Applications of Electronically Tunable OTA

A dissertation submitted in partial fulfilment of

the requirement for the degree of

MASTER OF TECHNOLOGY

IN

VLSI Design and Embedded Systems

by

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2012-14

CERTIFICATE

This is to certify that the thesis entitled "**Applications of Electronically Tunable OTA**" has been completed by **Kamlesh Kumar Tanwer** in partial fulfilment of the requirement of **Master in Technology** in **VLSI Design and Embedded Systems**. This is a record of his work carried out by him under my supervision and support. He has completed his work with utmost sincerity and diligence.

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ACKNOWLEDGEMENT

I wish to express my sincere gratitude to my supervisor **Dr. Malti Bansal**, Assistant Professor, Department of Electronics and Communication Engineering, Delhi Technological University, for her invaluable guidance and advice during every stage of this endeavor. I am greatly indebted to her for her continuing encouragement and support without which, it would not have been possible for me to complete this undertaking successfully. Her insightful comments and suggestions have continually helped me to improve my understanding.

I am also grateful to **Prof. Rajeev Kapoor**, HOD, Electronics and Communication Engineering Department, DTU for his immense support.

I would also like to acknowledge Delhi Technological University for providing the right academic resources and environment for this work to be carried out.

Last but not the least I would like to express sincere gratitude to my parents and my colleagues for constantly encouraging me during the completion of work.

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ABSTRACT

Over the past decade there has been increasing recognition of the importance of access to communication infrastructures and services for economic and social development. Generally, in communication systems, the processing of signal is accomplished in a single or a set of mixed-signal IC most of which are based on CMOS technology. OTA is a popular choice to implement filters, multipliers, oscillators, and several other important building blocks while satisfying the requirement of high frequency stability, wide input voltage range etc.

In the field of electric and electronic engineering, oscillators play an important role and have been widely applied in various aspects such as communications systems, instrumentation, measurement and signal processing, etc. The concept of oscillator design has been mainly on the requirement of multiple sinusoids which are 90◦ phase shifted, called Quadrature signal, for easy implementation with other circuits for example in the design of SSB modulator, etc. From the past, there have been attempts to synthesis the sine wave oscillator in both forms of current and voltage mode.

The Voltage-mode Quadrature oscillator using 3 operational transconductance amplifiers (OTAs) and 3 grounded capacitors and 3 grounded Resistor is presented. The proposed oscillator provides 2 sinusoidal output voltages with 90º phase difference. The condition of oscillation and the frequency of oscillation can be electronically controlled by adjusting the bias current of the OTA

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LIST OF ABBREVIATIONS &SYMBOLS

ASP	Analog Signal Processing
BPF	Band Pass Filter
BSF	Band Stop Filter
BW	Bandwidth
CMOS	Complementary metal-oxide semiconductor
CMRR	Common-mode rejection ratio
CO	Condition of Oscillation
dB	Decibel
Dec	Decade
FC	Folded Cascade
FO	Frequency of Oscillation
GaAs	Gallium arsenide
HPF	High Pass Filter
Hz	Hertz
IC	Integrated Circuit
LPF	Low Pass Filter
MHz	Mega Hertz
MIMO	Multiple input multiple outputs
NMOS	N-Channel metal-Oxide Semiconductor
OPAMP	Operational amplifier
OTA	Operational transconductance amplifier
PMOS	P-Channel metal-Oxide Semiconductor
QO	Quadrature Oscillator
RFIC’s	RF integrated circuits
SIMO	Single input multiple outputs
SoC	System on chip

SoI	Silicon on Insulator
SPICE	Simulation Program with Integrated Circuit Emphasis
THD	Total Harmonic Distortion
VCCS	Voltage controlled current source
VCVS	Voltage controlled Voltage source
A or G	Gain
C	Capacitance
C_{ox}	Gate oxide capacitance per unit area in F/m^2
f	Frequency in Hertz
g_m	Transconductance
I_D	Drain Current
L	Inductance or gate length
Q	Q factor
R	Resistance
T	Temperature in Kelvin
μ	Permeability
V_{CM}	Common Mode Input Voltage
V_{DD}	Positive supply voltage
V_{GS}	Gate-Source Voltage
V_{in}	Input Voltage
V_o	Output Voltage
V_{SS}	Negative supply voltage
V_{Th}	Threshold voltage
ω_o	Output pole frequency
W	Gate Width
Y	Admittance
Z	Impedance

CHAPTER 1

INTRODUCTION

1.1 Introduction

During the last two decades, research in the field of analog CMOS circuits has gained a lot of attention. Continuous efforts to improve CMOS technology enabled the integration of (largely digital) complete electronic systems on a single chip. Furthermore, analog signal processing can be favorable in terms of speed, chip area and power dissipation, especially for low and moderate precision circuit.

