

# Design of CMOS Differential Transimpedance Amplifier

Jevgenij Charlamov<sup>1</sup>, Romualdas Navickas<sup>1</sup>

<sup>1</sup>*Department of Computer Engineering, Vilnius Gediminas Technical University,  
Naugarduko St. 41, LT-03227 Vilnius, Lithuania  
jevgenij.charlamov@vgtu.lt*

**Abstract**—This paper explores the design of a differential transimpedance amplifier (TIA) integrated circuit for use in optical time domain reflectometer (OTDR) optical receivers (OR). All dependences between primary TIA design variables including OTDR laser pulse width, OR bandwidth, and TIA feedback resistance, are defined. Then, based on these dependencies, it is shown analytically that to achieve a stable system with no overshoots, the open-loop bandwidth of the fully differential amplifier (FDA) at the core of the TIA must exceed closed-loop TIA bandwidth by a factor of at least four. According to equations that are provided, it is shown that minimum amplifier noise occurs when detector and input transistor parasitic capacitances are equal. Schematics and short descriptions of all important TIA components, including the FDA and beta multiplier-based bias circuits, are given. The FDA has a bandwidth of 900 MHz. With a feedback resistance value of 20 k $\Omega$  the TIA has a closed-loop bandwidth of 125 MHz. TIA power consumption does not exceed 15 mW and at low bandwidths a variable feedback resistance allows the FDA to achieve up to 7 dB higher dynamic range. The area of the TIA integrated circuit is 100  $\mu\text{m} \times 250 \mu\text{m}$  excluding contact pads.

**Index Terms**—Circuit noise, CMOS integrated circuit, optical receiver, transimpedance amplifier.

## I. INTRODUCTION

Current to voltage (I/V) converters are very important building blocks in analog and mixed signal electronic circuits. There are many applications where such conversion is needed and that require optimization of different converter characteristics, as well as different design approaches. One of the possible I/V converter implementations is the transimpedance amplifier (TIA). Numerous papers have been published in past years dealing with the design of TIAs used in optical receivers (OR) for communications, where performance at the highest bandwidth is mostly considered [1]. Another possible TIA application, described in this paper, is in optical receivers used in optical time domain reflectometers (OTDR).

An OTDR is an opto-electronic device widely used to test optical fiber networks. Its operation is based on the Rayleigh light backscatter mechanism [2]. An OTDR launches certain width laser pulses while its optical receiver analyses backscattered light. Two key OTDR characteristics are

strongly related: attenuation dead zone (ADZ) and dynamic range (DR) [3]. Both depend on laser pulsewidth and OR bandwidth. Dynamic range also depends on laser power and optical receiver SNR. An ideal OTDR would have a zero ADZ and infinite DR. This would be achieved in the case of a noiseless optical receiver with infinite bandwidth and delta function type laser pulses. In practice, however, laser power is limited to 1 W by the nonlinear Kerr effect [4]. Therefore, to deliver more energy into the fiber under test, and thus achieve a higher SNR, pulsewidth must be increased. This means either ADZ or DR must be sacrificed to improve the other. In order to get a desired ADZ it is possible to select a given pulsewidth and then calculate the necessary OR bandwidth. One possible criterion for these calculations is that ADZ should exceed pulsewidth by a factor of two. Therefore, a key OTDR design goal is to maximize dynamic range as a function of pulsewidth, or in other words achieve the best SNR as a function of bandwidth. OTDRs commonly use pulsewidths from 5 ns to 10  $\mu\text{s}$ , which require OR bandwidths from 50 kHz to 130 MHz respectively [5].

TIAs used in commercial OTDRs are usually implemented with discrete components, which because of their large parasitic capacitances, limits maximum bandwidth to between 50 and 70 MHz [6]. A TIA implemented as an IC would have significantly lower parasitic capacitances, and therefore could have a higher maximum bandwidth, resulting in a shorter ADZ.

