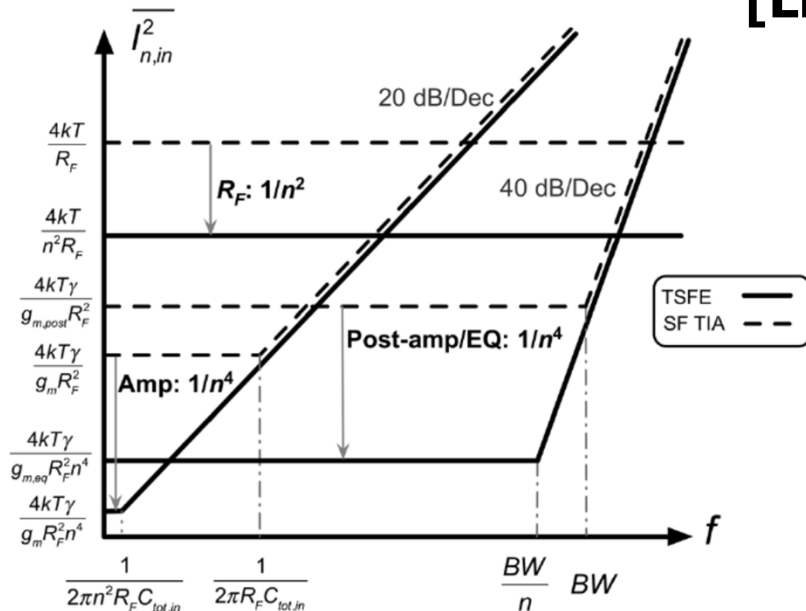
Low-BW TIA & CTLE Front-End



$$\overline{I_{n,\text{in,SF}}^2}(f) = \frac{4kT}{R_F} + \frac{4kT\gamma}{g_m R_F^2} + 4kT\gamma \frac{(2\pi C_{\text{tot,in}})^2}{g_m} f^2 + \frac{4kT\gamma}{g_{m,\text{post}} R_F^2} + \frac{4kT\gamma}{g_{m,\text{post}} R_F^2} \left(\frac{f}{BW} \right)^4$$

$$\begin{aligned} \overline{I_{n,\text{in,TSFE}}^2}(f) &= \frac{4kT}{R_F n^2} + \frac{4kT\gamma}{g_m R_F^2 n^4} \\ &+ 4kT\gamma \frac{(2\pi R_F n^2 C_{\text{tot,in}})^2}{g_m R_F^2 n^4} f^2 \\ &+ \frac{4kT\gamma}{g_{m,\text{eq}} R_F^2 n^4} + \frac{4kT\gamma}{g_{m,\text{eq}} R_F^2 n^4} \left(\frac{f}{BW/n} \right)^4 \\ &= \frac{4kT}{R_F n^2} + \frac{4kT\gamma}{g_m R_F^2 n^4} + 4kT\gamma \frac{(2\pi C_{\text{tot,in}})^2}{g_m} f^2 \\ &+ \frac{4kT\gamma}{g_{m,\text{eq}} R_F^2 n^4} + \frac{4kT\gamma}{g_{m,\text{eq}} R_F^2} \left(\frac{f}{BW} \right)^4 \quad (5) \end{aligned}$$

