Dynamic Damping in Transimpedance Amplifiers

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***Abstract*—This paper presents a design technique for optical receivers that adjusts the damping factor of a 2nd-order transimpedance amplifier (TIA) synchronously with the incoming data. This approach allows the fast response of low-damping factor while mitigating the intersymbol interference (ISI) associated with underdamped systems. A differential shunt-feedback TIA (SF-TIA) is optimized to reach its minimum input-referred noise, with and without cross-coupled inverters at its output. The negative transconductance introduced at the TIA's output improves bandwidth, noise performance and vertical eye opening (VEO). Dynamic damping is introduced by adding a triode-region transistor with a time-varying bias voltage across the outputs which modulates the damping factor of the system from -0.42 to 3.5 each unit interval. With dynamic damping, the TIA achieves more than twice the VEO compared to the optimized reference TIA.**

***Index Terms*—Optical Receivers, Linear Time Invariant (LTI) systems, Linear Periodic Time Varying (LPTV) systems, Shunt-Feedback Transimpedance Amplifier (SF-TIA), Cross-Coupled Inverters, Vertical Eye Opening (VEO)**

I. Introduction

Optical receivers use a transimpedance amplifier to convert the incoming photocurrent to a voltage. Usually, additional gain stages are used to further amplify the signal before applying it to a decision circuit such as a sense-amplifier based CMOS latch. Such a latch, or its CML counterpart uses positive feedback to regenerate a relatively small input signal (in the range of 10s of mV) into a full-rail logic signal. Although regeneration is known to be a power-efficient means to amplify a signal [1], this scheme still requires a TIA gain of at least 5-10 kΩ. Without sufficient gain in the receiver front-end, the latches will add a power penalty to the receiver's sensitivity above that predicted by only considering the input-referred noise of the front-end.

Conventional optical receivers consist of a stable, linear, or perhaps saturating, yet nonetheless time-invariant front-end. Regeneration is confined to the clocked decision circuit. This paper investigates the use of regeneration embedded in the analog front-end of the optical receiver. Inspired by observations of the dependence of a 2nd-order system's settling behavior on its damping factor, we modulate the damping factor of an inverter-based shunt-feedback TIA, a commonly used TIA. This creates a time-varying system, which for the small input currents generated by the photodiode (PD) is considered to be a *linear* periodically time-varying (LPTV) system.

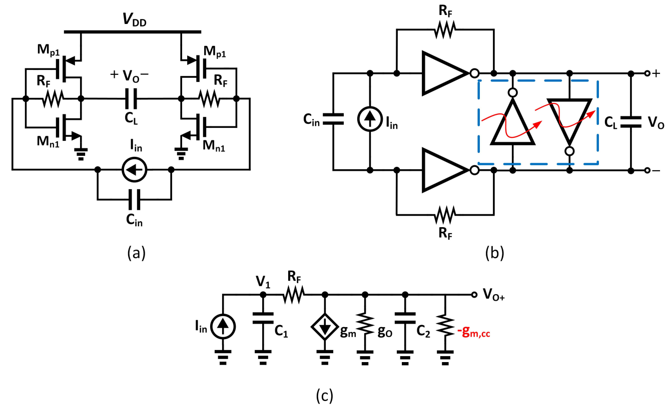

Fig. 1. Fully-differential shunt-feedback TIA (a) transistor-level schematic without cross-coupled inverters (b) high-level schematic with cross-coupled inverters (c) ac small-signal model of the differential half of circuit.

Shunt-feedback TIAs are optimal when designed for a damping factor $\zeta = \frac{1}{\sqrt{2}}$ [2]. However, when the voltage amplifier is implemented using a static-CMOS inverter, due to its low intrinsic gain, higher damping factor often results, leading to slower settling and a reduction in vertical eye opening (VEO). Cross-coupled inverters can be added to the output of a differential TIA increasing the voltage amplifier's gain, allowing the TIA to achieve an optimal damping factor. In this paper we investigate increasing their conductance dynamically and synchronously with the incoming data such that the damping factor is not only reduced to zero and but also becomes negative (i.e., regeneration) during a portion of the unit interval (UI). Dynamic damping factor adjustment in the front-end is shown to improve the effective gain of the front-end as well as improve the overall sensitivity of the front-end.

A static-CMOS inverter-based shunt-feedback TIA, optimized for minimum input-referred noise current serves as a reference design for this investigation. It has an effective gain of less than 2 kΩ. Adding a cross-coupled inverter at the output improves bandwidth and gain of the TIA. However, the effective gain is not more than 3 kΩ. Therefore, additional gain stages are needed which lead to more noise and power dissipation. By using the proposed dynamic damping factor system, the SF-TIA's effective gain increases to more than 5 kΩ thereby mitigating the need for follow-on gain stages.