In order for an OR to achieve optimal dynamic range at a given bandwidth, a certain  $R_f$  value must be used. At low bandwidths OR noise is dominated by feedback resistor thermal noise, so increasing  $R_f$  causes OTDR dynamic range to increase by the square root law. At a certain  $R_f$  value defined as optimal, further increases in  $R_f$  are not useful because they result in insignificant increases in SNR. At higher bandwidths transistor noise becomes dominant, so increasing  $R_f$  causes noise and signal to grow at the same rate, resulting in no signal to noise ratio improvement [5]. In practice  $R_f$  is also limited by parasitic capacitances, which limit bandwidth. This problem becomes more critical when a variable gain TIA is needed.

Taking into account all of the factors mentioned above, for bandwidths from 50 kHz to 130 MHz optimal  $R_f$  ranges from 20 k $\Omega$  to 800 k $\Omega$  [6].

High value feedback resistors are very susceptible to

electromagnetic interference (EMI), which in turn can cause measurement errors. In addition to good shielding, the best way to ensure that an OR has high EMI immunity is to use a differential TIA structure. This also results in high power supply and common mode rejection ratios.

Optical receivers generally comprise a photodiode and shunt-shunt feedback transimpedance amplifier. In order to achieve high dynamic range, OTDR optical receivers specifically use an avalanche photodiode because of its internal photo-electronic gain. Therefore, the main noise sources in OTDR optical receivers are feedback resistor noise, avalanche photodiode dark current noise (multiplied and non-multiplied), and input transistor channel thermal and flicker noise [6].

As noted above, because of lower parasitic capacitance, a TIA integrated circuit can achieve higher bandwidths than a TIA implemented using discrete components. For example, CMOS technology can be used to implement a TIA design with switchable feedback resistors and a bandwidth of 130 MHz, which is not possible using discrete components.

This paper summarizes the design of a variable gain, fully differential, transimpedance amplifier integrated circuit where all of the issues outlined above are taken into account.

TIA amplifiers used in OTDRs require large feedback resistance values. So while bipolar transistors have smaller  $1/f$  noise, their use in OTDR TIA designs is not practical because their high input bias currents can result in unacceptable offset voltage levels. This consideration determined the general choice of CMOS technology.

Reduction of IC channel width does not shrink the size of analog ICs at the same rate as digital ICs. However, such scaling increases cost significantly [7]. The specific choice of AMS 0.35  $\mu\text{m}$  CMOS technology was driven by these two competing factors.

## II. TIA DESIGN CONSIDERATIONS

The TIA discussed in this paper has a differential shunt-shunt feedback topology, with feedback resistors connected in series using single pole, double throw (SPDT) switches to reduce input parasitic capacitance (Fig. 1). Each switch connects its input to either the TIA output or to the next resistor. Five resistor pairs are needed to achieve the desired range of feedback resistance values.

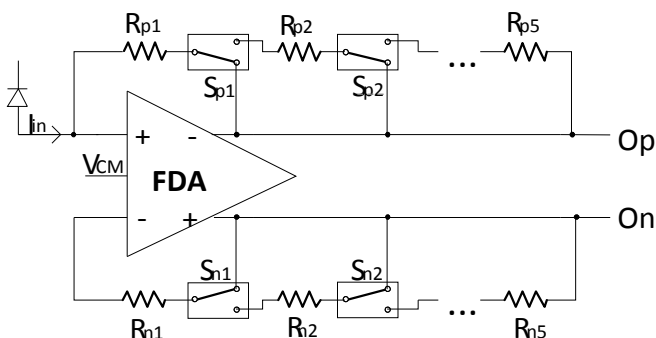


Fig. 1. TIA simplified schematic.

FDA input transistor size is calculated to minimize noise. This calculation begins by determining the ratio between detector capacitance  $C_D$  and transistor input capacitance  $C_{in}$  that results in minimum noise. Because detector capacitance

is usually fixed, this ratio is achieved by scaling input transistor size. Various results have been reported in the literature. One states that minimum noise occurs when detector capacitance is smaller. Another states the opposite opinion. A third says that these capacitances should be equal. However, the most recent research indicates that this ratio depends on a different set of design choices when either gain-bandwidth product or load capacitance is kept constant [8]. In the current case  $R_f$  and bandwidth are constants so input transistor noise plays a major role in determining the  $C_D/C_{in}$  ratio. The total input referred noise current of a shunt-shunt feedback transimpedance amplifier with FET input can be expressed as follows [8]

$$\overline{i_n^2} = \frac{4kT}{R_f} \times BW_n + \frac{4kT\Gamma(2f C_T)^2}{3g_m} \times BW_n^3, \quad (1)$$

where  $\Gamma$  is Ogawa's excess noise factor,  $g_m$  is FET transconductance,  $C_T = C_D + C_{in}$  is total input capacitance, and  $BW_n$  is noise bandwidth. The first term represents resistor noise current, which does not depend on input capacitance.

