Design of Variable Gain Transimpedance Preamplifier

by

Mohammad Saud Al-Juaid

A Thesis Presented to the

FACULTY OF THE COLLEGE OF GRADUATE STUDIES
KING FAHD UNIVERSITY OF PETROLEUM & MINERALS
DHAHRAN, SAUDI ARABIA

In Partial Fulfillment of the Requirements for the Degree of

MASTER OF SCIENCE

In

ELECTRICAL ENGINEERING

December, 1998
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DESIGN OF VARIABLE GAIN

TRANSIMPEDANCE PREAMPLIFIER

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MOHAMMAD SAUD AL-JUAIMD

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DECEMBER 1998
This thesis, written by Muhammad Saud Al-Juaid, under the direction of his Thesis Advisor and approved by his Thesis Committee, has been presented to and accepted by the Dean of the College of Graduate Studies, in partial fulfillment of the requirements for the degree of

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To my family & relatives,

To Mutlaq Al-Ghamdi & friends,

and to whoever supports me
ACKNOWLEDGEMENT

I am grateful to King Fahd University Petroleum and Minerals and especially for Electrical Engineering Department and College of Graduate Studies for support of this work.

I wish to express my appreciation and gratitude to Dr. C. B. Yahya, the Thesis Advisor, for his continuous and unlimited support and guidance throughout the thesis work.

I also wish to acknowledge other Thesis Committee members: Dr. H. A. Ragheb, Dr. M. A. Al-Sunaid and Dr. H. A. Al-Jamid for their advice and cooperation.
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ABSTRACT

Name: Muhammad Saud Al-Juaid
Title: Design of Variable Gain Transimpedance Preamplifier
Major Field: Electrical Engineering
Date of Degree: December 1998

This thesis discusses the design of a variable gain transimpedance preamplifier with high sensitivity, high gain, large bandwidth and wide dynamic range. Different preamplifier configurations are evaluated and compared. The Automatic Gain Control (AGC) concept is introduced. Also, the design of feedback based AGC for transimpedance preamplifier is provided. The common base front-end transimpedance preamplifier is studied in depth. Solutions are proposed and simulated to solve a DC instability of this preamplifier against the change in the feedback resistance. An optimized variable gain transimpedance preamplifier with common base input stage is proposed. The usage of Current Conveyor (CC) as a transimpedance preamplifier is investigated. A programmable CC based transimpedance preamplifier is proposed. Finally, an AGC for CC based programmable transimpedance preamplifier is also proposed.

Master of Science Degree
King Fahd University of Petroleum and Minerals
Dhahran, Saudi Arabia
December 1998
خلاصه الرسالة

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عنوان الرسالة: تصميم مضخم تحويل التيار إلى جهد ذو المقدرة الكهربائية المتغيرة

الخصص: الهندسة الكهربائية

تاريخ الشهادة: ديسمبر 1998

هذه الرسالة تناقش مضخم تحويل التيار إلى جهد ذو المقدرة الكهربائية المتغيرة. دوائر متعددة قمت وقومنت في هذا المجال. تعرضت الرسالة كذلك لمبدأ التكبير المتغير وتشخيص الأوتوماتيكي في مقدار التكبير. بعد ذلك نوقشت المضخم ذو القاعدة المشتركة، حيث عرضت مشكلة عدم استقرارية عمله عند محاولة تغيير مقدار تكبيره. حاولنا هذه المشكلة اقترحنا واعتبرنا برامج كمبيوتر متخصصة، ولقد روعي في تصميم هذا المضخم تحسين الخصائص المرغوبة الأخرى.

واخبرنا بذلك هذه الرسالة إمكانية استخدام دوائر ناقل التيار في هذا المجال، فقد اقترحنا واعتبرنا عدة دوائر مع مراعاة إمكانية تنفيذ التحكم الأوتوماتيكي لمقدار التكبير.

درجة الماجستير في العلوم

جامعة الملك فهد للبترول والمعادن

الظهران المملكة العربية السعودية

ديسمبر 1998
CHAPTER 1:

INTRODUCTION

In some electronic systems, the input signal is very weak. So, there should be a special circuit able to detect it, amplify it and make it usable for latter stages and processes. This circuit is called a preamplifier. The ability of detecting a signal is determined by the circuit sensitivity. To increase the circuit sensitivity then its noise contribution should be minimized. Also, the preamplifier should have a very high gain to amplify the weak signal to the acceptable level for a next stage. In some applications, the level of the input signal varies which may increase to a limit where a distorted output signal will be generated due to the high gain applied to it. The range between the lowest and the highest signal that the preamplifier can handle is known as the dynamic range. Wide dynamic range is essential requirements in many applications. Moreover, another important specification of the circuit is its bandwidth,
which can be generally defined as the range of frequencies the circuit can efficiently process.

A preamplifier with high gain, high sensitivity, wide dynamic range and large bandwidth is highly required in many applications. An important example of those applications is the optical communication system.

1.1 Optical Communication Systems:

Communication may be described as the transfer of information from a source to a destination over a transmission medium. A communication system is usually required when information is to be conveyed over any distance. Within a communication system, the information will be superimposed or modulated on to an electromagnetic wave acting as a carrier for the information signal. If the carrier is selected from the optical range of frequencies, the communication is called an optical communication. The transmission medium in this type of communications can be an optical fiber. [1]

Optical fiber communication offers extremely attractive features and advantages over the conventional electrical communication. Some of those features and advantages are listed next.

1. **Very High Bandwidth**: In optical communication, the optical carrier frequency is in the range of $10^{13}$ to $10^{16}$ Hz that yields an enormous potential transmission
bandwidth. While this available bandwidth is not yet fully utilized, a modulation at several GHz over a hundred-kilometers is possible without repeaters.

2. **Small Size and Weight:** The diameter of an optical fiber does not exceed the diameter of a human hair. Even when optical fibers are covered with protected coatings, they are very small and light. Because of that, there is an expansion to use optical fibers for signal transmission within aircraft, satellites and even ships. Also, they are used within cities to reduce duct congestion.

3. **Electrical Isolation:** Optical fibers do not exhibit earth loop and interference problems because they are fabricated from glass or a plastic polymer which are electrical insulators. Moreover, they are safe to be used in electrically hazardous environments.

4. **Immunity to Interference and Crosstalk:** Optical signals within fibers are not affected by being transmitted through an electrically noisy environment. Because it is easy to ensure that there is no optical interference between fibers, the crosstalk is negligible.

5. **Signal Security:** Any attempt to obtain optical signal from the fiber will be detected because no way to obtain the signal without drawing the optical power from the fiber. Moreover, the light within optical fibers does not radiate significantly. The above mentioned two features provide a high degree of signal security within optical fiber communications.

6. **Low Transmission Loss:** Optical fibers technology is producing optical fiber cables with very low attenuation or transmission loss which allow transmission for long distances without intermediate electronics (repeaters).

7. **System Reliability and Ease of Maintenance:** The common predicted lifetimes of optical components are of 20 to 30 years which make optical fiber communications
reliable. The reliability beside the low loss property, mentioned above, will reduce the maintenance time and cost for optical fiber communication systems.

![Optical Fiber Communication System](image)

**Fig 1.1: Optical fiber communication system**

Fig 1.1 shows the main blocks in the optical fiber communication system. The information source provides an electrical signal to the transmitter which electrically drives an optical source to modulate electrical signals over a lightwave carrier. The optical source provides the electrical-optical conversion. It may be either a Laser Diode (LD) or Light Emitting Diode (LED). The modulated optical signal travels via an optical fiber cable to a receiver at the other end. The receiver consists of an optical detector driving a further electrical stage and hence providing demodulation of the optical signals. Photodiodes (p-n, p-i-n, or avalanche) and, in some cases, phototransistors and photoconductors are used to detect optical signals and modulate them to electrical signals.
More detailed structure of the optical receiver is shown in Fig 1.2. The optical
detector block, as mentioned earlier, converts the received optical signal into electrical
current that is to be amplified to suitable signal level. The preamplifier provides initial
amplification and it is essential at this stage to keep additional noise to a minimum to
avoid corruption of the received signal. So, the preamplifier configuration and design,
which will be the main objective of this thesis, are major factors in determining the
receiver sensitivity. The main amplifier provides additional low noise amplification
for the signal. The equalizer stage is added to compensate for any nonlinear behavior
or distortion, caused by amplification stages. The final block in the receiver, i.e. the
filter, is to maximize the received signal to noise ratio while preserving the essential
features of the signal. [1]

Fig 1.2: Major elements of an optical fiber receiver.

1.2 Preamplifier Configurations:

For the preamplifier stage, one of the following configurations may be used: low
impedance, high impedance or transimpedance front end.
1.2.1 Low Impedance Front End Preamplifier:

Fig 1.3: Low impedance front-end preamplifier.

Low impedance front-end preamplifier is the simplest preamplifier structure. It consists of a voltage amplifier with low effective input resistance ($R_{in}$). The detector bias resistor ($R_{bias}$) and $R_{in}$ are minimized to obtain the optimum bandwidth. Unfortunately, an increase in the thermal noise will be introduced due to low values of resistors. [1]

1.2.2 High Impedance Front End Preamplifier:

The high impedance front end provides the optimum noise performance due to the use of high $R_{in}$ and $R_{bias}$. This is achieved at the expense of frequency performance. Moreover, it suffers from two drawbacks. First, it needs an equalization stage to compensate for distortion introduced to the signal because it is integrated over a large time constant. Second, it has low dynamic range due to the building of charges from low frequency signals on the input capacitance causing premature saturation of the amplifier at high input signal levels. [1]
1.2.3 Transimpedance Front End Preamplifier:

The transimpedance front-end preamplifier exhibits low noise performance with acceptable bandwidth and dynamic range. By applying a negative feedback via $R_f$, it overcomes the drawbacks of the high input impedance and operates as a current mode amplifier [1]. Its noise performance can be improved by increasing $R_f$ but it should be under the awareness of its effect on bandwidth and dynamic range. These conflicting performance aspects can be compromised based on the application requirements. In general, the transimpedance preamplifier is preferred in wide applications. [1]
1.3 Literature Survey:

Over the past two decades, excessive work was done to implement practical transimpedance preamplifiers and enhance their characteristics. Circuit design and device technologies are the major two factors where the improvement has occurred.

In the literature survey, the followings area were covered:

2. Dynamic Range Enhancement Techniques and circuits such as Automatic Gain Control (AGC) and Variable Gain Amplifier (VGA) [12-18].
3. FET-Based Transimpedance Preamplifiers [19-31].
4. Photo-Detector Technologies [32-44].

The Following subsections highlight major contributions in the first two areas. FET-Based transimpedance preamplifiers and Photo-detectors will not be covered in this thesis.