Although in most modern electronic systems, signal processing is performed in digital domain, but Analog Signal Processing (ASP) is still required since the physical world signal is represented in the analog form. As a result, most electronic equipments comprise analog processing circuitry that acts as interface between the digital world and the physical world. However, analog signal is processed for variety of purposes, such as to remove unwanted noise, to make the signal suitable for transmission or to extract certain meaningful information. The term ASP expresses plentiful of techniques that can be implemented to process analog signals including the theory and application of filtering, coding, transmitting, estimating, detecting, analyzing and reproducing analog signals.

Linear circuits, like amplifiers and filters, are indispensable analog building blocks. Their properties often critically determine system performance. In order to achieve high performance, circuits are usually designed in such a way, that the transfer function is mainly determined by a few carefully chosen components. *Passive* components, especially resistors and capacitors, are predominantly used for this purpose.

Although the use of passive components is preferable in many cases, there are also drawbacks. A major one is that electronic control of the value of these components is hardly possible. Such control is often desired in order to compensate for deviations from nominal component values due to fabrication tolerances, temperature variations etc. These deviations will change the transfer function of linear circuits, e.g. resulting in a shift of the pass-band of a filter.

Because electronic control is often needed, many MOS circuits with continuous electronic variability of the transfer function have been proposed. The present thesis deals with circuits that rely on the transconductance of a MOS transistor. Instead of relying on passive components, active devices now have a direct intended effect on the transfer. The transconductance of a MOS depends on gate geometry and biasing. Hence, designers both can dimension the nominal value of the transfer function, and adjust it electronically, as done in the well-known Transconductance-C filters

Operational Amplifiers (OPAMP) were the most popular choice as the building blocks for a wide range of such electronic circuits. Their popularity in circuit design was based on the fact that characteristics of the op-amp circuits with negative feedback (such as their gain) are set by external components with little dependence on temperature changes and manufacturing variations in the op-amp itself.

OPAMPs work well for low-frequency applications, such as audio and video systems. For higher frequencies, however, OPAMP designs become difficult due to their frequency limit [1], [2]. At those high frequencies, Operational Transconductance Amplifiers (OTAs) are deemed to be promising to replace OPAMPS as the building blocks. Theories of using OTAs as the building blocks for analog applications have been well developed [1], [2].To date, with much effort dedicated by analog IC researchers and the continuous scaling-down on commercial semiconductor technologies, the reported OTAs can work up to several hundred MHz’s.

OTA was the next choice to replace op-amp in most of their applications because of its ability to overcome the above mentioned shortcomings of the op-amp. Beside all these, OTA has a very useful feature of tunability.

Its transconductance can be changed by using a control voltage and hence can be used as controlled parameter in the circuits that needs to be tuned. Because of all these merits, nowadays OTA based circuits are extensively used to form several active basic building blocks for signal processing systems such as filters, multipliers, oscillators, controlled impedances, etc.

There are two modes depending upon the output: single ended and fully differential. Advantages of the differential structure include: improved output voltage swing, less susceptibility to common mode noise, and cancellation of even-order nonlinearities [3]. But the fully differential configuration lacks the current mirror capability of one of the current mirrors as achieved in the single ended configuration and hence is capable of producing only half of the bias current at output which can be solved by using a dual stage structure. Single-stage OTAs tend to be more power efficient than two-stage OTAs, because no power is wasted in driving the compensation capacitance in single-stage OTAs [4].