If we express  $g_m = 2 f_T C_{in}$  then (1) can be rewritten as

$$\overline{i_n^2} = \frac{4kT}{R_f} \times BW_n + \frac{(C_D + C_{in})^2}{C_{in}} \frac{8f kT\Gamma}{3f_T} \times BW_n^3, \quad (2)$$

where  $f_T$  is transistor cut-off frequency.

Taking the derivative of (2) with respect to  $C_{in}$  and setting result to zero reveals that minimum noise occurs when  $C_{in}$  equals  $C_D$ .

The typical capacitance of a reverse-biased, singlemode InGaAs avalanche photodiode with an active area of 30  $\mu\text{m}^2$  is 0.35 pF [9]. In AMS 0.35  $\mu\text{m}$  CMOS technology, an input capacitance value of 0.35 pF can be achieved by setting transistor size to 110  $\mu\text{m}$  by 0.7  $\mu\text{m}$ .

TIA stability is impacted by FDA bandwidth. Because OTDRs make analog signal measurements, TIA stability is more important in OTDRs than in digital communication systems. Ideally, OTDR TIA output pulses have minimum distortion and no overshoots. In control theory [10], percent overshoot ( $P.O.$ ) is defined as

$$P.O. = 100e^{-\zeta' f / \sqrt{1-\zeta'^2}}, \quad (3)$$

where  $\zeta'$  is a dimensionless damping factor. Setting  $\zeta'$  to 1 results in a  $P.O.$  value of zero, meaning that the system is critically damped. If we assume a) that a TIA is a second order system, where the first input pole ( $f_d = 1/2 R_f C_{in}$ ) is dominant, and the second pole ( $f_{nd} = 1/2 R_{out} C_{out}$ ), which is defined by voltage amplifier bandwidth, is non-dominant, and b) that  $R_{out} C_{out} \ll R_f C_{in}$ , then it is possible to write [11]

$$\frac{1}{R_{out} C_{out}} = 4\zeta'^2 \frac{A_0}{R_f C_{in}}. \quad (4)$$

Thus in order to meet a zero  $P.O.$  requirement, amplifier open-loop bandwidth must be at least four times greater than

the desired closed-loop bandwidth. Or in our specific case, to achieve a TIA bandwidth of 130 MHz, FDA bandwidth must be at least 520 MHz.

### III. REALISATION

The FDA described in this paper is implemented using a

two-stage, fully differential voltage amplifier with self-biased topology and an output stage that uses common mode feedback (Fig. 2).

The gain and phase frequency responses (Fig. 3) of the FDA show a unity gain frequency of 900 MHz with current consumption around 4 mA at a supply voltage of 3 V.