1.3.1 Bipolar Transistor Based Transimpedance Preamplifiers:

In 1986, Meyer proposed a wide-band low-noise monolithic transimpedance amplifier [2], see Fig. 1.6. Its first stage is a common emitter that has a Miller effect to create a dominant pole and make the circuit insensitive to the photodetector capacitance. Shunt-series feedback resistance over several stages was used to enable the use of a large $R_f$ which will reduce the input current noise. While a transimpedance gain of
6.6kΩ was obtained it was at the expense of the bandwidth which is around 200MHz.

The circuit was realized using silicon bipolar monolithic technology with a cutoff frequency \(f_T\) of 6GHz. A single power supply was used to bias the circuit.

![Circuit Diagram](image)

*Fig 1.6: Common emitter transimpedance preamplifier.*

After about a year, Wilson proposed a transimpedance preamplifier with a very low input resistance [3]. He used a common base configuration as the first stage, which has a very low input resistance. So, the preamplifier was insensitive to the photodiode capacitance and the frequency response was improved. The drawback of a common base is its low noise performance due to its unity current gain. In addition, two power supplies are used to bias the preamplifier. A bandwidth of 200MHz was achieved by using a bipolar transistor with \(f_T=2.6\)GHz.

The circuit proposed by Soda in 1992 [4] was very similar to Meyer's [2] but a shunt-shunt feedback resistance over a single stage was used. The good results achieved in
this preamplifier were based on using heterojunction bipolar transistor (HBT) technology with $f_T=41\text{GHz}$ and high transconductance. Also, the low value of $R_f$, $300\text{\Omega}$, contributed in getting a bandwidth of $12.7\text{GHz}$ but both gain and noise were not optimized. The reported transimpedance gain is $49\text{dB\Omega} (282\text{\Omega})$ and a sensitivity of $-15\text{dBm}$. The receiver sensitivity can be defined as the average optical power required for a bit error rate of $10^9$ which relates to the input-referred current noise. The circuit was biased by a single power supply.

There is a tradeoff in choosing the value of $R_f$ between the noise and gain performance on one hand and the bandwidth from the other side. Low noise and high gain require high $R_f$ while low $R_f$ is preferable for wide bandwidth. So, $R_f$ should be properly selected to compromise between them to reach desired applications’ requirements. As an alternative solution, Suzaki claimed that a compensation method could be used to improve the bandwidth while allowing high $R_f$ [6]. He used dual feedback $R_f1$ and $R_f2$ with a phase adjustment $C_p$ in parallel with $R_f2$. $R_f2$ and $C_p$ values should be carefully chosen to achieve the desired results. The transimpedance preamplifier was implemented by a silicon bipolar technology with $f_T=40\text{GHz}$. Two emitter followers were used as an output stage to achieve larger dynamic range without degrading bandwidth. A bandwidth of $11.2\text{GHz}$ was reported.

The use of a common base configuration in the preamplifier appeared again in the circuit proposed by Vanisri and Toumazou on 1995 [6]. They removed the collector resistance to eliminate its noise contribution and its associated pole. The circuit stability was improved but its biasing became sensitive to the value of $R_f$. A cascode
stage was added to amplify the signal and reduce the Miller effect. The devise technology used was GaAs HBT with $f_T=50\text{GHz}$. A transimpedance gain of $0.966\,k\Omega$ and a bandwidth of $10\text{GHz}$ were achieved at $R_f=1.2\,k\Omega$. The referred input current noise was $28\text{pA}/\sqrt{\text{Hz}}$ at $10\text{GHz}$. They argued that the common base preamplifier is better in transimpedance gain and noise performance than a common emitter preamplifier having the same bandwidth. That is because a common base will allow the use of a larger $R_f$ than a common emitter at the same bandwidth. The authors also proposed some other circuits using BiCMOS with $f_T =10\text{GHz}$ but they anticipated lower performance.

1.3.2 Dynamic Range Enhancement:

For a fixed gain preamplifier, as the input signal increases the output signal will increase until a certain limit at which saturation at the output will occur and a distorted output will be produced. The range between the lowest detectable input signal and the highest allowable input signal is called the dynamic range. The lower limit is determined by the amplifier noise performance, which was carefully considered in the above papers. Reducing the transimpedance gain can increase the other extreme, the maximum allowable input signal, but it is not preferable for low-level input signals. So, the bottom line is to have a high transimpedance gain for low input signals and low transimpedance gain for high input signals. In this case the transimpedance gain will not be fixed but it should be reduced as the input signal
increases. This idea is implemented with different techniques, like Automatic Gain Control (AGC) and current switching, used to widen the dynamic range.

With the participation of his team, Yamashita introduced a variable gain amplifier in 1986 [12]. They varied the open loop gain to increase the dynamic range, Fig. 1.7 (A). A differential amplifier stage was used to implement the circuit, Fig. 1.7 (B). So, by varying the input voltage of the differential amplifier (i.e. gain control) the gain can be changed. They considered the basic two transistors of a differential amplifier as two parallel amplifiers with $A_1$ and $A_2$ gains.

$$A_1 \propto K I R_L$$

$$A_2 \propto (1-K) I R_L$$

where $R_L = R_1 = R_2$. $K$ is proportional to the gain control voltage. This voltage is produced from an AGC control circuit detecting the level of the output signal. Dynamic range of 33dB was achieved for the transimpedance preamplifier by using this circuit.
Fig. 1.7 (A): Preamplifier with AGC: Block Diagram.

Fig. 1.7 (B): Preamplifier with AGC: Schematic diagram.
In 1988, Pietruszynski and others proposed an optical receiver circuit that contained many blocks including the preamplifier and its AGC [13]. The AGC was implemented by shunting the excess input current away from the first amplification stage of the preamplifier through $R_I$, Fig. 1.8. So, the total open loop gain will be reduced, for example, if the preamplifier contains three amplification stages with $A_0, A_1,$ and $A_2$. The overall open loop gain in normal operation will be $A_0A_1A_2$. While if a large input signal is detected, the overall open loop gain will be reduced to $A_1A_2$ by passing the first stage. A peak detector is required to produce a voltage controlling $R_I$ according to the level of the output signal.

![Fig. 1.8: Transimpedance preamplifier with shunting path ($R_I$)](image)

Widening the dynamic range by shunting a large input current to the output appeared again in Broeke’s and Nieuwkerk’s paper in 1993 [14]. It was implemented by using a current switch controlled by a voltage coming from a peak detector to determine whether the input signal will go through an amplification stage or to be bypassed to the output. A dynamic range of 73dB was reported for the proposed receiver.
Meyer and Mack introduced a transimpedance preamplifier with an AGC loop in 1994 [15] and a complementary work on 1996 [16]. The AGC loop was based on varying $R_f$ value according to the level of the output voltage detected by a peak detector. Since the transimpedance gain is approximately equal to $R_f$, especially for high open loop gain, so $R_f$ can control the transimpedance gain. As the input signal is increased $R_f$ should be decreased to avoid distortion at the output. In the frequency response, a decrease in $R_f$ value will shift the dominant pole towards the next pole and then a peak will be generated. A correction action was implemented by changing the value of the resistors controlling the other poles’ locations according to the change in $R_f$.

Khourramabadi and his team in their transimpedance preamplifier, Fig. 1.9, proposed on 1995 [17], implemented the same concept. Their circuit was much simpler and exhibited better results. Input currents ranged from few microamperes to hundreds of microamperes.

Fig. 1.9: Preamplifier with adaptive Transimpedance Gain.
In 1997, Yamazaki proposed a preamplifier with a dynamic range larger than 36dB [18]. It was based on a current bypass concept. It was implemented with a BJT differential pair.

1.4 Thesis Objective:

Based on the above literature survey, the following observations can be highlighted:

- There was no thorough comparative study of the various transimpedance preamplifier circuit configurations.
- There was no attempt to design a variable gain with a front-end common base transimpedance preamplifier.
- While Current Conveyor (CC) is an attractive current mode circuit offering valuable advantages in speed, bandwidth and gain [45-49], there was no study that investigates the potential of CC as a transimpedance preamplifier.

Consequently, the thesis work will include the following:

- Four different BJT transimpedance preamplifier configurations will be evaluated and compared based on their simulation results.
- The Automatic Gain Control (AGC) circuit will be addressed for the transimpedance preamplifier. Simple $R_f$-Based AGC will be designed and simulated to show the concept.
- The common base transimpedance preamplifier will be thoroughly studied to solve its bias sensitivity to $R_f$ variation. Solutions will be proposed and $R_f$ effect on
small signal will be investigated as well. Moreover, an optimized variable gain circuit will be designed and simulated.

- The use of Current Conveyor (CC) as a transimpedance preamplifier will be explored. Various CC-based transimpedance preamplifiers will be proposed and their performance will be simulated and evaluated. Also, the use of AGC for CC-based transimpedance preamplifier will be covered.
CHAPTER 2

PERFORMANCE EVALUATION OF VARIOUS
TRANSIMPEDEANCE PREAMPLIFIER
CONFIGURATIONS

In this chapter, four basic BJT transimpedance preamplifier configurations will be evaluated and compared. Those configurations are as follows:

1. Common Base Front-End Transimpedance Preamplifier [3].
2. Common Emitter Front End Transimpedance Preamplifier [2].
3. Cascode Front End Transimpedance Preamplifier.
4. Cascade Front End Transimpedance Preamplifier [15,16].

The evaluation is basically based on SPICE simulations. To achieve an objective evaluation, the same transistor and photo-detector models are used in all simulations.
Bandwidth, transimpedance gain, sensitivity (noise) and dynamic range are the main
criteria used in comparison. Circuits introduced under each configuration are designed for the same bandwidth, which will make the noise comparison easier.

2.1 Simulation Accuracy:

It is important to firstly validate the accuracy of the simulation and decide if it will be an acceptable reference of evaluations and decisions or not. A common base circuit with its experimental result [9] was selected from the literature, see Fig 2.1. This circuit was simulated with the available SPICE models. Simulation results are shown below.

![Common base front-end transimpedance preamplifier](image)

Fig 2.1(A): Common base front-end transimpedance preamplifier [9].
Fig 2.1(B): Experimental results for the circuit in Fig 2.1(A) [9].

Fig 2.2: SPICE simulation results.

Comparing the two figures, Fig 2.1 (B) and Fig 2.2, it is obvious that the simulation results highly matches the experimental. The bandwidth difference is due to the
frequency limitations of the used models. This issue is considered and highlighted over this report.

To highlight the effect of the transistor model on the frequency performance of the circuit, two models were chosen for comparison. The parameters for those models are included in the Appendix. Simulation results of the aforementioned circuit using both models are compared in Fig. 2.3. As shown in the figure, the two simulations coincide at low frequencies and differ at higher frequencies. It is due the effect of the transistor model on frequency performance of the circuit. So, the simulated frequency responses of the proposed circuits in this thesis are believed to be much better using sophisticated models.