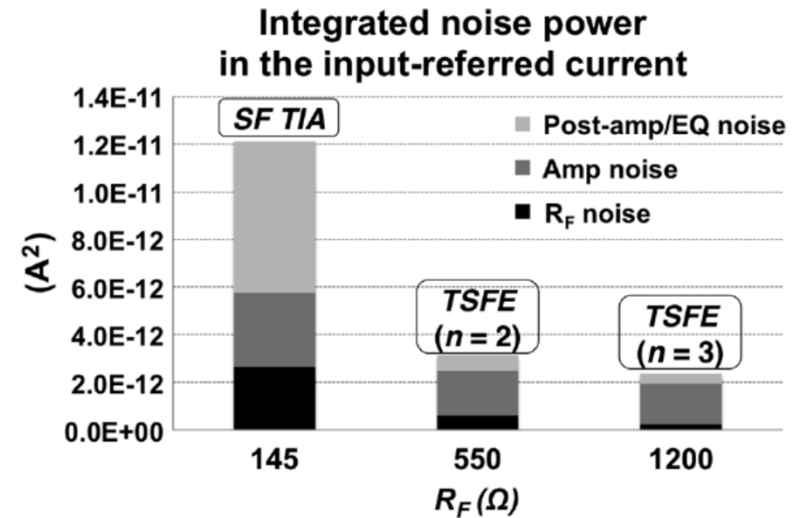
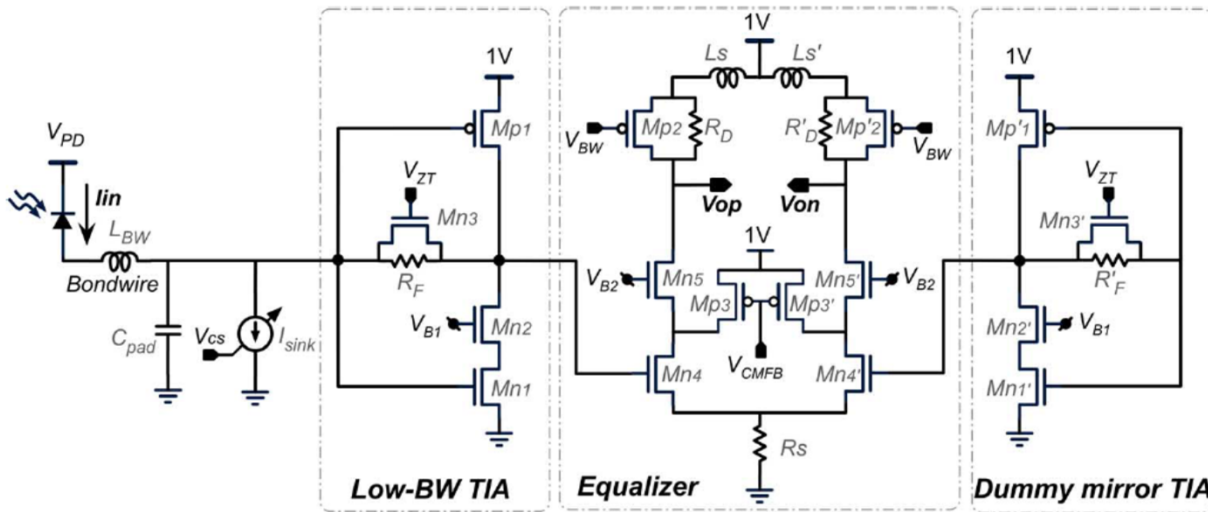
[Li JSSC 2014]