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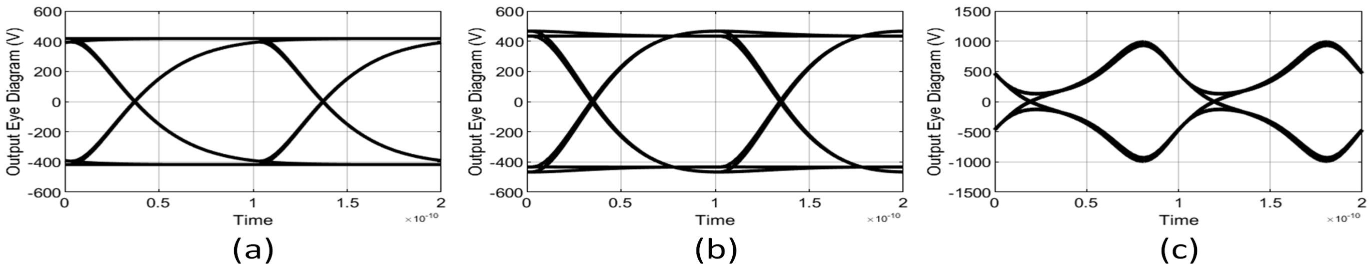

Fig. 2. Output of the shunt-feedback TIA model in SIMULINK, $f_{bit} = 10\ Gb/s$ (a) $A = 0$, $B = 0$ (b) $A = 0$, $B = 1$ (c) $A = 3$, $B = 1.2$, $\phi_{start} = 3.74\ rad$.

This paper is organized as follows: Section II presents analysis and results of a SF-TIA model in simulink with cross-coupled and g_m modulation at the output (LTI and LPTV systems). A transistor-level implementation and simulation results are then provided in Section III. Finally, Section IV summarizes and concludes the paper.

II. Shunt-feedback transimpedance amplifier and variable damping

A differential inverter-based shunt-feedback TIA is shown in Fig. 1 (a). It consists of static-CMOS inverters and feedback resistors R_F. The small-signal model of the half circuit is shown in part (c). Component parameters are as follows:

$$g_m = g_{m,n1} + g_{m,p1} \tag{1}$$

$$g_o = g_{o,n1} + g_{o,p1} \tag{2}$$

$$C_1 = C_{gs,n1} + C_{gs,p1} + 2C_{PD} + C_{pad} \tag{3}$$

$$C_2 = C_{db,n1} + C_{db,p1} + 2C_L \tag{4}$$

where C_{PD} is the capacitance of the PD, C_{pad} is the capacitance of the TIA's input pad and C_L is the load capacitance the TIA drives. C_{gd} of the MOSFETs has been ignored for preliminary modeling. The time-domain differential equations used to develop a SIMULINK model are:

$$i_{C_1} = C_1 \frac{dv_1}{dt} = i_{in} + G_F(v_{o+} - v_1) \tag{5}$$

$$i_{C_2} = C_2 \frac{dv_{o+}}{dt} = G_F(v_1 - v_{o+}) - g_m v_1 - g_o v_{o+} \tag{6}$$

These equations are represented in a behavioral model shown in Fig. 3. Fig. 2 (a) shows output of this model for random input data at $f_{bit} = 10\ Gb/s$, using $R_F = 1\ k\Omega$, $C_{in} = 100\ fF$, $C_L = 25\ fF$ and transistor parameters corresponding to $W_{n1} = W_{p1} = W_1 = 20\ \mu m$ in a 65 nm technology. In this case, $A = B = 0$, thereby disabling the feedback denoted by $g_{m,cc}$. The damping factor of this system depends on several parameters, namely [3]:

$$\zeta = \frac{1}{2} \frac{R_o C_2 + (R_o + R_F) C_1}{\sqrt{R_F R_o C_1 C_2 (1 + g_m R_o)}} \tag{7}$$

where $R_o = 1/g_o$.