Evolution of Active Blocks

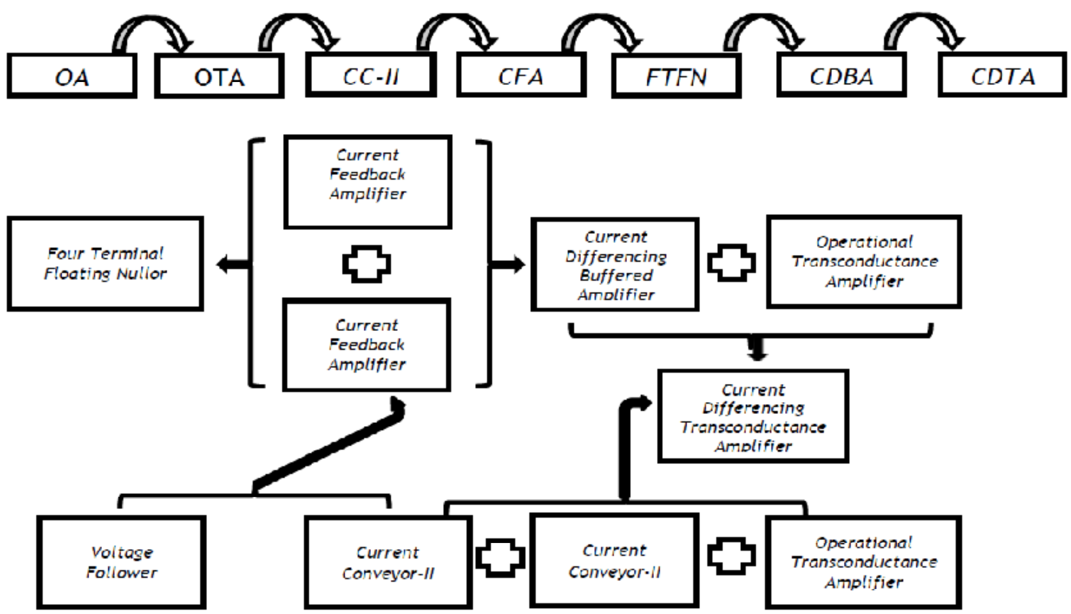


Figure 1.1: Evolution of Active Blocks

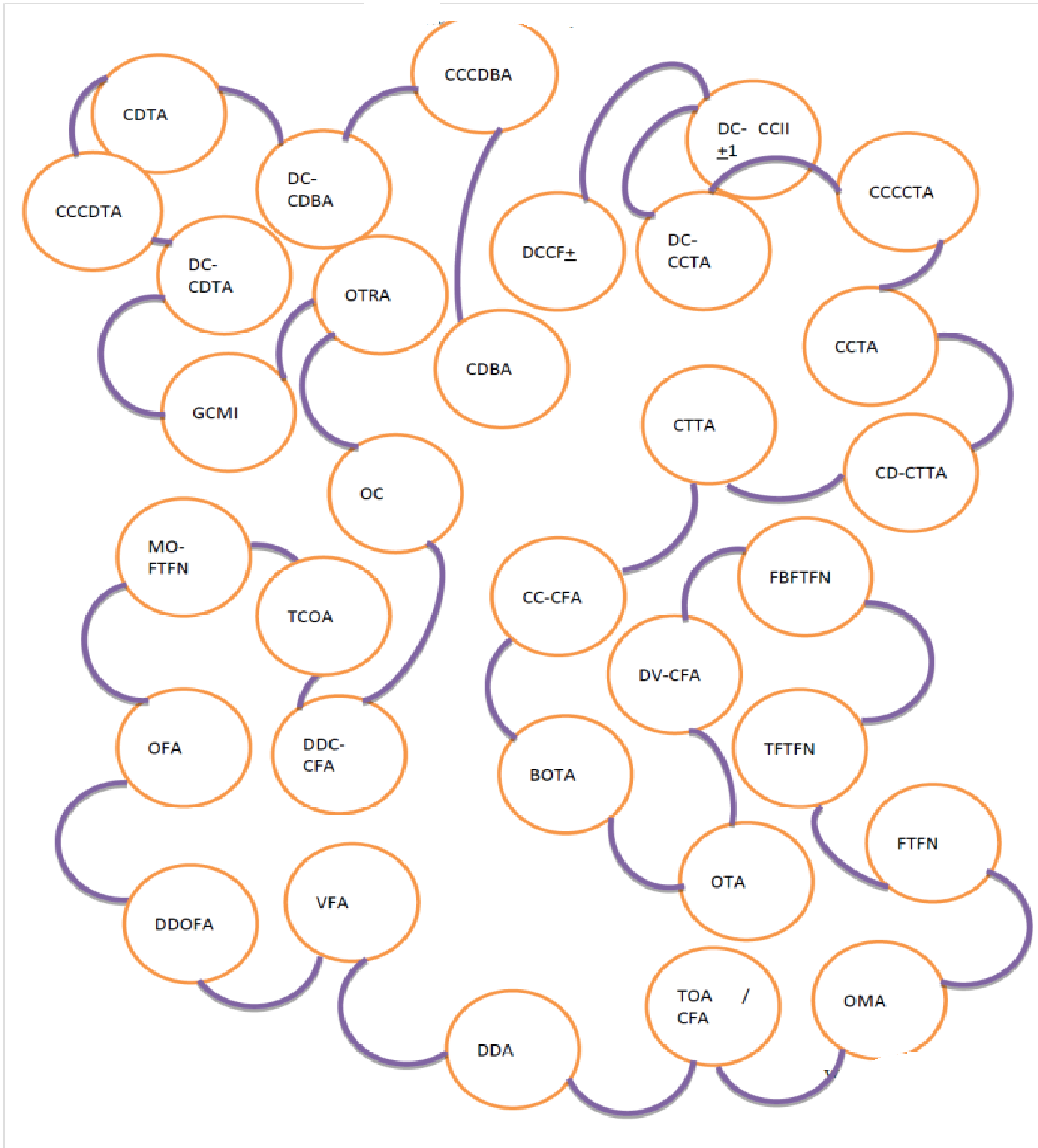


Figure 1.2: Evolution of active elements by suitable conditions of different active blocks

The evolution of active blocks is happening by modification of the basic elements such as VFA (Voltage Feedback Amplifier), CFA (Current Feedback Amplifier), OTA (Operational Trans conductance Amplifier), and particularly current conveyors (CC). The motivation behind such modifications is to increase the application potential of the element

- **OPAMP**

An operational amplifier, which is often called an op-amp, is a DC-coupled high-gain electronic voltage amplifier with a differential input and, usually, a single-ended output. Ideally the op-amp amplifies only the difference in voltage applied between its two inputs (V_+ andV_-), which is called the differential input voltage. The output voltage of the op-amp is given by the equation,

$$V_{out} = (V_+ - V_-).\, G_{open-loop} \quad (1.1)$$