![Graph showing frequency response](image)

Fig 2.3: SPICE simulation results with two different transistors' models.

In general, the spice simulations offer good estimation of the circuit performance and will be used through this thesis to support analysis and conclusions.
2.2 Common Base Front-End Transimpedance Preamplifier:

In a common base configuration, see Fig 2.3, the input signal is fed into the emitter node and the output is taken from the collector. A good aspect of the common base is its low input resistance. It will shift the pole associated with the photodetector internal capacitance to higher frequencies making the bandwidth insensitive to the photodetector’s capacitance. The unity current gain of the common base is a drawback from the sensitivity perspective. Moreover, it has lower transimpedance gain than other configurations that will be introduced later.

Fig 2.3: The Common Base Front End Transimpedance Preamplifier.

Fig 2.4 shows SPICE simulation results for the common base configuration. It exhibits a transimpedance gain of 2.84KΩ and a bandwidth of 300MHz for $R_f=15KΩ$. Its input referred current noise varies from 3.95 to 4.26pA/√Hz over its bandwidth. As
shown in Fig 2.5, for small $R_f < 10\, \text{k}\Omega$, Q2 will be saturated (Q1 is off or near to be off). For $R_f > 15\, \text{k}\Omega$, the circuit operates as a preamplifier but its gain is negligibly changes with $R_f$ variations. That is due to the low open loop gain ($R_{CI}$) compared to the value of $R_f$. This subject will be investigated in details in Chapter 4.

![Fig 2.4: Simulated transimpedance gain and noise of the common base configuration.](image)

![Fig 2.5: Simulated transimpedance gain of the common base preamplifier at different $R_f$ values.](image)
2.3 Common Emitter Front-End Transimpedance Preamplifier:

The common emitter configuration, see Fig 2.6, offers a greater gain than the common base. On the other hand, its frequency performance suffers from a Miller effect of the internal capacitance between the collector and the base \( (C_p) \). As a result, \((1+g_mR_1)\) is added to the effective input capacitance, which may dramatically degrade the bandwidth.

![Circuit Diagram](image)

*Fig 2.6: The common emitter front-end transimpedance preamplifier.*

To achieve a 300MHz bandwidth, \( R_f \) of 7.5KΩ is used. The transimpedance gain of this common emitter preamplifier is approximately equal to the value of \( R_f \), as shown in Fig 2.7. The input referred current noise varies from 1.75 to 2.13pA/√Hz over the bandwidth. A flat transimpedance characteristic is achievable as \( R_f \) exceeds 3KΩ. As indicated in Fig 2.8, the increase in \( R_f \) will cause an increase in the transimpedance gain on the expense of the bandwidth.
Fig 2.7: Simulated transimpedance gain and noise of the common emitter configuration.

Fig 2.8: Simulated transimpedance gain of the common emitter preamplifier at different $R_f$ values.

2.4 Cascode Front-End Transimpedance Preamplifier:

The cascode configuration combines common emitter and common base configurations to reduce the Miller effect and obtain high gain simultaneously. The
load resistance seen by the common emitter, i.e. Q₁ in Fig 2.9, is rₑ₂, which is very small. So, the reflected effect of Cᵦ to input will be small.

\[(1 + gₘ₁ rₑ₂) Cᵦ₁ ≈ 2 Cᵦ₁ \]  \hspace{2cm} (2.1)

where \( gₘ₁ = gₘ₂ ≈ 1/rₑ₁ \) \hspace{2cm} (2.2)

The diode-connected transistors Q₅, Q₆ and Q₇ are added to bias Q₂. To improve the dynamic range, transistor Q₈ is used to drop the DC voltage at the collector of Q₂ by \( V_{be} \).

![Fig 2.9: The cascode front-end transimpedance preamplifier.](image)

A bandwidth of 300MHz and a transimpedance gain of 17.7KΩ are achievable with \( R_f = 21KΩ \). The input referred current noise reaches 2.36pA/√Hz at the bandwidth upper limit. To avoid peaking in the transimpedance characteristic, \( R_f \) should be greater or equal to 19KΩ. While the transimpedance gain increases with the increase in \( R_f \), the bandwidth decreases.
Fig 2.10: Simulated transimpedance gain and noise of the cascode configuration.

Fig 2.11: Simulated transimpedance gain of the cascode preamplifier at different $R_f$ values.
2.5 Cascade Front-End Transimpedance Preamplifier:

The combination of a common collector and common emitter forms the cascade configuration. The common collector exhibits an excellent frequency performance due to the absence of the Miller effect. Its low voltage gain raises the need for the common emitter. While the common emitter, Q2 in Fig 2.12, suffers from the Miller effect, it is isolated from the input stage by the common collector Q1, which reduces the effective resistance at the base of Q2. [50]

![Circuit Diagram](image)

Fig 2.12: The cascade front-end transimpedance preamplifier.

For $R_f=31.5\,\text{K}\Omega$, a 300MHz bandwidth with a transimpedance of 30.46KΩ is obtained.

A range of input referred current noise varies from 1.45 to 1.92pA/√Hz over the bandwidth. An acceptable flat transimpedance characteristic is achievable for $R_f$ greater than 29KΩ. In general, the transimpedance gain is approximately equal to $R_f$ and the bandwidth decreases as $R_f$ increases.
Fig 2.13: Simulated transimpedance gain and noise of the cascade configuration.

Fig 2.14: Simulated transimpedance gain of the cascade preamplifier at different $R_f$ values.
2.6 Summary:

Table 2.1 summarizes the SPICE simulation results for these four configurations. The cascade configuration provides the highest gain-bandwidth product, which is the figure of merit for the preamplifier. Moreover, it has a better noise performance among other configurations. The narrow dynamic range of the cascade configuration is due to its high gain. In general, the cascade has been adopted in many practical transimpedance applications due to the aforementioned advantages [15,16, 51-53].

<table>
<thead>
<tr>
<th>Table 2.1: SPICE Simulation Results Summary</th>
</tr>
</thead>
<tbody>
<tr>
<td>Common</td>
</tr>
<tr>
<td>Base</td>
</tr>
<tr>
<td>Bandwidth</td>
</tr>
<tr>
<td>Transimpedance Gain</td>
</tr>
<tr>
<td>2.84KΩ</td>
</tr>
<tr>
<td>Noise</td>
</tr>
<tr>
<td>Dynamic Range</td>
</tr>
<tr>
<td>Power Dissipation</td>
</tr>
<tr>
<td>24mW</td>
</tr>
<tr>
<td>At Rf</td>
</tr>
</tbody>
</table>
CHAPTER 3

AUTOMATIC GAIN CONTROL FOR
TRANIMPEDANCE PREAMPLIFIER

3.1 Automatic Gain Control:

The level of the input signal differs due to of many factors. It may be a weak signal that should be amplified by high factor to be within useful output level. On the other hand, less amplification factor is needed for a strong signal. Applying high amplification factor to such a signal may cause a distorted signal to be generated at the output. To solve this conflict, an automatic gain control (AGC) should be added to the circuit. The AGC will control the gain based on the level of the input (or output) signal.

Since the output signal equals the input signal multiplied by a gain:

\[ V_o = T_Z I_{in} \]  \hspace{1cm} (3.1)
where $T_Z$ is the gain of the transimpedance preamplifier. So, AGC can be introduced in the following ways:

- Shunting high input signal over all or some amplification stages.
- Reducing the gain factor as the input (or output) signal becomes high.

The second technique can be triggered when input (or output) signals exceed a threshold value or it can be implemented to operate all the time simulating the following equation:

$$T_Z \propto \frac{1}{V_O}$$  \hspace{1cm} (3.2)

The circuit that will be introduced here is based on the latter technique.

Since a transimpedance preamplifier has a gain approximately equals to $R_f$, AGC will be implemented to control the effective $R_f$.

$$V_O = T_Z I_S$$

$$T_Z = R_f$$  \hspace{1cm} (3.3)

$$R_f \propto \frac{1}{V_O}$$  \hspace{1cm} (3.4)

If the input signal increases, the output will increase and may reach to saturation level. To correct it, the value of the gain ($=R_f$) should be reduced. Fig 3.1 indicates the main components in the AGC.
The peak detector detects the output signal level and sets the value of the variable resistance to a proper value. As $V_O$ increases, $R_{VAR}$ should decrease and hence the value of the effective feedback resistance. The coming subsections will describe the circuit design and functionality of the variable resistance and the peak detector.

### 3.2 Variable Resistance:

One way to implement the variable resistance is using NMOS. Figures 3.2 and 3.3 show the circuit and characteristic of an NMOS variable resistance.
The following conditions should be met to use NMOS as an active resistor [50]:

\[ V_{GS} > V_t \quad \text{to switch NMOS on} \]  
\[ \begin{align*} 
V_{DS} &< V_{GS} - V_t & \quad \text{to operated in the triode region} 
\end{align*} \]

The value of the active resistor can be calculated by the following formula:

\[ R_{DS} = \frac{1}{K[2(V_{GS} - V_t) - V_{DS}]} \]
where: \( K = \frac{1}{2} \mu_C \frac{W}{L} \)

The above formula indicates that \( VGS \) can control the value of \( R_{DS} \). The simulation results below emphasize this statement.

While \( V_{DS} \) has a small effect on the value of \( R_{DS} \), but it will be significant if the total harmonic distortion (THD) is considered, as shown next.
\[ R_{DS} = \frac{1}{K[2(V_{GS} - V_t) - V_{DS}]} \]
\[ 2(V_{GS} - V_t) = Z \]
\[ R_{DS} = \frac{1}{K[Z - V_{DS}]} \]
\[ R_{DS} = \frac{1}{KZ[1 - \frac{V_{DS}}{Z}]} \]
\[ \therefore \left| \frac{V_{DS}}{Z} \right| < 1 \]
\[ \frac{1}{1 - X} = \sum_{n=0}^{\infty} X^n, \text{ for } |X| < 1 \]
\[ \Rightarrow R_{DS} = \frac{1}{KZ} \sum_{n=0}^{\infty} \left( \frac{V_{DS}}{Z} \right)^n \] \hspace{1cm} (3.8)

Take \( V_{DS} = V\sin(wt) \)

\[ R_{DS} = \frac{1}{KZ} \sum_{n=0}^{\infty} \left( \frac{V}{Z} \sin(wt) \right)^n \]
\[ = \frac{1}{KZ} \left[ 1 + \frac{V}{Z} \sin(wt) + \left( \frac{V}{Z} \right)^2 \sin^2(wt) + \left( \frac{V}{Z} \right)^3 \sin^3(wt) + \ldots \right] \] \hspace{1cm} (3.9)
\[ = \frac{1}{KZ} \left[ 1 + \frac{V}{Z} \sin(wt) + \frac{1}{2} \left( \frac{V}{Z} \right)^2 (1 - \cos(2wt)) + \frac{1}{4} \left( \frac{V}{Z} \right)^3 (3\sin(wt) - \sin(3wt)) + \ldots \right] \]

Since \( \frac{1}{4} \frac{V^2}{Z^2} << \frac{V}{Z} \), so its contribution to the fundamental component will be ignored.