For the selected parameters, $\zeta = 0.92$ which is higher than the optimum value of $\frac{1}{\sqrt{2}}$. However, if a circuit technique could lower the MOSFET output conductance down to $\frac{2}{3}$ of its nominal value, the damping factor could be decreased. In low-speed designs, output conductance is decreased by increasing the channel length. However, this increases input capacitance, which is not acceptable in high-speed designs. In a fully-differential design as shown in Fig. 1 (b), cross-coupled inverters can be used to reduce the output conductance as [4]:

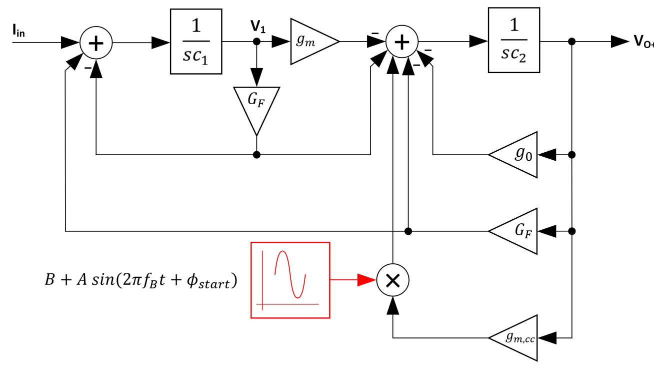

Fig. 3. SIMULINK model of shunt-feedback TIA.

$$g_{o,tot} = g_o - g_{m,cc} \tag{8}$$

where g_o is the output conductance of the main inverters and $g_{m,cc}$ is the transconductance of the cross-coupled inverters. Red arrows indicate proposed dynamic modulation of cross-coupled inverters which is discussed later. For modeling the shunt-feedback TIA with cross-coupled inverters at the output a cross-coupled inverter with conductance -1 m℧ is added in the block diagram of Fig. 3. In this case $A = 0$ and $B = 1$. Output of the SIMULINK model for random input data at $f_{bit} = 10\ Gb/s$ with the cross-coupled inverters is shown in Fig. 2 (b). For an input current amplitude of 1 A, the VEO of the output increased slightly from 800 V to 900 V.

Fig. 2 (c) shows the output of the SIMULINK model of the shunt-feedback TIA with cross-coupled inverters whose g_m is modulated. The values were a dc offset of 1.2, an amplitude of 3 and an initial phase of 3.74 rad. By modulating the damping factor, the VEO doubles to 1800 V. The damping factor ranges from -0.28 to 1.26. The system, no longer time-invariant, is nevertheless linear. The noise performance of this linear, periodically time-varying system is explored through simulation following the design of a transistor-level implementation.

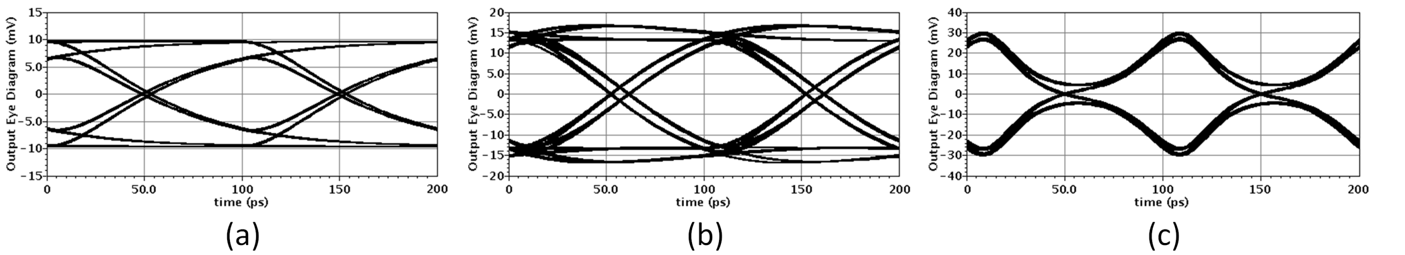

Fig. 4. $f_{bit} = 10\ Gb/s$, Output of the (a) SF-TIA (b) SF-TIA with cross-coupled inverters at the output (c) SF-TIA with g_m modulation at the output, $A = 0.3\ V$, $B = 1.15\ V$, $\phi_{start} = 310°$.

III. Transistor-level design and simulation

Fig. 1 (a) shows the implementation of the fully differential shunt-feedback transimpedance amplifier. The current source connected across the two inputs requires ac-coupling capacitors and offset compensation (not shown) to connect the PD since it usually has a reverse-bias voltage across it. For now, these modifications are not considered.

For random input data with $f_{bit} = 10\ Gb/s$, $C_{in} = 100\ fF$ and $C_L = 25\ fF$, input-referred noise current for different sizes of transistors ($W_{n1} = W_{p1} = W_1$) and different values of R_F for each W_1 was calculated. This calculation was done by referring the output noise voltage to the input using the effective gain of the TIA, following the approach in [5]. Since the TIA may introduce ISI, but has no follow-on equalization, the effective gain is the ratio between the output VEO and the peak-to-peak input signal. The input-referred noise current reaches its minimum value of $0.47\ \mu A_{rms}$ with $W_1 = 23\ \mu m$ and $R_F = 1.15\ k\Omega$. The output eye diagram is shown in Fig. 4 (a). The effective gain is $1.3\ k\Omega$.