Where V_+ is the voltage at the non-inverting terminal and V_- is the voltage at the inverting terminal and G open-loop is the open-loop gain of the amplifier.

The ideal operation is difficult to achieve and the non-ideal conditions often raise limitations like finite impedances and drift, their primary limitation being not especially fast. The typical performance degrades rapidly for frequencies greater than 1MHz, although some models are designed especially for higher frequencies. High input impedance at the input terminals (ideally infinite) and low output impedance at the output terminal(s) (ideally zero) are important typical characteristics. The other important fact about op-amps is that their open-loop gain is huge. This is the gain that would be measured from a configuration in which there is no feedback loop from output back to input. A typical open-loop voltage gain is ~ 10^4 -10^5 .

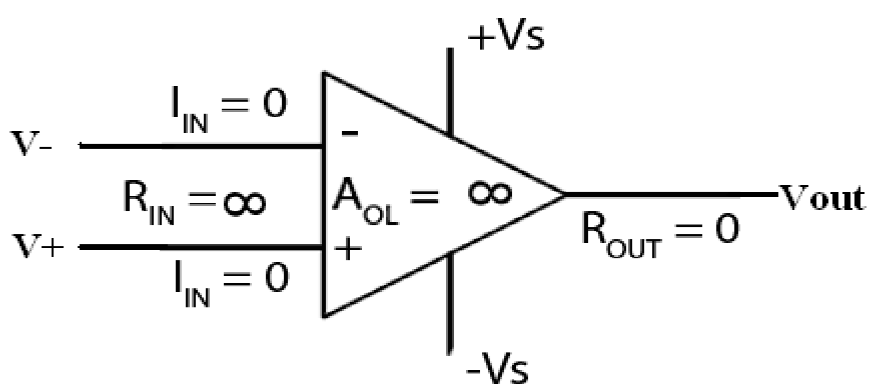


Figure 1.3: Ideal internal circuit of Op-amp

An ideal op-amp is usually considered to have the following properties, and they are considered to hold for all input voltages:

- Infinite open-loop gain.
- Infinite voltage range available at the output (V_{out}) (in practice the voltages available from the output are limited by the supply voltages Vs_{+} and Vs_{-})
- Infinite bandwidth
- Infinite input impedance
- Zero input current
- Zero input offset voltage
- Infinite slew rate
- Zero output impedance
- Infinite Common-mode rejection ratio (CMRR)
- Infinite Power supply rejection ratio for both power supply rails.

- CURRENT CONVEYOR

The current conveyor (CC) is the basic building block of a number of applications both in the current and voltage and the mixed modes. The principle of the current conveyor of the first generation was published in 1968 by K. C. Smith and A. S. Sedra. Two years later, today's widely used second-generation CCII was described in, and in 1995 the third-generation CCIII [2]. However, initially, during that time, the current conveyor did not find many applications because its advantages compared to the classical operational amplifier were not widely appreciated.

An IC Current Conveyor, namely PA630, was introduced by Wadsworth in 1989 and about the same time, the now well known CFOA AD844 was recognized to be internally a CCII+ followed by a voltage follower. Today, the current conveyor is considered a universal analog building block with wide spread applications in the current mode, voltage mode, and mixed mode signal processing. Several generations of current conveyors have been defined over the years. Undoubtedly, the second generation conveyor (CCII) is the more well known of the device. The terminal relations of a CCII can be characterized by

$$\begin{bmatrix} V_x \\ I_y \\ I_z \end{bmatrix} = \begin{bmatrix} \beta & 0 & 0 \\ 0 & 0 & 0 \\ 0 & \pm\alpha & 0 \end{bmatrix} \begin{bmatrix} V_y \\ I_x \\ V_z \end{bmatrix}$$

where $\alpha = 1 - \epsilon_i$ and $\beta = 1 - \epsilon_v$,$|\epsilon_i| << 1$ and $|\epsilon_v| << 1$, represent the current and voltage tracking errors, respectively, where the subscripts x, y, and z refer to the terminals labeled X, Y and Z in fig1 The CCII is defined in both a positive and a negative version where the +sign in the matrix is used for the CCII+ type conveyor and the –sign is used for the CCII- type conveyor. Its features find most applications in the current mode, when its voltage input y is grounded and the current, flowing into the low-impedance input x, is copied by a simple current mirror into the z output.

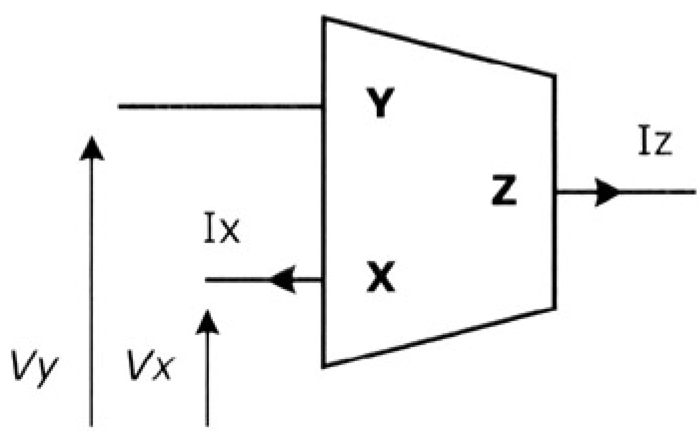


Figure1.4: Block diagram of CCII

- OPERATIONAL TRANSRESISTANCE AMPLIFIER

As signal processing extends to higher frequencies, traditional design methods based on voltage op-amps are no longer adequate. It is well known that a traditional operational amplifier has bandwidth which is dependent on the closed-loop voltage gain. The attempt to overcome this problem has led to a renewed interest in circuits which operate in current mode.