From the above formula, the second and third harmonics are as follows:

\[ HD_2 = \frac{\frac{V^2}{Z^2}}{\frac{V}{Z}} = \frac{1}{2} \frac{V}{Z} \]
\[ HD_3 = \frac{\frac{V^3}{Z^3}}{\frac{V}{Z}} = \frac{1}{4} \frac{V^3}{Z^2} \]

Since \( \frac{V}{Z} < 1 \), as the order of harmonic distortion increases, its contribution to the total harmonic distortion decreases.
In practical communication systems, the total harmonic distortion should be less than 1%. The following conditions should be met to achieve this level:

\[
THD = \sqrt{HD_2^2 + HD_3^2 + HD_4^2 + ...}
\]  

(3.10)

The second harmonic distortion should be less than 1% because it is the main contributor to the total harmonic distortion.

\[
\Rightarrow HD_2 < 0.01
\]

\[
\Rightarrow \frac{1}{2}V < 0.01
\]

\[
\Rightarrow Z > 50V
\]

\[
\Rightarrow 2(V_{GS} - V_i) > 50V
\]  

(3.11)

For a circuit with a 5V power supply and an NMOS with \( V_i = 1V \), the amplitude of \( V_{DS} \) should be less than

\[
\frac{2(5-1)}{50} = 160mV
\]

If the harmonic distortion is not considered to this extent and instead the engineering sense applied for the formula

\[
2(V_{GS} - V_i) >> V
\]

\[
\Rightarrow 2(V_{GS} - V_i) \geq 10V
\]  

(3.12)

With the previous assumptions, the amplitude of \( V_{DS} \) should be less than 800mV.

If there is a DC component with \( V_{DS} \), it should be considered with \( Z \).

\[
\Rightarrow 2(V_{GS} - V_i) - V_{DS} > 50V
\]  

(3.13)
3.3 Transimpedance Preamplifier with Variable Resistance:

Connecting the variable resistance in parallel with $R_f$, see Fig 3.5, will control the effective feedback resistance and hence the preamplifier's gain. The drawback of changing the effective $R_f$ is the expected introduction of a peak especially if the second pole is near the dominant pole. As shown earlier the common emitter provides a wide range over which $R_f$ can be varied without introducing a peak. So, a variable resistance for this configuration and its simulation results will be included here.

![Fig 3.5: Transimpedance preamplifier with a variable feedback resistance.](image_url)

The range over which $R_{ds}$ can be varied is controlled by the allowable range for $(V_{gs} - V_t)$. To make the value of the $R_{ds}$ independent of $V_{ds}$, $(V_{gs} - V_t)$ should be much greater than $V_{ds}$. 


\[ R_{DS} = \frac{1}{K[2(V_{GS} - V_i) - V_{DS}]} \]
\[ \approx \frac{1}{K[2(V_{GS} - V_i)]} \]  
\[ (3.14) \]

To decrease the value of \( R_{DS} \) by a ratio, \((V_{GS} - V_i)\) should be increased by the same ratio.

Assume \((V_{GS} - V_i) = 500 \text{mV}\)

To make \( R_{DS} = 30 \text{K}\Omega \)

\( 30 \text{K}\Omega = \frac{1}{K[2(500m)]} \)

\[ \Rightarrow K = 33.33 \mu \text{A/V}^2 \]

For \( \frac{1}{2} \mu_nCox = 10 \mu \text{A/V}^2 \)

\[ \frac{W}{L} = 3.333 \text{ take } W = 10 \mu \text{m and } L = 3 \mu \text{m}. \]

If the value of \((V_{GS} - V_i)\) increases to 5V, \( R_{DS} \) will decrease to 3K\Omega accordingly. If a single 5V-power supply is to be used, \((V_{GS} - V_i)\) will be less than 5V.

From results in Fig 3.6, two things can be verified:

1. \( V_{GS} \) will control \( R_{DS} \), which is in parallel with \( R_f \), and hence the overall transimpedance gain. If the \( V_G \) is set by an external source, then this circuit can be considered as programmable transimpedance preamplifier.

2. The effective feedback resistance can be changed over a range without introducing a peak that means no need for auxiliary circuits to be added to eliminate the peak.
If the THD is considered, the conditions indicated equation 3.11 must be met to keep THD less than 1%. The following table addresses this point.

<table>
<thead>
<tr>
<th>$V_{GS}$</th>
<th>$V_{DS}$</th>
<th>THD</th>
<th>$2(V_{GS}-V_t)&gt;50V_{DS}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.23V</td>
<td>230mV</td>
<td>3.6%</td>
<td>No</td>
</tr>
<tr>
<td>2.23V</td>
<td>125mV</td>
<td>1.2%</td>
<td>No</td>
</tr>
<tr>
<td>3.23V</td>
<td>88mV</td>
<td>0.67%</td>
<td>Yes</td>
</tr>
<tr>
<td>3.73V</td>
<td>76mV</td>
<td>0.56%</td>
<td>Yes</td>
</tr>
</tbody>
</table>

3.4 Peak Detector:

In the above variable gain control circuit, the gain is controlled by the value of $V_{GS}$. In automatic gain control the value of the $V_{GS}$ should be automatically set based on the
level of the output voltage. So, a peak detector is needed to implement an AGC. The peak detector is a circuit that will detect the peak value of the input signal and produce a constant voltage accordingly. The simplest form of a peak detector consists of a diode and a capacitor with load resistance [50] but the 0.7 V needed to bias the diode may hide most of the output signal especially if it is weak. To solve this problem, a differential pair will be used in the peak detector circuit as shown in Fig 3.7.

![Fig 3.7: Peak detector circuit.](image)

The differential pair will function as a rectifier where the $V_{b2}$ should equal to the DC component of the input signal.

The common emitter stage is used to correct for the negative sign in the relation between the input and the output of the differential pair. The value of the capacitor should be chosen to meet the following condition:

The discharge time constant

$$RC >> T \quad (3.15)$$
Where $T$ is the period of the sinusoidal signal ($T$: is to be chosen for the practically minimum signal period in optical communication). The discharge resistance in the above circuit is equal the parallel combination of the input resistance of the NOMS and the output resistance of the buffer stage (QF6).

$$R = R_G \!//\! r_e$$

(3.16)

Since $R_G$ is very high and $r_e$ very small so the discharge resistance will be approximately equal to $r_e$. This means a large value of capacitance should be used to meet the discharging requirement. An external capacitor can be used for this purpose because large value of capacitance is impractical for integration. Another alternative solution is to connect a small capacitor with a high impedance node.

The charge time should not necessarily be much smaller than the period because the circuit should have enough time to settle and it is not expected to instantaneously follow the input signal level. The following simulation result shows the performance of the peak detector of Fig 3.8.
3.5 Transimpedance Preamplifier with AGC:

The three parts, i.e. the transimpedance preamplifier, variable resistance and peak detector, are linked together producing a transimpedance preamplifier with the AGC functionality, as shown in Fig 3.9. Its simulation results are provided in Fig 4.10. If the input signal increases from 2μA to 200μA, the automatic gain control circuit will act accordingly to reduce the effective feedback resistance from 8.1KΩ to 2.2KΩ. To insure the final circuit will work properly, the dc voltage at the output node of the preamplifier should match the reference voltage at the differential pair. Moreover, the voltage produced at the output of the peak detector should be within the desired range of $V_{GS}$. 
Fig 3.9: Transimpedance preamplifier with an AGC circuit.

Fig 3.10 (A): Simulation results at $I_{sp}$=2μA.
Fig 3.10 (B): Simulation Results at $I_{sp} = 200 \mu A$. 
CHAPTER 4:

VARIABLE GAIN COMMON BASE TRANSMIMPEDANCE PREAMPLIFIER

4.1 Introduction

As mentioned earlier in this thesis, the CB transimpedance preamplifier has a very low input resistance that will make the frequency response insensitive to the photodetector capacitance. Moreover, the common base has no miller effect since its base is grounded. On the other hand, it does not accept a wide range of $R_f$ values. So, $R_f$ can't be effectively used to control the transimpedance gain. That is because $R_f$ contributes to the circuit biasing and changing its value will dramatically affect the circuit operation. In addition, the common base has not been well investigated by authors compared to common emitter and cascade.
Due to the above, the common base front-end transimpedance preamplifier will be studied in this chapter. The effect of $R_f$ on biasing the CB circuit will be investigated. In addition, a solution will be proposed to make the circuit probably operating over a wide range of $R_f$. Such a solution is needed to effectively implement a variable gain or AGC for CB transimpedance preamplifier. The effect of $R_f$ on small signal performance will be investigated as well. At the end of this chapter, an optimized circuit will be proposed.

4.2 $R_f$ and DC biasing of CB Transimpedance Preamplifier:

![CB Transimpedance Preamplifier Circuit Diagram](Fig 4.1: CB Transimpedance Preamplifier)

From Fig 2.4, the overall gain of the CB transimpedance preamplifier is not equal to $R_f$ and negligibly changed with its value, provided the circuit in operation mode. That is because of the low open loop gain as mentioned earlier. To solve this issue an additional gain stage should be added as shown in Fig 4.1 [3].
This circuit has a critical problem preventing it to be used as a variable gain transimpedance preamplifier. This problem is the circuit DC instability due to the variation in the feedback resistance, as shown in Fig 4.2.

Fig 4.2 (A): Transimpedance gain over $R_f$ range ($R_{EF}=6\text{K}\Omega$, $R_{CF}=10\text{K}\Omega$, $R_{EF}=1\text{K}\Omega$, $R_{CF}=3\text{K}\Omega$, $R_{EF}=0.5\text{K}\Omega$, $C_f=10\text{fF}$, $V_{EE}=-5\text{V}$, $V_{CC}=5\text{V}$)

Fig 4.2 (B): Transimpedance gain vs. $R_f$. 
Problem Definition:

The current following through $R_{E1}$, i.e. is constant and determined by the value of $V_{EE}, R_{E1}$ and $V_{E1}$ which is constant and approximately equal to 0.8V. This current is equal to the some of the currents following through $Q_1$ and $R_f$. So, if there is a change in $R_f$ the current biasing $Q_1$ will change accordingly.