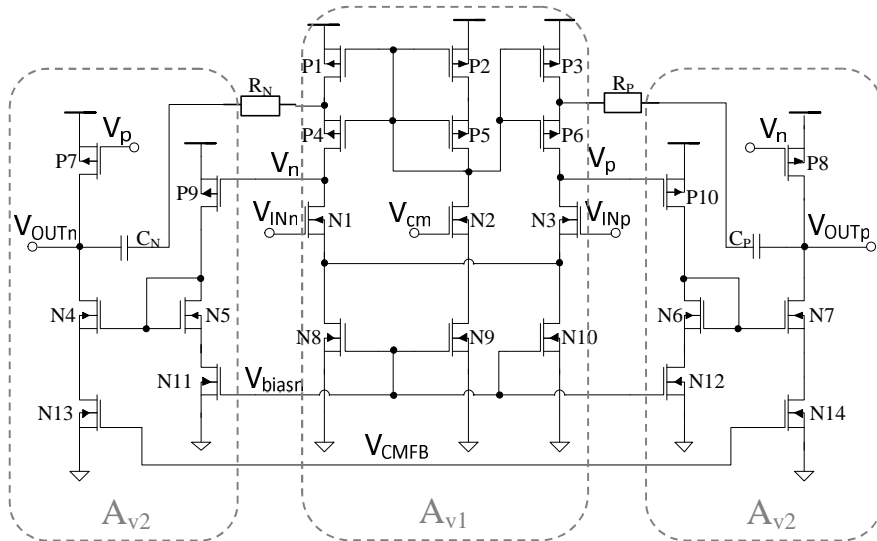


Fig. 2. FDA schematics.

TABLE I. DIFFERENTIAL AMPLIFIER TRANSISTORS DIMENSIONS.

MOSFET	Width, $\mu\text{m}$	Length, $\mu\text{m}$
N1–N3	110	0.7
N4, N7	30	0.35
N5–N6	10	0.35
N8–N12	20	0.7
N13–N14	100	0.35
P1–P6	120	0.7
P7–P8	90	0.35
P9–P10	30	0.35

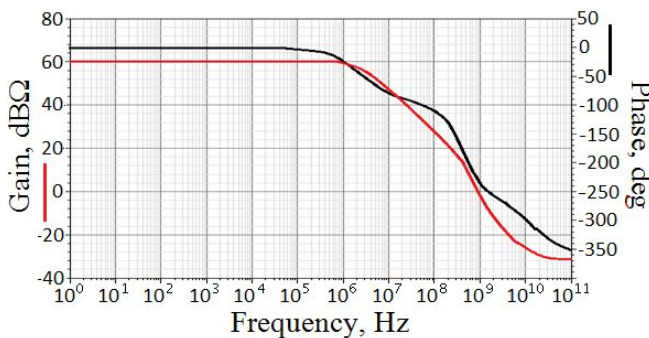


Fig. 3. FDA gain and phase frequency responses.

In order to generate stable differential amplifier tail current, a beta-multiplier based generator was selected (Fig. 4).

This circuit uses positive feedback, so to avoid oscillation gain must be less than 1. The generated current is 116  $\mu\text{A}$  using a 3 V supply and is stable from 2.5 V to 3.5 V.

Current sensitivity to voltage change is only 0.6  $\mu\text{A/V}$  (Fig. 5). The circuit is self-biasing and therefore to avoid an unwanted bias state requires a start-up circuit, which is implemented using the three transistors shown on the left in Fig. 4. Each SPDT switch shown in Fig. 1 is implemented

using a pair of simple transmission gates.

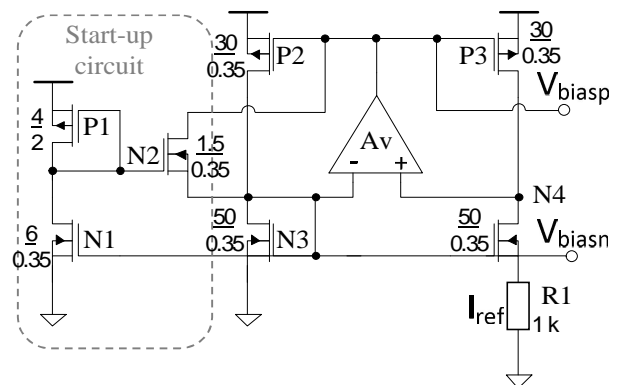


Fig. 4. Beta multiplier based bias voltage generator simplified schematic.

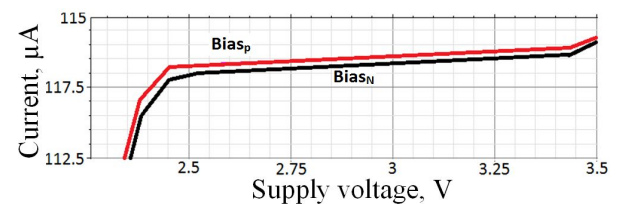


Fig. 5. Generated tail current dependency on supply voltage variations.