Fig. 5 shows the SF-TIA with the cross-coupled inverters and a triode-region MOSFET to realize g_m modulation. Before introducing g_m modulation, a fully-differential SF-TIA with cross-coupled inverters at the output is designed for minimum input-referred noise current. Again $I_{n,in}$ for different sizes of transistors (W_1 and $W_{n2} = W_{p2} = W_2$) and values of R_F was calculated. The input-referred noise reaches its minimum value of $0.37\ \mu A_{rms}$ using $W_1 = 30\ \mu m$, $W_2 = 4.5\ \mu m$ and $R_F = 1.4\ k\Omega$. Fig. 4 (b) shows the output eye diagram with effective gain of $2.49\ k\Omega$.

To explore the performance of the TIA using dynamic damping (g_m modulation), M_r is biased by a sinusoidal signal ($V_b = B + A\ sin(2\pi f_{bit} + \phi_{start})$). Assuming M_r is in triode, the transconductance of the cross-coupled circuit is given by:

$$g = -g_{m,cc} + g_r = -(g_{m,n2} + g_{m,p2}) + \mu_n C_{ox} \frac{W_r}{L} \left[B + A\ sin(2\pi f_{bit} + \phi_{start}) - V_{O,DC} - V_{t,n}\right] \quad (9)$$

Therefore the total conductance of the cross-coupled inverters and M_r is modulated by the sine wave. By selecting appropriate values for W_2, W_r, B, A and ϕ_{start}, the damping factor of the system swings between negative and positive values. The output eye digram of the LPTV system is shown in Fig. 4 (c). Here, $W_1 = 30\ \mu m$, $W_2 = 15\ \mu m$, $W_r = 4\ \mu m$, $R_F = 1.4\ k\Omega$ and $V_b = 1.15 + 0.3\ sin(2\pi f_{bit} + 310°)\ V$. This corresponds to variations in g and ζ along the UI of $-8.9\ m\mho < g < 18.8\ m\mho$ and $-0.42 < \zeta < 3.5$ each unit interval.

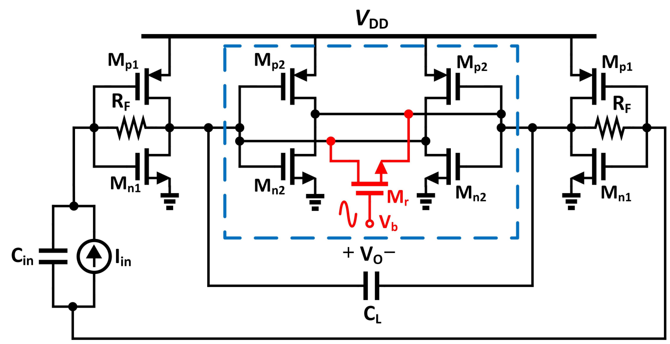

Fig. 5. Fully-differential SF-TIA including cross-coupled inverters and g_m modulation.

For an LPTV system, due to a time varying operating point, the mean-squared noise at the output varies over each UI, rendering conventional ac noise analysis inaccurate. Instead, transient noise analysis is used. The mean-squared output noise is calculated along the UI by:

$$v_n(t) = \sqrt{\frac{\Sigma_{n=0}^{N} v_o^2(t + nT_b)}{N}} \quad (10)$$

where N is the number of UIs used in the estimate, T_b is the bit period and v_o results from a transient noise simulation in Spectre when no input signal is applied. Note that we are assuming that since input and output signals are relatively small the system is still linear. Therefore, the output when we have an input signal will be the sum of the output due to the signal and the output due to noise. Fig. 6 shows the VEO and output noise voltage of the LPTV system in one unit interval. The maximum vertical eye opening gain is $5.24\ k\Omega$ at a sampling time of 0.57 UI. The input-referred noise current at this sampling time is $0.45\ \mu A_{rms}$.

Although the proposed technique of damping factor modulation requires a signal synchronous with the incoming data, the gain of the LPTV system is higher than $5\ k\Omega$ over a range of more than $45°$, allowing generation from a 4-stage differential ring oscillator.