For High $R_f$

If $R_f$ is increased, then the current following through it will decrease. Since $I_{RE1}$ is constant so $Q_1$ will compensate for the decrease. In other words, the current following through the emitter of $Q_1$, i.e. $I_{E1}$ will increase.

Taking

$$I_{C1} \approx I_E$$

$$\therefore V_{C1} = V_{CC} - I_{C1} R_{C1}$$

(4.1)

So, an increase in $I_{C1}$ will cause an increase in the voltage drop across $R_{C1}$ and bring $V_{C1}$ to a lower level. It may reach a level where it is not enough to bias next stages, i.e. $Q_2$ and $Q_3$.

From the simulation results shown in Fig 4.3 and Table 4.1, at $R_f=3K\Omega$ all transistors are operating in active region as it is obvious form their $V_{BES} (\approx 0.8V)$ and $V_{CES} (>0.3V)$. If $R_f$ is increased to $17K\Omega$, the collector current at $Q_1$ will increase from $0.327mA$ (at $R_f=3K\Omega$) to $0.436mA$. Consequently, $Q_3$ is turned off ($V_{BES}=1.59mV$) due to the drop in $V_{C1}$ caused by the increase in $I_{C1}$, as indicated by equation 4.1.
**For Low \( R_f \)**

A decrease in \( R_f \) will cause an increase in the current following through it. Since

\[
I_{RE1} = I_{RF} + I_{EI}
\]

and \( I_{RE1} \) is constant so \( I_{EI} \) will decrease. Consequently the voltage drop across \( R_{CI} \) will decrease causing an increase in the voltage level of \( V_{CI} \). It may reach to a value which is too high for biasing following stages causing \( Q_3 \) to saturate.

As shown in Fig 4.4 and Table 4.1, for \( R_f = 1K \Omega \), the collector current at \( Q_3 \) will decrease form 0.327mA (at \( R_f = 3K \Omega \)) to 0.191mA. Per equation 4.1, a decrease in \( I_{CI} \) will cause an increase in \( V_{CI} \). As a result, \( Q_3 \) enters the saturation region (\( V_{CE3} = 0.107V < 0.3V \)).
To sum up, $R_f$ dramatically affects the DC bias of the circuit. The circuit as is will not be suitable for variable gain or AGC circuits where $R_f$ is to be varied over wide range.

The main objective of next section is to design a CB transimpedance preamplifier that will be DC stable over a wide range of $R_c$. Eliminating the peaking introduced with varying $R_f$ will be discussed at the end of this chapter.
### Table 4.1: DC simulation results

#### (A) $R_f = 1 \Omega$

<table>
<thead>
<tr>
<th>NAME</th>
<th>$Q_0Q_2$</th>
<th>$Q_0Q_1$</th>
<th>$Q_0Q_4$</th>
<th>$Q_0Q_3$</th>
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4.3 Proposed Solutions for $R_f$ Variation:

The effect of $R_f$ on circuit biasing is due to its contribution to the value of $V_{CI}$. It may decrease the level of $V_{CI}$ to below the minimum level to turn $Q_2$ and $Q_3$ on. Moreover, it may increase $V_{CI}$ to a level where $Q_3$ will enter a saturation region. To solve this issue the level of $V_{CI}$ should be kept approximately constant regardless of the change in $R_f$. $V_{CI}$ should be designed to be sufficient to bias consequent stages, e.g. $Q_1$ and $Q_2$ in Fig 4.1.

The following equations show the relation between $V_{CI}$ and $R_f$ in a circuit similar to Fig 4.1:

\[ V_{CI} = V_{CC} - R_{C1}I_{C1} \]  \hspace{1cm} (4.2)

\[ I_{RC1} \approx I_{C1} \approx I_{E1} \]  \hspace{1cm} (4.3)

Provided $Q_1$ and $Q_2$ are in active region and $I_{B2} << I_{C1}$.

\[ I_{E1} = I_{RE1} - I_{Rf} \]  \hspace{1cm} (4.4)

\[ \Rightarrow I_{E1} = \frac{V_{E1} - V_{EE}}{R_{E1}} - \frac{V_{E4} - V_{E1}}{R_f} \]  \hspace{1cm} (4.5)

Substitute 4.3 and 4.5 in 4.2

\[ V_{CI} = V_{CC} - \frac{R_{C1}}{R_{E1}} (V_{E1} - V_{EE}) + \frac{R_{C1}}{R_f} (V_{E4} - V_{E1}) \]  \hspace{1cm} (4.6)
To preserve the value of $V_{CI}$ while $R_f$ is varied, one of the following scenarios may be implemented:

a. Changing $R_{CI}$ and $R_{E1}$ with a change in $R_f$.

b. Changing $R_{CI}$ with a change in $R_f$.

c. Reducing the voltage drop across $R_f$.

4.3.1 Varying $R_{CI}$ and $R_{E1}$:

If $R_f$ is increased (or decreased), $R_{CI}$ should be increased (or decreased) accordingly to correct for the change in the third term of equation 4.6. Since the change in $R_{CI}$ will also effect the second term of the equation so $R_{E1}$ should be changed to keep $V_{CI}$ constant. In this case, two variable resistors are needed to correct for the change in $V_{CI}$ due to $R_f$ variation.

An example for this solution is shown in Fig 4.5 where $R_{CI}$, $R_{E1}$ and $R_f$ are made equal. Simulation results are shown in Fig 4.6.

$$R_{CI} = R_{E1} = R_f \quad (4.7)$$

$$\Rightarrow V_{CI} = V_{CC} - (V_{E1} - V_{EE}) + V_{E4} - V_{E1} \quad (4.8)$$

$$\Rightarrow V_{CI} = V_{CC} - 2V_{E1} + V_{EE} + V_{E4} \quad (4.9)$$
Fig 4.5: Proposed circuit for the first solution.

Fig 4.6 (A): Simulation results of Fig 4.5: Transimpedance gain with $R_f$ varied from 1KΩ to 31KΩ.

Fig 4.6 (B): Simulation results of Fig 4.5: Gain, noise and bandwidth vs. $R_f$. 

\[ R_f = 1k, 3k, 5k, \ldots, 31k \]

Graphs showing frequency response with different $R_f$ values.
4.3.2 Varying $R_{E1}$:

As mentioned earlier, if $R_f$ is increased (or decreased), the third term in equation 4.6 and hence $V_{C1}$ will change. The change in $V_{C1}$ can be corrected by making $R_{E1}$ follow the change in $R_f$. So, a change in the second term will compensate for that happened in the third and hence $V_{C1}$ will remain constant. In this scenario, just one variable resistance is needed to correct for the effect of varying $R_f$ on the circuit biasing. It should be considered that there is a high limit for $R_f$ after which the overall gain of the preamplifier will not be equal to $R_f$ but less.

Open loop gain:

$$A \approx -g_{m3}R_{C1}R_{C3}$$  \hspace{1cm} (4.10)

$$\beta = \frac{1}{R_f}$$  \hspace{1cm} (4.11)

To make the overall gain is equal to $R_f$, the following condition should be met:

$$A\beta \gg 1 \Rightarrow g_{m3}R_{C1}R_{C3} \gg R_f$$  \hspace{1cm} (4.12)

Circuit and simulation results are shown in Fig 4.7 and 4.8 respectively.
Fig 4.7: Proposed circuit for the second solution.
(Here, $R_C1$ determines the upper limit of $R_f$ & $R_{2f}$ before the circuit biasing get effected)

Fig 4.8 (A): Simulation results of Fig 4.7: Transimpedance gain with $R_f$ varied from 1KΩ to 31KΩ.
4.3.3 Reducing $V_{RF}$

Back to equation 4.6, the effect of $R_f$ on $V_{CI}$ can be minimized if

$$R_{CI} \frac{V_{E4} - V_{E1}}{R_f} \approx 0$$  \hspace{1cm} (4.12)

$$\therefore R_{CI} \neq 0 \Rightarrow \frac{V_{E4} - V_{E1}}{R_f} \approx 0$$  \hspace{1cm} (4.13)

$$\therefore R_f \text{ to be varied} \Rightarrow V_{E4} \approx V_{E1}$$  \hspace{1cm} (4.14)

So, no correction action is required here to maintain the value of $V_{CI}$ constant over a variation of $R_f$. The only thing should be considered is to design the circuit with $V_{E4}$ approximately equals $V_{E1}$. 
Fig 4.9 shows a circuit implemented according to this solution. In this circuit a shift level transistor (Q42) is added and biased by a negative voltage to reduce the voltage drop across $R_f$. In other words, $V_{E42}$ is designed to be approximately equal to $V_{E1}$. In equation 4.14, $V_{E4}$ will be replaced by $V_{E42}$. Simulation results are shown in Fig 4.10.

Fig 4.9: Proposed circuit for the third solution:
(Q42 is used and biased by negative voltage to allow $V_{E42}$ to be equal to $V_{E1}$.)

Fig 4.10 (A): Simulation results of Fig 4.9: Transimpedance gain with $R_f$ varied from 1KΩ to 31KΩ.
Fig 4.10 (B): Simulation results of Fig 4.9: Gain, noise and bandwidth vs. $R_f$

Rf is logarithmically varied from 1ohm to 100kohm

Gain=89.434K at Rf=100Kohm

Gain=0.98 at Rf=1ohm

Fig 4.10 (C): Simulation results of Fig 4.9: Transimpedance gain with $R_f$ logarithmically varied from 1Ω to 100KΩ.

From the above results, $R_f$ can be varied from 1Ω to 100KΩ without affecting the DC operation of the circuit, see Table 4.2 DC biasing currents and voltages. Also, the overall transimpedance gain of the preamplifier is approximately equal to the value of $R_f$. 
Table 4.2: DC simulation results for the third DC solution.

(A) $R_f=1\Omega$

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(B) $R_f=1\Omega$

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(C) $R_f=100\Omega$

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To sum up, the DC biasing of the CB transimpedance preamplifier can be made insensitive to the change of \( R_f \) by minimizing the current flowing through \( R_f \). In other words,

\[
I_{R_f} = \frac{V_{R_f}}{R_f} \approx 0
\]  

(4.15)

So, \( V_{R_f} \) should be made small to satisfy the above condition regardless of the value of \( R_f \). All this is needed to implement a variable gain transimpedance preamplifier. Its gain is to be controlled by \( R_f \) without affecting the DC biasing.

4.4 Effect of \( R_f \) Variation on Small Signal:

As mentioned earlier, the main purpose of varying \( R_f \) is to implement a variable gain transimpedance preamplifier that can be used in an AGC or programmable (variable) gain circuits. For a variable gain control, one or more circuit elements should be varied to control the gain. The feedback resistor is the proper element that can be varied to control the gain in a transimpedance preamplifier. In this section, the effect of varying \( R_f \) on gain and other circuit performance aspects will be studied.