### IV. RESULTS

Gain vs. frequency response curves of the designed TIA for different feedback resistance values are shown in Fig. 6. Gains and 3 dB bandwidths for these feedback resistance values are shown in Table II. Bandwidth ranges from 6 MHz at 800 k $\Omega$  to 125 MHz at 20 k $\Omega$ .

Monte-Carlo simulation results (Fig. 7) show a PSRR between positive and negative TIA outputs, that is no worse than 250 dB, at a frequency of 1 kHz.

Equivalent input noise currents, integrated over frequency for different  $R_f$  values are shown in Fig. 8. At low

frequencies, where  $R_f$  defines overall noise, the noise current mostly depends on and is inversely proportional to feedback resistor value. At middle frequencies breakpoints in all curves are observed. Here transistor noise, which does not depend on  $R_f$  value, starts to dominate. At high frequencies all curves converge because overall noise is defined only by transistor noise.

TABLE II. BW AND GAIN VS. FEEDBACK RESISTANCE.

$R_f$ , k	Bandwidth, MHz	Gain, dB
20	125	87.25
50	62	95
100	36	100.8
200	20	106.6
400	10	112.4
800	6	118.3

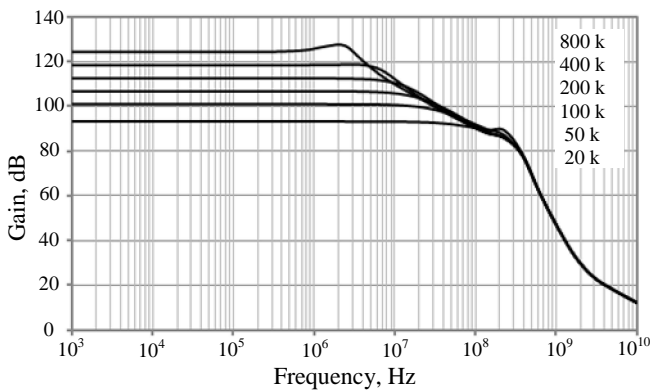


Fig. 6. TIA gain frequency response with different feedback resistances.

Since OTDR measurements must be highly accurate, power supply rejection ratio (PSRR) is another important TIA characteristic.

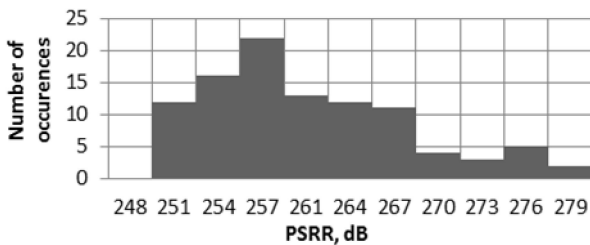


Fig. 7. TIA PSRR worst case simulation results.

Transistor noise does not depend on feedback resistor value, so when it is the dominant noise source, increasing  $R_f$  increases noise and signal equally [5]. It follows then that

above some  $R_f$  value there will be no further increase in TIA signal-to-noise ratio and consequently no further increase in dynamic range. Simulations were performed and plotted to visualize this model of TIA performance (Fig. 9).

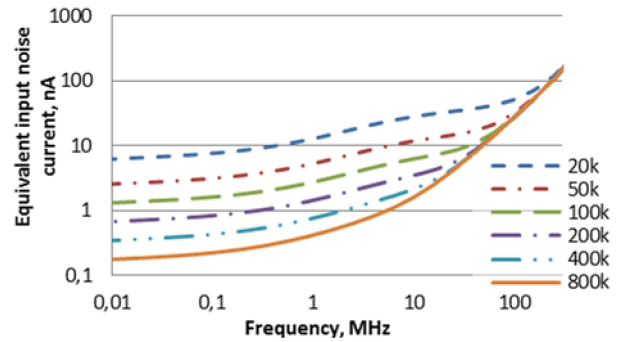


Fig. 8. TIA equivalent input noise current.

