

A Fiber Optic Application: Wideband Shunt Feedback Transimpedance Amplifier

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Abstract:

A fiber optic Transimpedance Amplifier (TIA) circuit application is reported. The circuit is realized in MOS/BJT transistor technology. A TIA bandwidth of 319 MHz is achieved with a gain bandwidth product of 1.977 GHz. The transimpedance R_T of this amplifier was found to be 4.04 k Ω which is equivalent to 72.12 dB Ω . A total input-referred noise current of 2.59 pA/ $\sqrt{\text{Hz}}$ was calculated with an input-current overload level of 29.4 μA . This data was realized through the Multisim 2001 software environment in conjunction with mathematical and physical calculations.

Keywords: Transimpedance, TIA.

Introduction:

Fiber optic communication systems require the use of sensitive current-input preamplifiers to condition the electrical signal received in the form of current pulses from a p-i-n diode or avalanche photodiode detectors (APD). A commonly used topology is the transimpedance feedback amplifier whose (relatively) low input impedance and flat gain characteristic is well suited to this application [1-4]. Important specifications for such amplifiers include input current noise, bandwidth, gain flatness and input current overload level [1-4].

For the design of fixed-gain transimpedance amplifiers for these applications, there is a direct trade off between input current noise and input-current overload level via the value of the shunt feedback resistor used in the signal path. For this reason, in applications requiring both large overload levels and low noise (very wide dynamic range), various means have been employed to vary the gain of the amplifier in response to the input signal level [5-8]. The transimpedance front end circuit can be compared to other major front end circuits as follows [9]:

Table 1. Receiver front-end comparison [9]

Connection	Bipolar	FET	High Impedance	Transimpedance
Circuit Complexity	Simple	Simple	Complex	Moderate
Relative Noise	Moderate	Moderate	Very Low	Low
Bandwidth	High	Low	Moderate	High
Dynamic Range	Moderate	Moderate	Narrow	Wide

In this paper, a specific and new design is introduced for the shunt feedback TIA amplifier. However, the general physical and mathematical concepts of such amplifiers is commonly available [10]. The generic format of the shunt feedback TIA amplifier indicates that the photodetector is connected to the input of the inverting voltage amplifier, which has a feedback resistor R_F leading from its output to the input. The photodetector capacitance C_D and the input capacitance C_I appear in parallel (from AC point of view) in which they can be combined as a single capacitance $C_T = C_D + C_I$ [10].

Simulation and Calculation Results

In Fig. 1, the shunt feedback TIA amplifier is implemented with transistors Q_1 through Q_4 . Transistor Q_1 is primarily responsible for gain and after this stage, there is a cascade of three emitter followers ($Q_2 - Q_4$), which buffer and level shifts the output signal. For better input to output isolation, the last stage has two outputs, one for the feedback signal and one for the output signal which is not needed in this case..

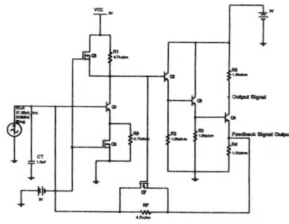


Fig. 1: A schematic of the shunt feedback TIA amplifier.

In the circuit shown in Fig. 1, the photodetector (simulated as a current source) has an input capacitance of 0.5 pF and the amplifier input capacitance is also 0.5 pF. Using Multisim 2001 software environment to simulate the circuit operation, the feedback amplifier has a low frequency gain of 18.6 dB which is around $A = 6.2$ in terms of voltage gain. The transimpedance of the TIA amplifier at low frequencies can therefore be calculated as [10]:

$$R_T = \frac{A}{A+1} \times R_F = \frac{6.2}{7.2} \times 4700 = 4.04 \text{ k}\Omega$$

The value of R_T is lower than but close to the value of the feedback resistor (4.7 k Ω) which is expected. The value of 4.04 k Ω can be transformed into dB Ω unit as $20\log R_T$ and that is equal to 72.12 dB Ω .

