ECEN689: Special Topics in Optical Interconnects Circuits and Systems
Spring 2016

Lecture 5: Transimpedance Amplifiers (TIAs)

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

• HW2 due today
• Exam1 is on Mar 10
  • 2:20-3:45PM (10 extra minutes)
  • Closed book w/ one standard note sheet
  • 8.5”x11” front & back
  • Bring your calculator
  • Covers material through lecture 4
• Reading
  • Sackinger Chapter 5
  • Razavi Chapter 4
Agenda

• Optical Receiver Overview

• Transimpedance Amplifiers
  • Common-Gate TIAs
  • Feedback TIAs
  • Common-Gate & Feedback TIA Combinations
  • Differential TIAs

• Integrating Optical Receivers
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
Transimpedance Amplifier (TIA)

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

Transimpedance $Z_T = \frac{\Delta v_o}{\Delta i_i}$ (Ω)

Also expressed in units of dBΩ by $20 \log|Z_T|$
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]

\[ R_T = R_{in} = R_L \]

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

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

\[
\bar{V}_{n,\text{out}}^2 = \int_0^\infty \bar{I}_n^2 Z_T^2 \, df = \int_0^\infty \frac{4kT}{R_L} \left( \frac{R}{1 + j2\pi f RC} \right)^2 \, df = \frac{kT}{C_D}
\]

\[
\bar{I}_{n,\text{in}}^2 = \frac{\bar{V}_{n,\text{out}}^2}{R_L^2} = \frac{kT}{R_L^2 C_D}
\]

\[
I_{n,\text{in},\text{rms}} = \frac{\sqrt{KT/C_D}}{R_L}
\]
Agenda

• Optical Receiver Overview

• Transimpedance Amplifiers
  • Common-Gate TIAs
  • Feedback TIAs
  • Common-Gate & Feedback TIA Combinations
  • Differential TIAs

• Integrating Optical Receivers
Common-Gate TIA

- Input resistance (input bandwidth) and transimpedance are decoupled

\[ R_T = R_D \]
\[ R_{in} = \frac{r_o + R_D}{1 + (g_m + g_{mb})r_o} \approx \frac{1}{g_m} \]
Common-Gate TIA Frequency Response

Neglecting transistor $r_o$: \[
\frac{v_{out}}{i_{in}} = \frac{R_D}{1 + s \left( \frac{C_{in}}{g_{m1} + g_{mb1}} \right) \left( 1 + s R_D C_{out} \right)}
\]

- Often the input pole may dominate due to large photodiode capacitance ($100 - 500\text{fF}$)

[Razavi]
Common-Gate TIA Noise

- Both the bias current source and RD contribute to the input noise current.
- RD 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.

\[ \overline{I_{n,\text{in}}} = 4kT \left( \frac{2}{3} g_m + \frac{1}{R_D} \right) \left( \frac{A^2}{\text{Hz}} \right) \]

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

Neglecting transistor \( r_o \):
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

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

[Diagram of Regulated Cascode (RGC) TIA]

[Park ESSCI RC 2000]
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

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

\[ A_2 = g_{m2} R_2 \quad A_3 = -g_{m3} R_3 \]
Agenda

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

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

With Finite Bandwidth Amplifier:

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

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

\[ R_T = \frac{A}{A + 1} R_F \]

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

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

\[ R_{in} = \frac{R_F}{A + 1} \]
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

![Butterworth and Bessel responses](image)

2nd-Order TIA Frequency Response

- Butterworth
- Bessel
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:

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

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

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

**Maximum** \( R_T \) \( \leq \frac{A \omega_A}{C_T \omega_{3dB}^2} \)  

[Mohan J SSC 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

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

Assuming that the source follower has an ideal gain of 1

\[
A = g_{m1}R_D \\
R_f = \frac{g_{m1}R_D}{1 + g_{m1}R_D}R_F \\
R_{in} = \frac{R_F}{1 + g_{m1}R_D} \\
R_{out} = \frac{1}{g_{m2}(1 + g_{m1}R_D)}
\]
CMOS Inverter-Based Feedback TIA

- CMOS inverter-based TIAs allow for reduced voltage headroom operation
- Multiple inverter stages in feedback provide higher gain at the cost of reduced stability
- Diode-connected transistor loads allow for high-frequency internal poles
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-Refereed 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
  - Recall the noise bandwidth tables
Input-Refereed 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

\[ i_{sens}^{pp} = 2Q_i^{rms} \]

- To determine the total optical receiver sensitivity, we need to consider the detector noise and responsivity.
Averaged Input-Reflected Noise Current Density

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

\[
\dot{i}_{\text{avg}} = \frac{i_{\text{rms}}}{i_{\text{n,TIA}}} = \frac{i_{\text{n,TIA}}}{\sqrt{BW_{3dB}}}
\]

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

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

\[
I_{n,TIA}^2(f) \over \text{the TIA bandwidth}.
\]
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
  \[ I_{n,D}^2 = 4kT\Gamma g_m \]
  \( \Gamma \) 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

$$i_{n,TIA} \frac{v_{out}}{i_{n,D}} = \frac{v_{out}}{Z_T}$$

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

$$\left( \frac{i_n, D}{i_{n,TIA}} \right)^{-1}$$

$$i_{n, D} = g_m v_{n,TIA} = \frac{g_m i_{n,TIA}}{s C_T + \frac{1}{R_F}} = \frac{g_m R_F}{1 + s R_F C_T} i_{n,TIA}$$

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

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

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

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

Uniform and $f^2$ component!
Total Input-Referred FET Feedback TIA Noise

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

- 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
Common-Gate & Feedback TIA

- 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 gm 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 J SSC 2013]
Agenda

- Optical Receiver Overview
- Transimpedance Amplifiers
  - Common-Gate TIAs
  - Feedback TIAs
  - Common-Gate & Feedback TIA Combinations
  - Differential TIAs
- Integrating Optical Receivers
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 \( C_T = C_D + C_I \)

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

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 \( C_T = C_D + C_I \),

\[
Z_T(s) = \frac{v_{OP} - v_{ON}}{i_i} = \frac{\left( \frac{2A}{A+2} \right)R_F}{1 + \frac{sC_T R_F \frac{A}{2} + 1}{A}}
\]
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
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
  - TIA $I_D \propto \left( R_T C_{IN} \right)^2 f_{3dB}^4$
  - LA $I_D \propto f_{3dB}^2$

\[ V_A = V_{GS1} + V_{GS2} \approx 0.8 \times VDD \]
\[ A \approx g_{m1} R_D = \frac{\alpha(VDD - V_A)}{VOD} \approx \frac{\alpha(0.2 \times VDD)}{VOD} \]
Integrating Receiver Block Diagram

[Emami VLSI 2002]
Demultiplexing Receiver

- Demultiplexing with multiple clock phases allows higher data rate
  - Data Rate = \#Clock Phases \times 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
Residual SA Offset = 1.15mV

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

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

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

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

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

<table>
<thead>
<tr>
<th>Gb/s</th>
<th>$P_{avg}$ (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>-9.8</td>
</tr>
<tr>
<td>16</td>
<td>-7.8</td>
</tr>
</tbody>
</table>
Integrating Receiver Sensitivity

• Test Conditions
  – 8B/10B data patterns (variance of 6 bits)
  – Long runlength data (variance of 10 bits)

• BER < $10^{-10}$

[Palermo J SSC 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]
Next Time

- Main/Limiting Amplifiers