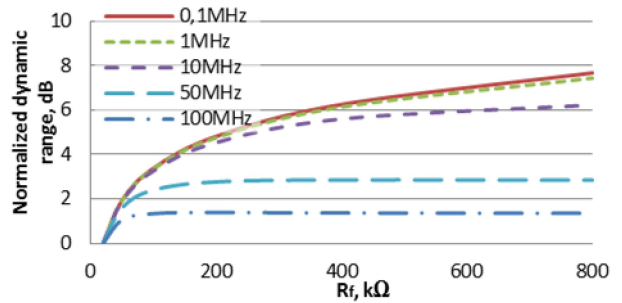


Fig. 9. TIA normalized dynamic range vs. bandwidth.

At a bandwidth of 100 MHz, the simulated TIA Normalized Dynamic Range vs.  $R_f$  curve approximately reaches its maximum value at a feedback resistance value of 50 k . At a bandwidth of 50 MHz, a similar break point occurs when  $R_f$  equals 100 k . At a bandwidth of 10 MHz a breakpoint is observed when  $R_f$  equals about 300 k . However, in the 10 MHz and lower bandwidth curves, growth in dynamic range slows above their break point, but does not approach zero. These simulations further show that at low bandwidths increasing gain in a transimpedance amplifier can extend dynamic range up to about 7 dB. These simulations show good agreement with theoretical calculations [5].

The dimensions of the designed integrated circuit (Fig. 10) are  $100 \times 250 \mu\text{m}$ , excluding contact pads. More than 30 % of the topology area is occupied by feedback resistors.

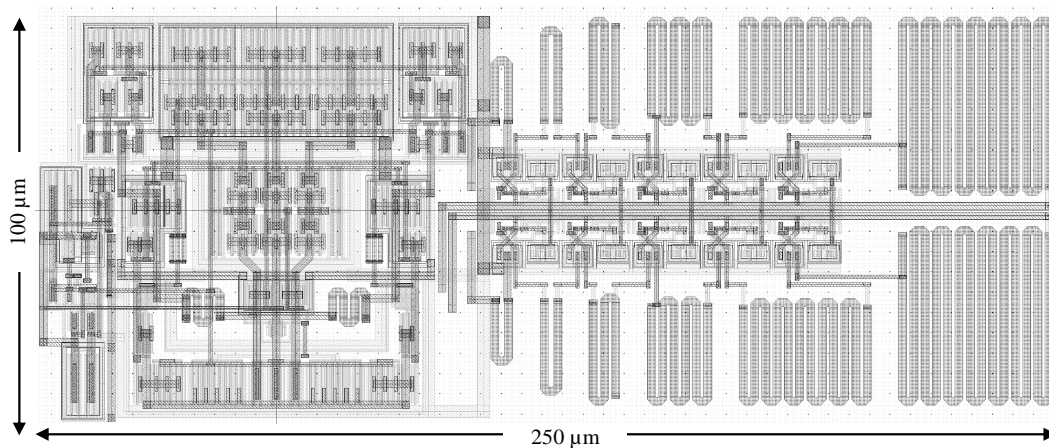


Fig. 10. Designed TIA integrated circuit.

## V. CONCLUSIONS

The design of an IC of variable gain, differential transimpedance amplifier for use in OTDR optical receivers is presented. Analytical expressions show that to achieve minimum noise, photodiode parasitic and transistor capacitances must be equal. Also, zero overshoot is achieved when the open loop bandwidth of the fully differential voltage amplifier is at least four times higher than closed loop TIA bandwidth. Simulation results show that maximum bandwidth of the TIA IC would be about two times greater than that of discrete now used in commercial OTDRs, and equal 125 MHz using a feedback resistance of 20 k $\Omega$  and a designed input parasitic capacitance of 1 pF. The use of five, switchable feedback resistor pairs, ranging in value from 20 k $\Omega$  to 800 k $\Omega$ , allows  $R_f$  to be set to a value that is close to optimal for any given OTDR laser pulse width. Variable gain provides up to 7 dB of dynamic range enhancement at low bandwidths. Power consumption of the IC is about 15 mW, and worst case simulated power supply rejection ratio is 250 dB at 1 kHz. The area of the IC is 0.025 mm<sup>2</sup>, excluding contact pads.

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