Given that $A = 6.2$, $R_F = 4.7 \text{ k}\Omega$ and $C_T = C_D + C_I = 1 \text{ pF}$, the TIA bandwidth BW_{3dB} can therefore be calculated as [10]:

$$BW_{3dB} = \frac{1}{2\pi} \frac{\sqrt{2A(A+1)}}{R_F \cdot C_T} = \frac{1}{6.28} \cdot \frac{\sqrt{2 \times 6.2(6.2+1)}}{4700 \times 1 \times 10^{-12}} = 319 \text{ MHz}$$

Given that a frequency response has been obtained (using Multisim 2001 electronic workbench) ranging from 1 Hz to 2.95 GHz in logarithmic scale as in Fig. 2. The time constant T_A becomes [10]:

$$T_A = \frac{R_F \cdot C_T}{2A} = \frac{4700 \times 1 \times 10^{-12}}{12.4} = 379 \text{ ps}$$

In Fig. 3, a simulated frequency range matches the calculated TIA amplifier bandwidth. However, the curve will intercept the frequency axis at 420 MHz which represent a frequency pole at which the feedback TIA gain becomes zero as calculated below:

$$\frac{1}{2\pi \cdot T_A} = \frac{1}{6.28 \times 379 \times 10^{-12}} = 420 \text{ MHz}$$

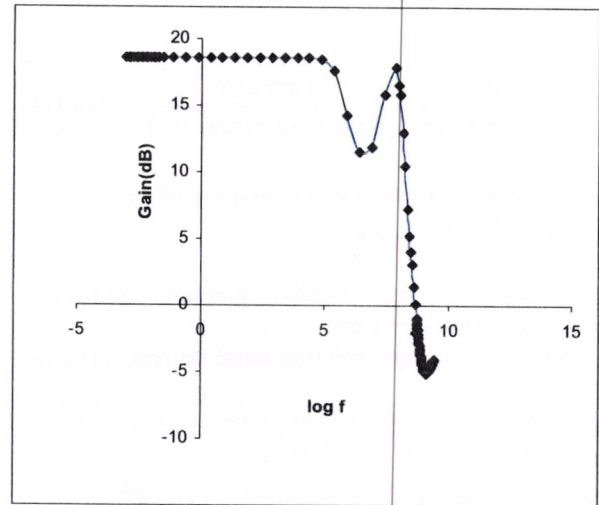


Fig. 2: TIA amplifier gain (dB) versus frequency range on logarithmic scale.

Fig. 3, represents a part of the frequency response reported in Fig. 2, in which the gain is clearly indicated as a function of the bandwidth (in MHz) of the TIA amplifier.

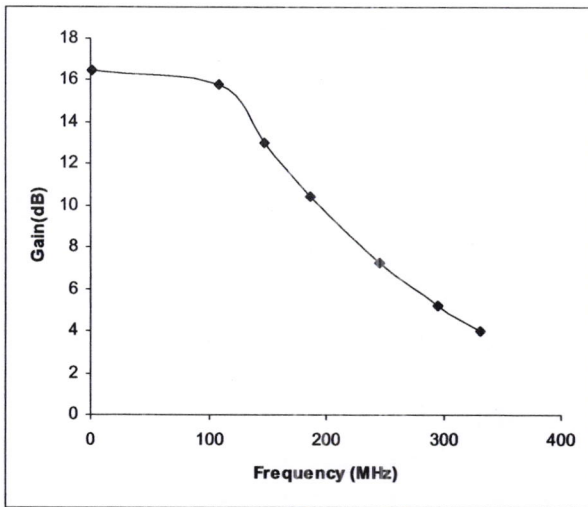


Fig. 3: Gain versus bandwidth of the TIA amplifier simulated using Multisim 2001 software.

A gain bandwidth product which is the area above which the curve stands in Fig. 2 can be established as:

$$A \cdot f_A = 6.2 \times 319 \times 10^6 = 1.977 \text{ GHz}$$

To confirm the value of R_T calculated earlier (4.04 k Ω) in terms of system design, it should be less than or equal to:

$$R_T \leq \frac{A \cdot f_A}{2\pi \cdot C_T \cdot (BW_{3dB})^2} = \frac{1.977 \times 10^9}{6.28 \times 1 \times 10^{-12} (319 \times 10^6)^2} = 4.06 \text{ k}\Omega$$

The TIA's input impedance at low frequencies is:

$$R_i = \frac{R_F}{A+1} = \frac{4700}{7.2} = 652.77 \Omega$$

Now, given a typical bipolar junction transistor parameters $\beta = 100$, $I_C = 1 \text{ mA}$, $R_b = 80 \Omega$, the total input-referred noise current formula is as:

$$\frac{\bar{i}_{nTIA}}{\sqrt{BW_{3dB}}} = \sqrt{\frac{4kT}{R_F} + \frac{2qI_C}{\beta}} = \sqrt{\frac{4 \times 1.381 \times 10^{-23} \times 300}{4700} + \frac{2 \times 1 \times 10^{-19} \times 1 \times 10^{-3}}{100}}$$

$$\frac{\bar{i}_{nTIA}}{\sqrt{BW_{3dB}}} = \sqrt{6.72 \times 10^{-24}} = 2.59 \times 10^{-12} = 2.59 \frac{\text{pA}}{\sqrt{\text{Hz}}}$$