The following formulas can be derived for the basic circuit shown if Fig 4.1:

**Open loop gain (A):**
\[
A = \frac{\frac{\text{vo}}{\text{in}}}{\frac{\text{ve1}}{\text{in}} \cdot \text{ve2} \cdot \text{ve3}} \cdot \frac{\text{vo}}{\text{ve3}} \quad (4.16)
\]

\[
A \approx -g_m R'_{\text{C1}} R'_{\text{C3}} \frac{R'_{\text{E1}}}{r_{\text{e1}} + R'_{\text{E1}}} \quad (4.17)
\]

Where \( R'_{\text{C1}} \approx R_{\text{C1}} / R_{\text{in2}} \approx R_{\text{C1}} \) provided \( R_{\text{in2}} = (\beta_2 + 1) R_{\text{E2}} + r_{\pi2} >> R_{\text{C1}} \)

\[
R'_{\text{E1}} = R_{\text{E1}} / R_f
\]

\[
R'_{\text{C3}} \approx R_{\text{C3}} / R_{\text{in4}} \quad \text{and} \quad R_{\text{in4}} = r_{\pi4} + (\beta_4 + 1)(R_{\text{E4}} / R_f)
\]

**Feedback Factor (B):**

\[
\beta = -\frac{1}{R_f} \quad (4.18)
\]

**Loop Gain:**

\[
\beta A = g_m R_{\text{C1}} R_{\text{C3}} \frac{1}{R_f} \frac{R_{\text{E1}} / R_f}{r_{\text{e1}} + R_{\text{E1}} / R_f} \quad (4.19)
\]

**Input Resistance:**

\[
R_{\text{in}} = r_{\text{e1}} / R_{\text{E1}} / R_f \quad (4.20)
\]

**Output Resistance:**

\[
R_o = R_{\text{E4}} / R_f \div \frac{R_{\text{C3}} + r_{\pi4}}{\beta_4 + 1} \quad (4.21)
\]

**Adding the feedback effect:**

**Overall Gain (A_{of}):**

\[
A_{of} = \frac{A}{1 + \beta A} \quad (4.22)
\]

To make the overall transimpedance gain linearly proportional to \( R_f \) the following condition should be achieved:

\[
\beta A >> 1 \Rightarrow A_{of} \approx R_f \quad (4.23)
\]
For high $R_f$:

$$g_{m3}R_{C1}R_{C3} \gg R_{high}$$  \hspace{1cm} (4.24)

For Low $R_f$:

$$g_{m3}R_{C1}R_{C3} \gg r_{ei}$$  \hspace{1cm} (4.25)

Meeting equation 4.24 for high $R_f$ will satisfy 4.25.

Overall Input Resistance:

$$R_{in} = \frac{R_{in}}{1 + \beta A} \approx \frac{r_{ei}R_f}{g_{m3}R_{C1}R_{C3}} \text{ provided } \beta A >> 1$$  \hspace{1cm} (4.26)

For transimpedance preamplifier, it is desired to have a very low input resistance.

Applying equation 4.24 will guarantee the achievement of low input impedance for high $R_f$.

$$R_{in} \ll r_{ei}$$  \hspace{1cm} (4.27)

Overall Output Resistance:

$$R_{of} = \frac{R_o}{1 + \beta A} \approx \frac{R_{E4} // R_f // \frac{R_{C1} + r_{\pi 4}}{\beta_4 + 1}}{g_{m3}R_{C1}R_{C3} \frac{1}{R_f} \frac{R_{E1} // R_f}{r_{ei} + R_{E1} // R_f}}$$  \hspace{1cm} (4.28)

A low output resistance is another important specification for a transimpedance preamplifier.

For low $R_f$:
\[ R_{of} \approx \frac{r_{e1} R_{fow}}{g_m R_c R_C} \]  

(4.29)

From equation 4.25;

\[ R_{of} \ll R_{fow} \]  

(4.30)

For high \( R_f \)

\[ R_{of} \approx \frac{R_{f_{\text{high}}}}{g_m R_c R_C} \frac{R_C + r_e}{\beta + 1} \quad \text{provided} \quad \frac{R_C + r_e}{\beta + 1} \ll R_{E4} \]  

(4.31)

From equation 4.24:

\[ R_{of} \ll \frac{R_C + r_e}{\beta + 1} \]  

(4.32)

which is very low provided \( R_C < 50 \Omega \).

**Noise Performance:**

The noise is inversely proportional to the value of \( R_f \) so it will increase as \( R_f \) is decreased. It is not an issue because \( R_f \) will be decreased to reduce the gain when a high-level signal is received. So, the signal to noise ratio will not be affected.

\[ SNR = \frac{i_s(t)}{i_n(t)} \]  

(4.33)

where \( i_s(t) \) is the input current signal and \( i_n(t) \) is the input referred noise.

Moreover, for optimum noise performance resistance values in the first two stages should be high. Also, current levels in \( Q_1 \) and \( Q_2 \) should be minimized.
Frequency Response:

The open loop poles for the transimpedance preamplifier can be described in terms of their time constants as follows:

\[
\tau_{RC1} = \left[ \frac{RC1}{(r_{\pi2} + (\beta_2 + 1)R'_{E2})} \right] \cdot \left[ C_{\mu1} + \frac{C_{\pi2}}{1 + g_{m2} \cdot R'_{E2}} \right] 
\]
(4.34)

\[
\tau_{RE2} = \left[ \frac{RE2}{\beta_2 + 1} \right] \cdot \left[ C_{X3} + \frac{R_{C3}}{(1 + g_{m3}R_{C3}) \cdot C_{\mu3}} \right] 
\]
(4.35)

\[
\tau_{RC3} = \left[ \frac{RC3}{(r_{\pi4} + (\beta_4 + 1) \cdot R_{E4} // R_f)} \right] \cdot \left[ C_{\mu3} + \frac{C_{\pi4}}{1 + g_{m4} \cdot R_{E4} // R_f} \right] 
\]
(4.36)

Due to the low input and output resistance, poles at input and output nodes will be too high. So, the frequency response of the circuit will be insensitive to the internal capacitance of the photo detector, up to some limit.

While the dominant poles are insensitive to the variation of \( R_f \) but the effective bandwidth will change with the value of \( (1 + \beta_4) \) which is \( R_f \) dependent.

The following remarks are obtained from previous formulas and discussion:

1. Input and output resistances satisfy the requirements for a transimpedance preamplifier, i.e. low \( R_{in} \) and \( R_{out} \), provided \( r_{el} \) and \( r_{ed} \) are very small.

2. With high open loop gain, the overall gain approximately equals to \( R_f \).

3. While the open loop poles are insensitive to \( R_f \), the overall frequency response does. Peaking may be introduced with the change in \( R_f \) if a dominant pole is brought toward the second dominant pole.
4. Replacing the common emitter stage with a cascode will enhance the frequency performance of all circuits provided that the dominant pole is at the base of Q3 and $C_{\text{e3}}$ is not much higher than $(1 + g_{m3} R_C 3) C_{\mu3}$, which is the usual case.

5. To optimize the noise performance, the first two stages (Q1 and Q2) should be designed with high resistors (in KΩ range) and low level of currents (in tenths of mA).

4.5 Optimized CB Transimpedance Preamplifier:

In the circuit shown in Fig 4.11, the common emitter stage is replaced by a cascode to reduce the miller effect at Q3. To achieve good results with this circuit, the following requirements should be met:

1. $V_{E1} = V_{E4}$ to allow $R_f$ to be varied without affecting the DC bias.

2. $R_{C1}, R_{E1}$ and $R_{E2}$ should be in KΩ range to minimize the noise.

3. $I_{C1}$ and $I_{C2}$ should be designed to be tenths of mA.

4. $I_{C3}$ and $I_{C4}$ should be greater than 1mA.

5. High loop gain, i.e. $g_{m3} R_C 3 R_{C1} \gg \text{max } R_f$

6. $V_{C1}$ should be enough to bias transistors in following stages, e.g Q2 and Q3.

7. $R_{C1} > R_{C3}$ to make $w_{R_{C1}}$ dominant provided $C_{\mu1} > C_{\mu3}$:

$$\tau_{R_{C1}} = \left( \frac{R_{C1}}{(r_{x2} + (\beta_2 + 1) R'e2)} \right) \cdot \left[ C_{\mu1} + \frac{C_{x2}}{1 + g_{m2} \cdot R'e2} \right] \approx R_{C1} C_{\mu1}$$  \hspace{1cm} (4.37)

$$\tau_{R_{E2}} = \left( \frac{R_{E2}}{\beta_2 + 1} \right) \cdot \left[ C_{x3} + 2C_{\mu3} \right] \approx \frac{R_{C1} + r_{x2}}{\beta_2 + 1} \cdot \left[ C_{x3} + 2C_{\mu3} \right]$$  \hspace{1cm} (4.38)
\[ t_{RC3} = \left[ \frac{1}{\tau_{RC3}} \right] \approx R_{C3} C_{\mu 32} \left( 1 + \frac{C_{E4}}{1 + g_{m4} R_{E4} / R_f} \right) \]

The overall bandwidth assuming \( w_{RC1} \) is the dominant can be written as follows:

\[ f_{3dB} = (1 + \beta A) \cdot f_{RC1} \approx (1 + \frac{g_{m3} R_{C1} R_{C3}}{R_f}) \cdot \frac{1}{2\pi R_{C1} C_{\mu 1}} \approx \frac{g_{m3} R_{C3}}{2\pi R_f C_{\mu 1}} \]

(4.40)

It is obvious from the above equation that the overall bandwidth depends on \( R_f \).

Higher bandwidth is achievable at lower \( R_f \), see simulation results in Fig 4.12(B).

A level shift transistor is used in Fig 4.11 to make \( V_{CL} \) greater than \( V_{BL} \) and hence to enforce \( Q_1 \) to be active.

---

**Fig 4.11:** Proposed CB transimpedance preamplifier.

\((R_{E1}=8.8K, R_{C1}=7.35K, R_{E2}=1K, R_{C2}=4.38K, R_{E3}=4K, C_f=10\text{ femtoF})\)
Fig 4.12 (A): Simulation results for Fig 4.11: Transimpedance gain with $R_f$ varied from $1\,\text{k}\Omega$ to $31\,\text{k}\Omega$.

Fig 4.12 (B): Simulation results for Fig 4.11: Gain, noise and bandwidth vs. $R_f$. 
Fig 4.12 (C): Simulation results for Fig 4.11: Transimpedance gain with $R_f$ logarithmically varied from 1Ω to 100KΩ.