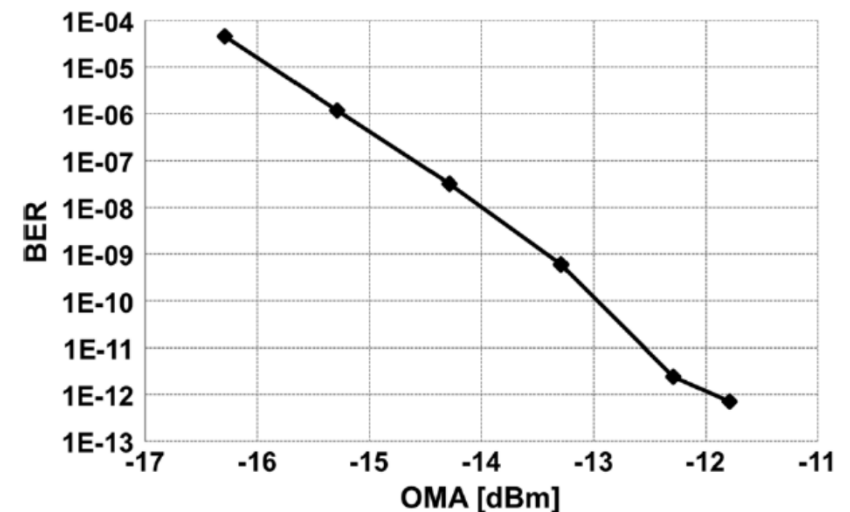
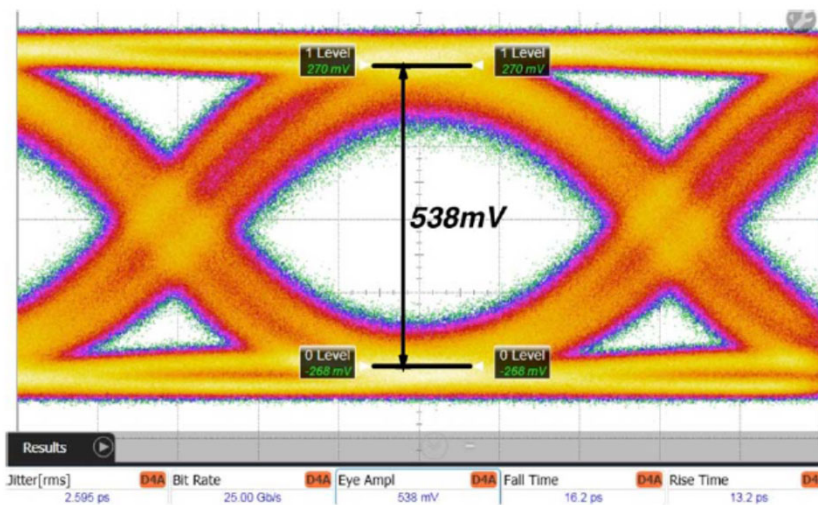
- Significant reduction in feedback resistor noise
- Low-frequency input and post amplifier noise is also reduced

Low-BW TIA & CTLE Front-End

[Li JSSC 2014]