As for the input overload current (peak to peak), it can be established as the maximum permissible voltage swing (from simulated data) per the input resistance R_i :

$$i_{ovl}^{pp} = \frac{V_i(\max)}{R_i} = \frac{19.2 \text{ mV}}{652.77 \Omega} = 29.4 \mu\text{A}$$

There is however, a phase angle Φ between the input and the output signals and is shown in relation to the frequency range of the TIA amplifier bandwidth in Fig. 4. In Fig. 5, the relationship between the phase change angle Φ and the gain in dB.

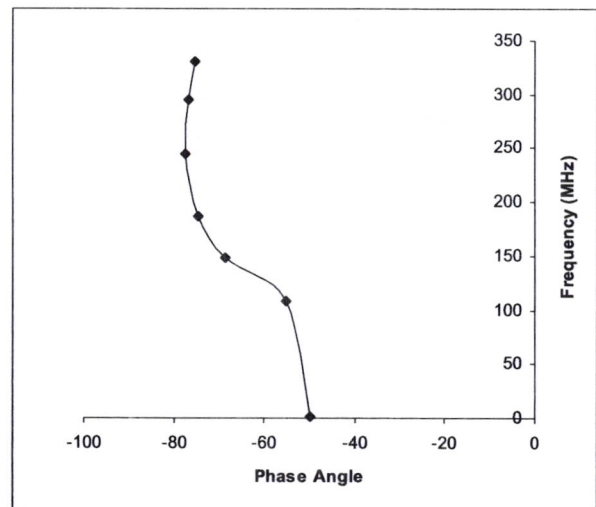


Fig. 4: Phase angle versus TIA amplifier bandwidth.

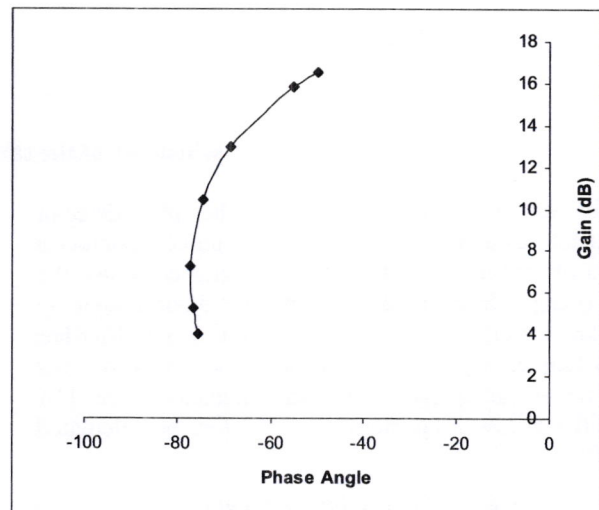


Fig. 5: Phase angle versus TIA amplifier gain (dB).

In comparison between data published in this work and the BiCMOS transimpedance amplifier done by Meyer and Mack [11], the following is established in Table 2.:

Table 2.

	This Work	Meyer and Mack [11]
TIA Bandwidth	319 MHz	128 MHz
Input-Referred Current Noise	2.59 pA/ $\sqrt{\text{Hz}}$	1.17 pA/ $\sqrt{\text{Hz}}$
Input Resistance R_i	652.77 Ω	240 Ω
Gain Bandwidth Product	1.977 GHz	-
Input-Current	29.4 μA	20 μA
Overload Level		

Discussion

It is important first to discuss the calculated TIA bandwidth (319 MHz) for the circuit in Fig. 1 in more details, to start with, the voltage gain obtained at 319 MHz is around 4.3 dB from simulated results which is about 1.6 in linear voltage gain. That means that the linear voltage gain is above unity (one) which makes the gain at this frequency quite useful. However, the range of 319 MHz starts from 1MHz at 1.65 dB of gain. It is also very important to point out that there is almost a perfect agreement between calculated bandwidth of 319 MHz and simulated data of Fig. 3.