From the above simulation, this circuit has the following advantages:

1. Its overall transimpedance gain is equal to the value of $R_f$ over a wide range of $R_f$ variations (1Ω to 100KΩ).

2. If an AGC is to be added to this circuit, then $R_f$ can be safely used to control the gain without affecting the DC biasing. It may be implemented by a voltage-controlled-resistance. A MOSFET can do the job.

3. The input referred current noise is low.

4. Frequency performances are acceptable considering the non-sophisticated transistor model used in simulations.

5. While there are two power supply sources but the maximum power dissipation is 34mW. The level of currents flowing through the last two stages, i.e. $Q_3$ and $Q_4$, mainly controls the amount of power dissipated.

6. Output DC level is designed to be near the middle of the power source level to allow for a wide dynamic range.
7. Using higher value of power supplied- if it is practical- will enhance the dynamic range and noise performances after proper scaling of resistors.

The frequency advantage of this circuit will not be visible due to the limitation of transistor model used. Peaking is introduced with $R_f$ variations. There are few techniques to eliminate the peaking. One way is to reduce the open loop gain but the overall gain will not be proportional to $R_f$ at high values. The circuit usability in variable gain circuits will be restricted. An alternative way for removing the peaking without affecting the gain is to utilize $C_f$ as a compensation capacitor. Its proper value can be calculated by:

$$
C_f = \frac{1}{2\pi f_p R_f}
$$

where $f_p$ is the smallest peaking frequency. Refer to Fig 4.13 for simulation results.

To use a single power supply, the base voltages at $B_1$ and $B_{31}$ should be greater than zero. So, $Q_1$ and $Q_{31}$ are biased by $V_{CC}$ via diode-connected transistors as shown in Fig 4.14, see Fig 4.15 for simulation results. By this connection, new poles will be generated at $B_1$ and $B_{31}$, which will affect the frequency response of the circuit. Moreover, the input-referred noise will increase due to the contribution of transistors connected at $B_1$. Also, the use of 3V-power supply is possible with modifications in resistance values. It will be on the expense of dynamic range and noise performances.
Fig 4.13: Simulation results for Fig 4.11 with $C_f=40\text{ femtoF}$.  

Fig 4.14: CB transimpedance preamplifier with a single power supply.
Summary:

The main objective of this chapter was to design a variable gain transimpedance preamplifier with common base front-end configuration. Obstacles in the way of this design are the low open loop gain of the common base and the bias sensitivity to the $R_f$ variations. The first issue was resolved by adding a common emitter or a cascode. The DC bias sensitivity to $R_f$ value was studied and solutions were proposed. Moreover, the effect of $R_f$ on the small signal was analyzed. Based on results, an optimized circuit was introduced. The circuit advantage in the frequency performance was not clear due to the limitation of transistor model used in simulations. To eliminate peaking introduced by the aforementioned issue, a compensation capacitor was used. The proposed circuit was stable from the DC point of view over very wide $R_f$ range (1Ω to 100KΩ) with overall gain approximately equals $R_f$. So, it is an excellent candidate to be used as a programmable transimpedance preamplifier where a single element ($R_f$) is to be programmed to control the gain. Also, it can be integrated with an AGC circuit to produce a very wide dynamic range preamplifier. Its
noise performance is mainly controlled by the value of $R_f$. It was generally within the acceptable range and can be optimized with high $R_f$. In addition to the above, it consumes low power around 34mW so it is appropriate for applications where low power consumption is desired. A transimpedance circuit with a single power and maintaining approximately the same level of performance was also proposed.
CHAPTER 5:

CURRENT CONVEYOR AS A PROGRAMMABLE TRANSIMPEDANCE PREAMPLIFIER

5.1 Introduction:

Current mode circuits offer valuable advantages in speed, bandwidth and accuracy. A circuit can be defined as a current mode if its signal is represented by time varying current. One of these circuits gaining considerable attention is the Current Conveyor circuit.

A current conveyor is a device with three terminals labeled X, Y and Z. Actual development of current conveyors started on 1968 [45, 46]. During the past three decades different configurations were implemented. Those are first generation current
conveyor (CCI), second generation current conveyor (CCII) and third generation current conveyor (CCIII). The following matrix describes the relations between current conveyor terminals for the three different configurations:

\[
\begin{bmatrix}
i_y \\
v_x \\
i_z
\end{bmatrix} =
\begin{bmatrix}
0 & a & 0 \\
1 & 0 & 0 \\
0 & b & 0
\end{bmatrix}
\begin{bmatrix}
v_y \\
i_x \\
v_z
\end{bmatrix}
\]

(5.1)

\[a = 1\] for CCI
\[a = 0\] for CCII
\[a = -1\] for CCIII

The value of \(b\) can be +1 or -1 determining the polarity of the current conveyor. For example, in the CCII:

- If \(b = +1\), then \(I_x\) and \(I_z\) are flowing in the same direction, all into or out of the conveyor. In this case, CCII will be labeled as CCII⁺.

- If \(b = -1\), then \(I_x\) and \(I_z\) are flowing in different directions. One will flow out of the conveyor and the other will follow into the conveyor. This type of conveyor is labeled as CCII⁻.

Fig 5.1 shows the general block diagram representation of the current conveyor. Current directions will be set by the values of \(a\) and \(b\). [45], [47]
A controlled current conveyor (CCC) was proposed by Fabre [48] and mainly implemented for CCII, noted as CCCII. The driving concept behind controllability is the dependence of the internal resistance at port X (Rx) and hence Vx on the bias current (Io). The following formula shows this dependency:

\[ R_x = \frac{V_r}{2I_0} \]  

(5.2)

The symbolic representation of the CCCII is shown in Fig 5.2.

The equation 5.3 expresses the matrix relations for CCCII parameters.

\[
\begin{bmatrix}
    i_y \\
    v_x \\
    i_z
\end{bmatrix} =
\begin{bmatrix}
    0 & a & 0 \\
    1 & R_x & 0 \\
    0 & b & 0
\end{bmatrix}
\begin{bmatrix}
    v_y \\
    i_x \\
    v_z
\end{bmatrix}
\]  

(5.3)
Fig 5.2: Symbolic representation of CCCII.

In this chapter, the usage CCCII as a programmable transimpedance preamplifier will be investigated. Circuits with single stage and multi stages will be covered. Also, the Automatic Gain Control (AGC) concept will be applied using current conveyor blocks.

CCII circuit model [49] shown in Fig 5.3 will be used as the basis for formula extraction and manipulation. The circuit shown in fig 5.4 is a transistor level implementation of the CCCII that will be used in simulation [48]. Moreover, the complementary spice models Q2N3904 and Q2N3906 will be used for the NPN and PNP pairs. Simulation results are mainly to verify the concept not to obtain optimized results.
5.2 Programmable Transimpedance Preamplifier:

A single stage of a CCCII can be used as a transimpedance preamplifier as shown in the Fig 5.5.
In circuit (a), input signal is applied at port Y and output is taken from port Z.

\[ v_y = R_1 \cdot i_{in} \quad \text{provided} \quad R_y \gg R_1 \quad \text{and} \quad i_y \approx 0 \quad (5.3) \]

\[ i_x = \frac{v_y}{R_x} = \frac{R_1}{R_x} i_{in} \quad (5.4) \]

\[ i_z = i_x = \frac{R_1}{R_x} i_{in} \quad \text{provided} \quad R_z \gg R_2 \quad (5.5) \]

\[ v_o = \frac{R_1 R_2}{R_x} i_{in} \quad (5.6) \]

\[ R_{in} = R_1 // R_y \approx R_1 \quad (5.7) \]
\[ R_0 = R_2 R_2 \approx R_2 \] (5.8)

From equation 5.6, the circuit has a transimpedance gain of \( (R_t R_2 / R_X) \) where \( R_X \) is the internal resistance at port X. In the CCCII, \( R_X \) is controllable by the value of \( I_o \) based on the following equation:

\[ R_X = \frac{V_t}{2I_o} \]

Simulated relation between \( I_o \) and \( R_X \), as shown in Table 5.1, agrees with the above equation. Also, Fig 5.3.2 shows the effects of \( I_o \) on transimpedance gain.

<table>
<thead>
<tr>
<th>( I_o )</th>
<th>( R_X )</th>
<th>( R_{out}=R_2 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.2mA</td>
<td>62.0</td>
<td>99.9</td>
</tr>
<tr>
<td>0.4mA</td>
<td>31.2</td>
<td>99.7</td>
</tr>
<tr>
<td>0.6mA</td>
<td>20.8</td>
<td>99.5</td>
</tr>
<tr>
<td>0.8mA</td>
<td>15.7</td>
<td>99.3</td>
</tr>
<tr>
<td>1.0mA</td>
<td>12.5</td>
<td>99.2</td>
</tr>
<tr>
<td>1.2mA</td>
<td>10.4</td>
<td>98.9</td>
</tr>
<tr>
<td>1.4mA</td>
<td>8.9</td>
<td>98.7</td>
</tr>
<tr>
<td>1.6mA</td>
<td>7.8</td>
<td>98.4</td>
</tr>
<tr>
<td>1.8mA</td>
<td>6.9</td>
<td>98.2</td>
</tr>
<tr>
<td>2.0mA</td>
<td>6.2</td>
<td>98.0</td>
</tr>
</tbody>
</table>
As shown above, the increase in \( I_o \) will increase the gain because \( R_x \) will be decreased.

In the other circuit, Fig 5.5 (B) input is applied at port X and output voltage from port Z.

\[
\frac{v_o}{i_{in}} = R_2
\]

(5.9)

\[
R_{in} = R_x
\]

(5.10)

\[
R_o = R_2//R_Z \approx R_2
\]

(5.11)
Since the transimpedance gain is not a function of \( R_X \), so it can not be programmed by \( I_O \). Simulation results expressing the relationship between the transimpedance gain and \( I_O \) are shown in Figs 5.7.

![Graph showing gain vs. \( I_O \) for the circuit in Fig 5.5(B).]

Fig 5.7: Gain vs. \( I_O \) for the circuit in Fig 5.5(B).

Table 5.2 summarizes simulation results highlighting input and output resistance and the relation between \( R_X \) and \( I_O \).

As a transimpedance preamplifier specification, the input and output resistance should be small. To apply this on circuit (a), \( R_I \) and \( R_2 \) should be made small but they will dramatically reduce the gain. Buffer stages at input and output of the circuit will solve the issue. Implementations of those buffers using CCCII are addressed later in this chapter.
Table 5.2: Input resistance ($R_2$) and output resistance with $R_2=100\Omega$ and $I_0$ varies from 0.2mA to 2.0mA for the circuit in Fig 5.5(B).