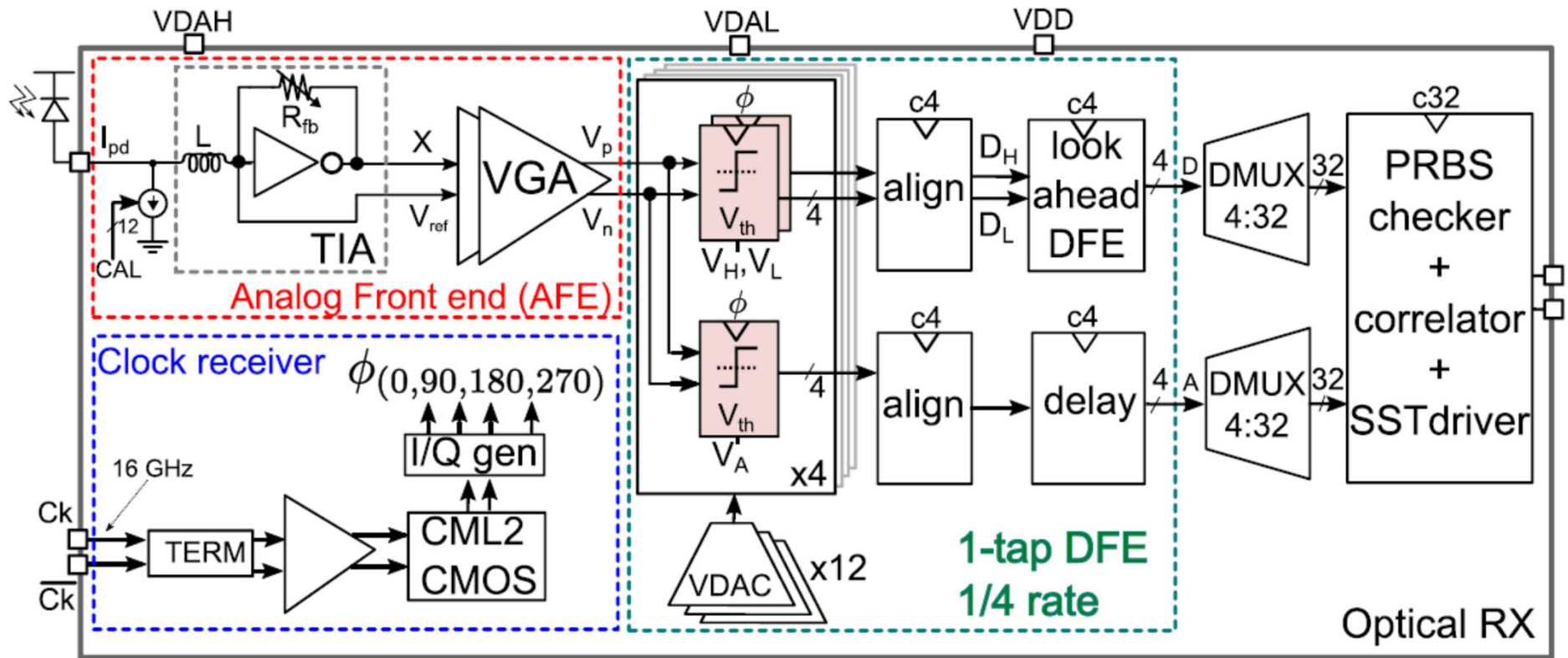


25Gb/s Eye Diagram



Low-BW TIA & DFE RX

[Ozkaya JSSC 2017]

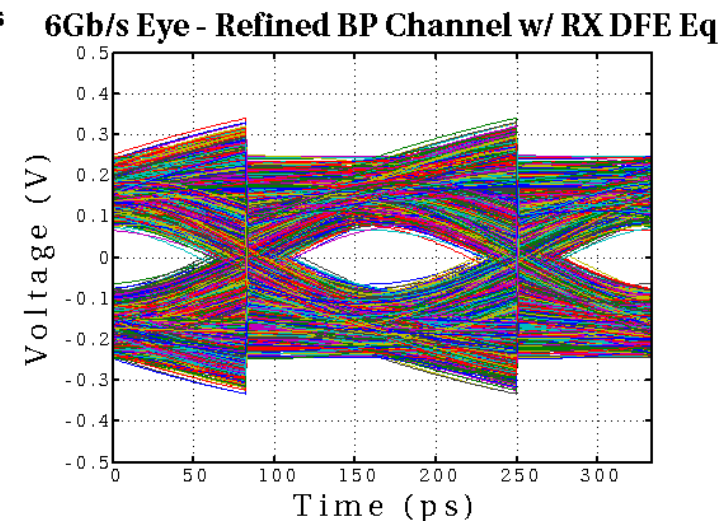