In comparison with the work published by Meyer and Mack [11], the TIA bandwidth is further extended by a factor of 2.49. On the other hand, the input-referred current noise compared to the scale of bandwidth involved is improved despite the fact that it is 2.59 pA/√Hz compared to 1.17 pA/√Hz by Meyer and Mack [11]. A trade-off always exists between the input-referred current noise and the amount of input-referred overload current which means that if one of these currents is increased, the other one will decrease, hence an optimum level is the best option for this case. A 29.4 μA of input-current overload level is achieved which is also an improvement to that of 20 μA done by Meyer and Mack [11]. In effect, a wide dynamic range has been obtained in this work which is one of the important characteristics of this work. This is achieved by the very low input capacitance of 1 pF with a shunt feedback set up. The value of the transimpedance obtained in the simulated circuit of Fig. 1, which is 4.04 kΩ indicated a very low input impedance of 652.77 Ω which is exactly why this form of transimpedance is useful given the large bandwidth and gain bandwidth product that were achieved.

It can be seen from Fig. 1, that the signal path is implemented with bipolar junction transistors (BJT), while the bias network is implemented with MOS transistors. This difference in transistor use takes advantage of the BJT's superior high frequency performance and accurate matching and as for the MOS transistors use, it is because of their high output impedance, virtual zero gate current and low noise which is clearly very important for this application.

If we assume that the single pole angular frequency ω_o for the TIA amplifier is:

$$\omega_o = \sqrt{\frac{A+1}{R_F \cdot C_T \cdot T_A}} = \sqrt{\frac{7.2}{4700 \times 1 \times 10^{-12} \times 379 \times 10^{-12}}} = 2.01 \text{ GHz}$$

For the above data, the bandwidth according to a butterworth frequency response is equal to:

References:

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$$BW_{3dB} \leq \frac{\omega_o}{2\pi} = \frac{2.01 \times 10^9}{6.28} = 319 \text{ MHz}$$

This result confirms earlier calculation of BW_{3dB} . To achieve near ideal passband (without peaking, i.e near flat) for this TIA amplifier, the feedback amplifier bandwidth must be larger than:

$$f_A \geq \frac{2A}{2\pi \cdot R_F \cdot C_T} = \frac{2 \times 6.2}{6.28 \times 4700 \times 1 \times 10^{-12}} = 419 \text{ MHz}$$

To reach the value of 419 MHz, a new technology is needed to realize an amplifier with the gain bandwidth product of:

$$A \cdot f_A = 6.2 \times 419 \times 10^6 = 2.597 \text{ GHz}$$

There are some design difficulties related to the transimpedance limit with present technology (without using faster systems). Getting a 419 MHz bandwidth with a gain bandwidth product of 2.597 GHz would require reducing the value of the feedback resistor and hence reducing the transimpedance of the amplifier. That may cause higher input-referred current noise and lower input-current overload level. In effect, it would mean narrower dynamic range.

In Fig. 4, phase angles give good indications as to where it can be selected for particular applications in relation to the TIA bandwidth, while the range of phase angles indicated in Fig. 5, represent an important factor at different levels of TIA amplifier gain.

Conclusion:

A wideband shunt feedback transimpedance amplifier can be achieved, in which a TIA bandwidth of 319 MHz with only 2.59 pA/√Hz of input-referred noise current has been reached. An optimum level of input-current overload level off 29.4 μA can be obtained in a trade-off with the input-referred noise current. A transimpedance amount of 4.04 kΩ represents a value at which the TIA's amplifier functions in a way that makes a gain bandwidth product of 1.977 GHz. In conclusion, a wideband, low noise and moderate transimpedance is achievable in a shunt feedback TIA amplifier configuration.

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الملخص

في هذا البحث تطبيق لمكبر ذو الممانعة البينية (Transimpedance) والمستخدم في منظومات الاليف البصرية . هذه الدائرة تم استحداثها باستخدام مجموعة متكونة من ترانزستورات BJT مع ترانزستورات MOS . تم الحصول على عرض حزمة استقبال اشارة الدخل (لمكبر ذو الممانعة البينية) بقيمة 319 ميكا هيرتز و كذلك بقيمة 1.977 كيكا هيرتز لعامل ضرب عرض الحزمة في الربح . أما بالنسبة لممانعة الدائرة المعرفة بما بين الممانعة فقد وجدت على أنها تساوي 4.04 كيلو أوم وبما يكافئ 72.12 ديسبل أوم . بالإضافة الى هذا فقد تم حساب تيار ضوضاء الدخل (المرجع) وقد وجد بأنه يساوي 2.59 بيكو أمبير لكل هيرتز (تحت الجذر) بينما تم الحصول على مستوى تيار الدخل (ما فوق الحمل) وكان 29.4 مايكرو أمبير. هذه المعلومات أعلاه تم انجازها بواسطة استخدام نظام المحاكاة (ملتيزيم 2001) لنمذجة الدوائر الالكترونية اضافة الى حسابات فيزيائية ورياضية .