<table>
<thead>
<tr>
<th>$I_0$</th>
<th>$R_{in}=R_X$</th>
<th>$R_{out}=R_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.2mA</td>
<td>61.4</td>
<td>99.9</td>
</tr>
<tr>
<td>0.4mA</td>
<td>30.8</td>
<td>99.7</td>
</tr>
<tr>
<td>0.6mA</td>
<td>20.6</td>
<td>99.5</td>
</tr>
<tr>
<td>0.8mA</td>
<td>15.5</td>
<td>99.3</td>
</tr>
<tr>
<td>1.0mA</td>
<td>12.4</td>
<td>99.2</td>
</tr>
<tr>
<td>1.2mA</td>
<td>10.4</td>
<td>98.9</td>
</tr>
<tr>
<td>1.4mA</td>
<td>8.9</td>
<td>98.7</td>
</tr>
<tr>
<td>1.6mA</td>
<td>7.9</td>
<td>98.4</td>
</tr>
<tr>
<td>1.8mA</td>
<td>7.0</td>
<td>98.2</td>
</tr>
<tr>
<td>2.0mA</td>
<td>6.3</td>
<td>98.0</td>
</tr>
</tbody>
</table>

For circuit (b), the input resistance $R_x$ is small which meets the specification. Output resistance, on the other hand, is $R_2$ which is controlling the gain. So, there is a contradiction between the requirements of high transimpedance gain and a low output resistance. A buffer is needed to solve this trade off.

As an advantage of circuit (a) over (b), its gain is greater by a factor of $(R_I / R_X)$, provided $R_I > R_X$. Moreover, it can be programmed by the value of $I_0$. 
5.3 Multistage CC-Based Programmable Transimpedance Preamplifier:

Multi stage CCCII, shown in Fig 5.8, will provide the low input and output resistances desired for a transimpedance preamplifier. At the same time, it will maintain the programmability of the transimpedance gain.

\[ R_{in} = R_{X1} \]  \hspace{1cm} (5.12)

\[ R_{o} = R_{X3} \]  \hspace{1cm} (5.13)

\[ Gain = \frac{V_o}{i_{in}} = \frac{R_{Z1}}{R_{Z1} + (R_{1} \parallel R_{Y2})} \cdot \left( \frac{1}{R_{X2}} \right) \cdot \frac{R_{Z2}}{R_{Z2} + (R_{2} \parallel R_{Y3})} \cdot \left( \frac{1}{R_{X3} + R_L} \right) \cdot \frac{R_L}{R_{X2}} \]  \hspace{1cm} (5.14)

\[ Gain \approx \frac{R_1}{R_{X2}} \cdot R_2 \]
It is clear from the above equations and associated simulation results, shown in Fig 5.9 and Fig 5.10, that this transimpedance preamplifier has the following features:

- Programmable Transimpedance gain,
- Low input resistance \( (R_{xi}) \) which can be controlled by \( Io_1 \),
- Low output resistance \( (R_{xo}) \), controllable by \( Io_3 \),
- Transimpedance gain is insensitive to a load resistance.

Fig 5.9 (A): Simulated transimpedance gain for the transimpedance preamplifier in Fig 5.8.
Fig 5.9 (B): Simulated gain vs. $I_{o2}$ for the transimpedance preamplifier in Fig 5.8.

Fig 5.10: Simulated gain vs. $R_L$ for the transimpedance preamplifier in Fig 5.8.

Implementing transimpedance preamplifier with CCCII blocks only and without external resistors is possible as shown in Fig 5.11.
Fig 5.11: Programmable transimpedance preamplifiers with no external resistors.

\[\begin{align*}
R_{in} &= R_{X1}/R_{X2} \approx R_{X1} \quad \text{(can be controlled by } I_{O1}) \\
R_o &= R_{X4} \quad \text{(can be controlled by } I_{O4}) \\
\text{Transimpedance Gain} &= \frac{R_{X1}}{R_{X2}}\frac{R_{X3}}{}
\end{align*}\]

Provided \(R_L >> R_{X4}\).

So, \(I_{O1}, I_{O2}\) and \(I_{O3}\) can program transimpedance gain, see Fig 5.12 to 5.14.

Fig 5.12: Simulated gain vs. \(I_{O1}\) for the transimpedance preamplifier in Fig 5.11.
Fig 5.13: Simulated gain vs. $I_{O2}$ for the transimpedance preamplifier in Fig 5.11.

Fig 5.14: Simulated gain vs. $I_{O3}$ for the transimpedance preamplifier in Fig 5.11.
5.4 Automatic Gain Control (AGC) for Current Conveyor Based Transimpedance Preamplifier:

The purpose of an automatic gain control (AGC) is to reduce the amplification if the level of input signal increases to a limit where a distorted output may be generated. The programmability feature of CCCII-based transimpedance preamplifier can be used to automatically control its gain. Fig 5.15 shows the main blocks in AGC circuit with CCCII based transimpedance preamplifier.

![Diagram of CCCII based transimpedance preamplifier with AGC.]

Fig 5.15: CCCII based transimpedance preamplifier with AGC.

The circuit in Fig 5.8 shows an example of CCCII based transimpedance preamplifier. As discussed earlier, its gain equals \((R_1 \times R_2 / R_{x2})\) and \(R_{x2}\) is controllable by \(I_{O2}\) in inversely basis, \((I_{O2}\) will be noted here as \(I_O)\). If \(V_O\) increases and becomes near distortion level, the gain should be decreased. Decreasing \(I_O\) should increase \(R_x\) and hence decrease the gain. The relation between \(V_O\) and \(I_O\) can be realized by a voltage controlled current source. Fig 5.16 shows a CCCII based voltage controlled current source.
Fig 5.16: CCCII based Voltage Controlled Current Source (VCCS).

\[
i_Z = \frac{V_{in}}{R_X}
\]  \hspace{1cm} (5.18)

if \( V_{in} = v_c \) \( \Rightarrow \) \( i_z = \frac{v_c}{R_{cc}} \)  \hspace{1cm} (5.19)

Increasing \( v_c \) will cause an increase in \( i_Z \). So, \( i_Z \) can not be taken as \( I_O \). Instead, the modified circuit shown in Fig 5.17 will be used.
Fig 5.17: Modified CCCII based VCCS.

\[ Io = I - \frac{v_c}{R_{xc}} \quad \text{provided} \quad \frac{v_c}{R_{xc}} \leq I \quad \text{and} \quad v_c \geq 0 \]  \hspace{1cm} (5.20)

Here, an increase in \( v_c \) will reduce \( Io \) and hence the transimpedance gain. The amount of change in \( Io \) with the change in \( v_c \) can be controlled by \( Io_C \), which controls \( R_s \).

An alternative implementation for the modified CCCII based VCCS is shown in Fig 5.18.
Fig 5.18: Alternative circuit for modified CCCII based VCCS.

\[ I_0 = \frac{V - v_c}{R_{sc}} \quad \text{provided} \quad v_c \leq V \quad \text{and} \quad v_c \geq 0 \]  \hspace{1cm} (5.21)

An increase in \( v_c \) will result in a decrease of \( I_0 \) from the initial value \( (v_c/(R+R_{sc})) \) by a factor of \( (1/(R+R_{sc})) \). The slope of change in \( I_0 \) and its initial value are controllable by \( I_{OC} \).

Using one of these circuits within an AGC circuit will make it programmable.

Fig 5.19 shows a CCCII based transimpedance preamplifier combined with an AGC.
A current based AGC can be implemented by passing $i_{23}$ through a proper peak detector and then through a current controlled current source (CCCS). Output current of CCCS will control $I_o$ and hence the preamplifier gain.

There are alternative schemes for implementing AGC. Since the transimpedance gain is $(R_1 * R_2 / R_x)$ so $R_1$ or $R_2$ can be used to control the gain. In this case, the control voltage coming out of peak detector circuit should be fed to a voltage-controlled resistance. $R_1$ or $R_2$ may be replaced by that voltage controlled resistance. The drain to source resistance ($r_{ds}$) of a MOSFET is a simple example of a voltage-controlled resistance.
CHAPTER 6

CONCLUSION and FUTURE WORK

6.1 Conclusion:

The followings are general conclusion remarks obtained as a result of the work done in this thesis:

- Different front-end transimpedance preamplifier configurations, used in optical communication applications, were evaluated. Those configurations are common base, common emitter, cascode and cascade. Bandwidth, transimpedance gain, noise and dynamic range are the major performance criteria used as evaluation basis. Moreover, those configurations were evaluated on their performance over a range of $R_f$ values. The common base was the worst from this perspective. In general, the cascade exhibits better performance among others.

- The VGA or AGC concept for transimpedance preamplifier was introduced. A peak detector and variable resistance were designed and simulated. They were
integrated with a transimpedance preamplifier to show the concept. The need for a transimpedance preamplifier accepting a wide variation of $R_f$ was highlighted.

- The DC bias sensitivity of a common base transimpedance preamplifier to $R_f$ value was studied and solutions were proposed. Moreover, the effect of $R_f$ on the small signal was analyzed. Based on results, an optimized circuit was introduced. The circuit advantage in the frequency performance was not clear due to the limitation of transistor model used in simulations. The proposed circuit exhibited a DC stability over a very wide range of $R_f$ (1Ω to 100KΩ) with overall gain effectively controlled by $R_f$. So, it is an excellent candidate to be used as a variable (programmable) transimpedance preamplifier where a single element ($R_f$) is to be used to control (program) the gain. Also, it can be integrated with an AGC circuit to produce a very wide dynamic range preamplifier. Its noise performance is mainly controlled by the value of $R_f$. It was generally within the acceptable range and can be optimized with high $R_f$. In addition to the above, it consumes low power around 34mW so it is appropriate for applications where low power consumption is desired. A transimpedance circuit with a single power and maintaining approximately the same level of performance was also proposed.

- The use of Current Conveyor (CC) as a transimpedance preamplifier was explored. Various configurations were proposed. Attractive features such as high transimpedance gain, low input and output resistances were achieved. Moreover, they are programmable so they can be tailored to meet application requirements. In addition, the CC-based AGC implementation for CC-based transimpedance preamplifier was proposed. Fair judgement of frequency and noise performances
of the proposed circuits can not be reached due to the use of non-sophisticated transistor pair.

6.2 Future Work:

As an enhancement and future work, the following areas may be addressed:

- Investigate performance aspects (especially frequency) of the optimized common based transimpedance preamplifiers by using higher frequency transistor models.
- Investigate the CC-based transimpedance preamplifier’s performance (especially frequency and noise) using high frequency CC circuits or transistor pairs.
APPENDIX

(PSPICE Transistors’ Parameters)
PSPICE Transistor Models:

Transistor models' parameters used in PSICE Simulations are listed in the following tables [55].

Table A: Available High Frequency NPN Transistor Model.

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REFERENCES:


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