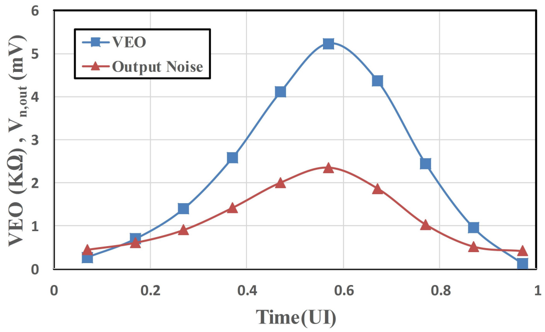

Fig. 6. VEO and output noise voltage of the LPTV system in one UI.

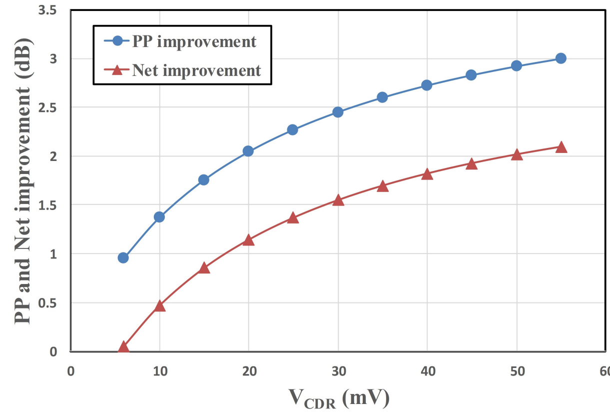

Fig. 7. Power Penalty (PP) and net performance improvement.

Table I summarizes results of simulations. Adding the cross-coupled inverter and M_r increases the output noise voltage. In the SF-TIA with g_m modulation, the input-referred noise current is larger compared to the reference design with cross-coupled inverters but no modulation. However, g_m modulation provides significantly increased VEO gain eliminating the need for extra stages of amplification. The highest value of the Gain/Power ratio is obtained by the LPTV system implying that having a variable damping factor decreases the power dissipation of the overall receiver.

TABLE I
RESULTS SUMMARY

SF-TIA including:	Main inverters	Main and X-coupled inverters	Main and X-coupled inverters with modulation
Size of transistors (μm)	$W_1 = 23$	$W_1 = 30$ $W_2 = 4.5$	$W_1 = 30$ $W_2 = 15$ $W_r = 4$
Data rate (Gb/s)	10	10	10
f_{3dB} (GHz)	3.27	5	—
AC gain (kΩ)	1.93	2.68	—
VEO gain (kΩ)	1.3	2.49	5.24
$I_{n,in}$ (μA_{rms})	0.47	0.37	0.45
Power (mW)	2.65	4.03	5.29
Gain/Power (kΩ/mW)	0.49	0.62	0.99
$i_{pp,min}$ (μA_{pp}) for V_{CDR} = 25 mV	31.3	15.2	11

The sensitivity of a receiver is function of both the input referred noise of the analog circuitry and the minimum voltage swing (V_{CDR}) required at the input of the clock-and-data recovery system that follows the TIA as given by:

$$i_{pp,min} = SNR \times I_{n,in} \times PP \tag{11}$$

where SNR is the signal-to-noise ratio required to achieve the targeted bit-error rate. An $SNR = 14$ is needed for $BER = 10^{-12}$. PP denotes the power penalty incurred by the swing needed at the CDR, given by [6]:

$$PP = 1 + \frac{V_{CDR}}{SNR \times I_{n,in} \times Gain} \tag{12}$$

The LPTV system has an $I_{n,in}$ degradation of 0.9 dB compared to the cross-coupled inverter-based TIA. The net performance improvement in dB is given by the difference between the improvement in PP and the degradation in $I_{n,in}$. Fig. 7 shows the results. Notice that for all but the smallest values of V_{CDR} the proposed technique provides a net improvement.

IV. Conclusion

This work presents damping factor modulation in a 2nd-order system to take advantage of both low and high damping factor systems (high-speed and low ISI). A simple SF-TIA and a SF-TIA with a cross-coupled inverter at the output are designed for minimum input-referred noise current in a 65 nm technology. By adding a time-varying resistor at the output, the damping factor of the system is changed synchronously with the incoming data, giving rise to twice the gain, and a net improvement in receiver sensitivity for reasonable output voltage swing requirements.

One reason for the increased noise in the proposed design is that high damping factor is implemented by shunting the cross-coupled inverters with a conductance. The net conductance is the difference between that of the cross-coupled inverters and M_r. However, their noise contributions add. A lower noise approach would be to directly modulate the cross-coupled inverters. Future research is aimed at designing inverter structures whose transconductance can be modulated without injecting signal current into the outputs of the TIA.

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