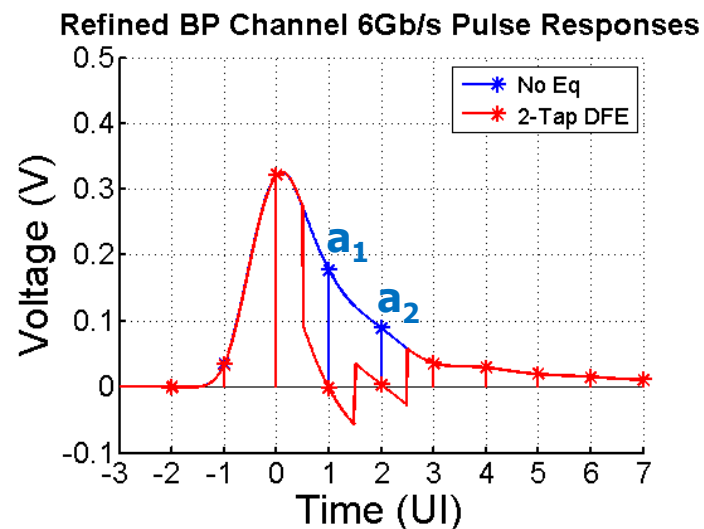
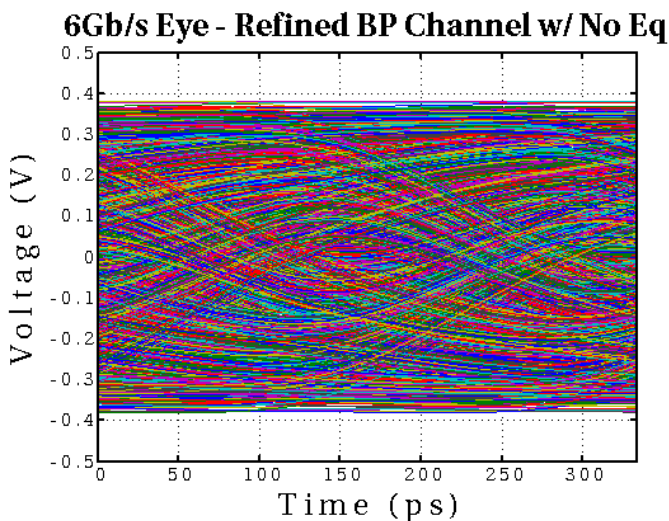
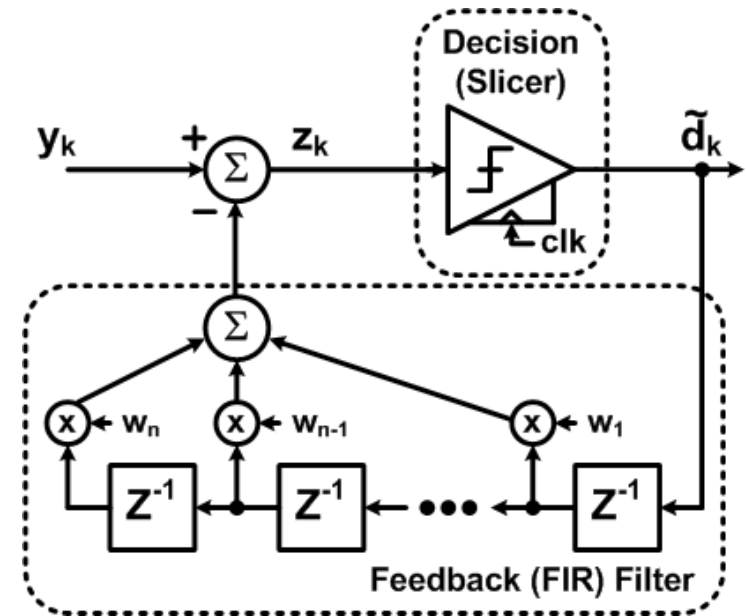


- In a similar manner, a high-gain low-bandwidth TIA is utilized
- The resultant ISI is cancelled by a subsequent 1-tap loop-unrolled DFE

DFE Example

- If only DFE equalization, DFE tap coefficients should equal the unequalized channel pulse response values $[a_1 \ a_2 \ \dots \ a_n]$
- With other equalization, DFE tap coefficients should equal the pre-DFE pulse response values
 - DFE provides flexibility in the optimization of other equalizer circuits
 - i.e., you can optimize a TX equalizer without caring about the ISI terms that the DFE will take care of

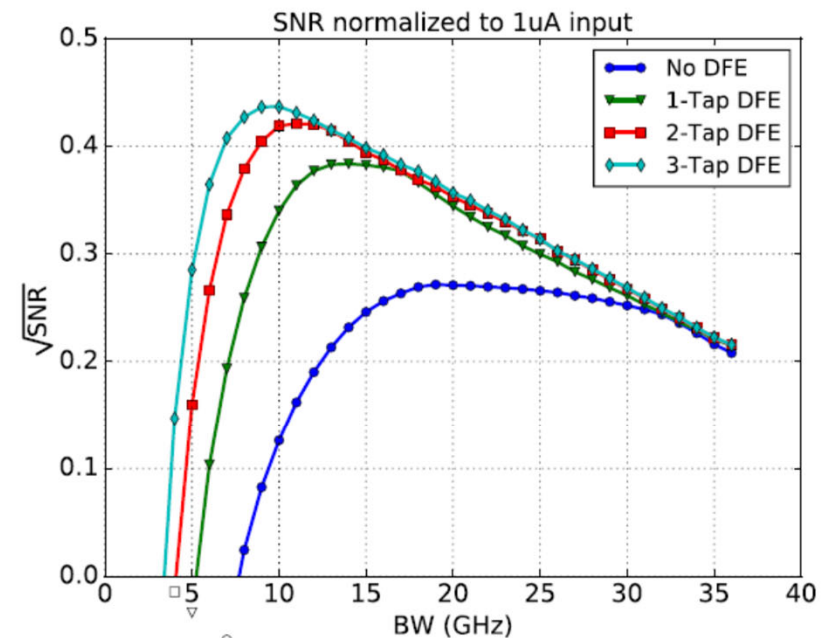
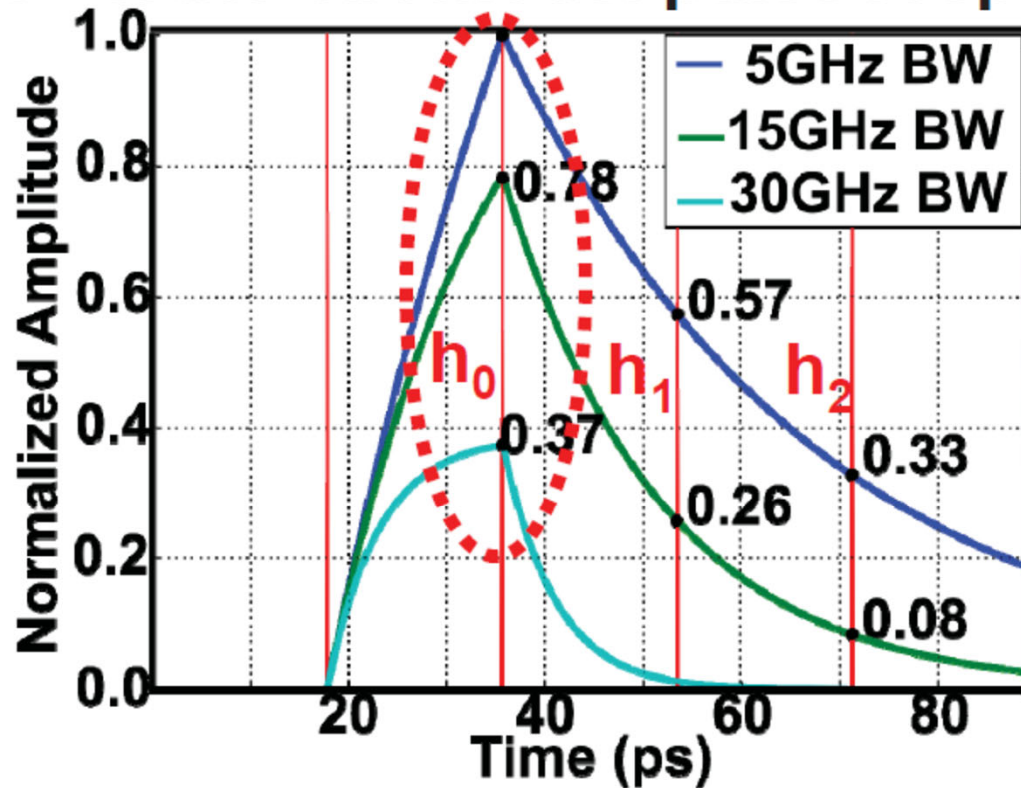
$$[w_1 \ w_2] = [a_1 \ a_2]$$



Low-BW TIA & DFE RX

[Ozkaya JSSC 2017]

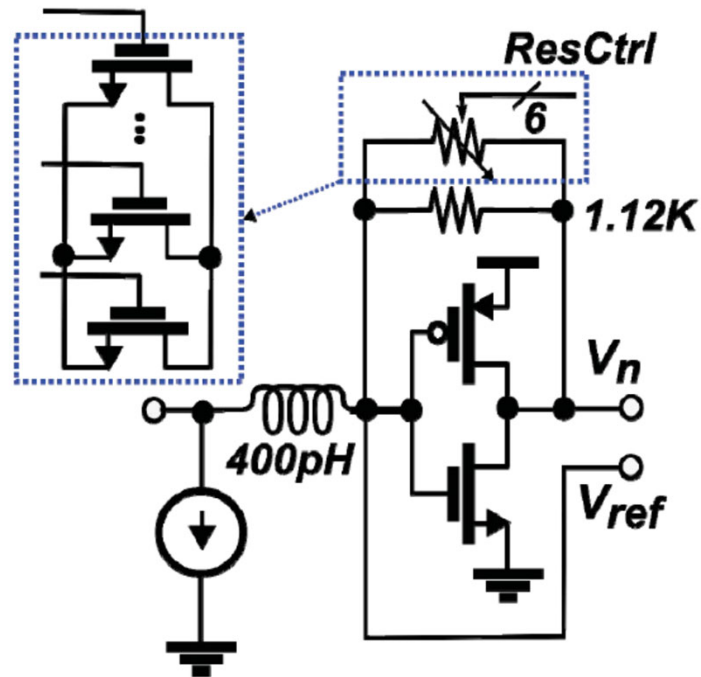
1st order TIA model pulse response (56Gb/s)



- As RF is increased, the main cursor increases and the SNR improves as ISI is cancelled by a DFE
- Large performance benefit with a low-complexity 1-tap DFE

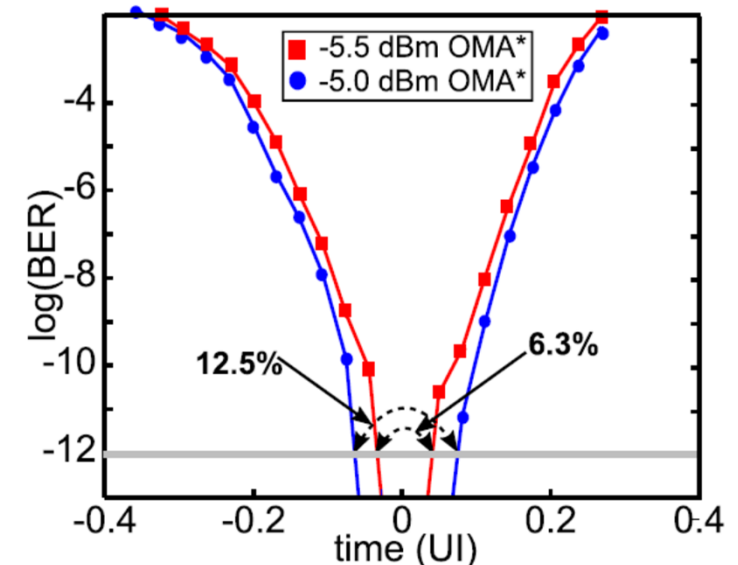
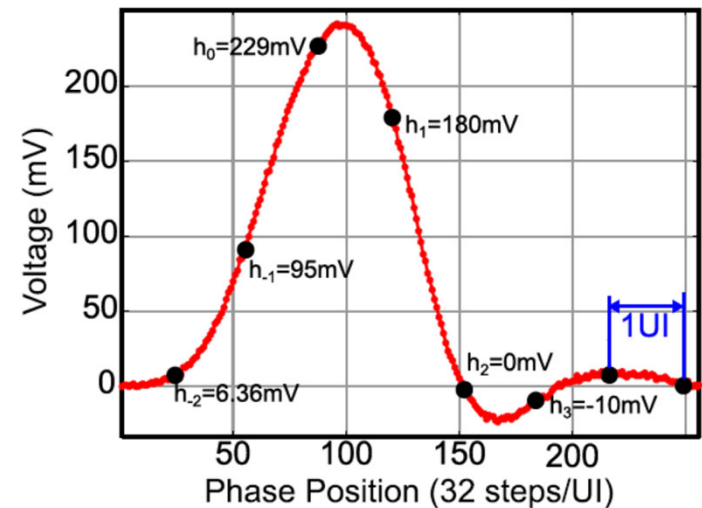
Low-BW TIA & DFE RX

[Ozkaya JSSC 2017]



- Self-referenced TIA is used for differential generation
- Actual 64Gb/s pulse response has a significant pre-cursor ISI tap, which requires a 2-tap TX FFE

64Gb/s Pulse Response & Timing Margin



Next Time

- Main/Limiting Amplifiers