The OTRA is a current mode device that uses current mirrors and common source amplifier to give a current difference signal as input which in turn produce an appropriate voltage signal as output. The OTRA is a three terminal analog building block. Both the input and output terminals are characterized by low impedance. The circuit symbol of the OTRA is illustrated in Fig.1.5.The port relations of an OTRA can be characterized by the following matrix form [3].

$$\begin{bmatrix} v_p \\ v_n \\ v_z \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ R_m & -R_m & 0 \end{bmatrix} \begin{bmatrix} i_p \\ i_n \\ i_z \end{bmatrix}$$

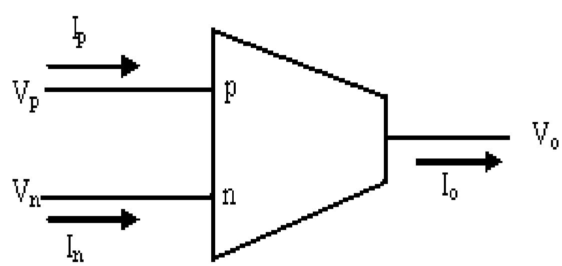


Figure 1.5: Circuit symbol of OTRA

It eliminates response limitations incurred by capacitive time constants leading to circuits that are insensitive to the stray capacitances at the input terminals. For ideal OTRA, the Transresistance gain, R_m, approaches infinity and external negative feedback must be used which forces the input currents to be equal [3]. Thus the OTRA must be used in a negative feedback configuration. Practically the Transresistance gain is finite and its effect should be considered.

The important advantages offered by OTRA are:

- Since the OTRA has one output terminal with low impedance and two input terminals that are virtually grounded, most effects of parasitic capacitances disappear and the remainder can be compensated without adding any extra components.
- Using current feedback techniques, OTRAs have a bandwidth almost independent of the closed loop-voltage gain.
- Due to the input terminals being virtually grounded, they are cascadable

• OPERATIONAL TRANSCONDUCTANCE AMPLIFIER

The operational transconductance amplifier (OTA) is an amplifier whose differential input voltage produces an output current [2-6]. Thus, it is a voltage controlled current source (VCCS). There is usually an additional input for a current to control the amplifier's transconductance. The OTA is similar to a standard operational amplifier in that it has a high impedance differential input stage and that it may be used with negative feedback.

The OTA is not as useful by itself in the vast majority of standard op-amp functions as the ordinary op-amp because its output is a current. One of its principal uses is in implementing electronically controlled applications such as variable frequency oscillators and filters and variable gain amplifier stages which are more difficult to implement with standard op-amps. In the ideal OTA, the output current is a linear function of the differential input voltage, and is given by:

$$I_{out} = (V_{in+} - V_{in-}).\, g_m \tag{1.2}$$

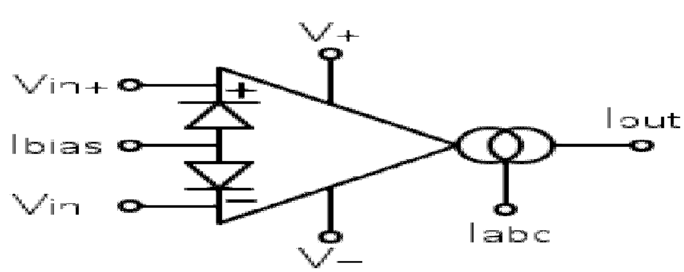


Figure 1.6: OTA Model

The amplifier's output voltage is the product of its output current and its load resistance:

$$V_{out} = I_{out} \,.\, R_d \tag{1.3}$$

The voltage gain is then the output voltage divided by the differential input voltage:

$$G_{voltage} = R_d \cdot g_m \tag{1.4}$$

The transconductance of the amplifier is usually controlled by an input current, denoted, I_{abc}.The amplifier's transconductance is directly proportional to this current. This is the feature that makes it useful for electronic control of amplifier gain, etc.

- CURRENT DIFERENCING BUFFERED AMPLIFIER

CDBA, current differencing buffered amplifier, is a multi-terminal active component with two inputs and two outputs [1-2]. Its block diagram can be seen from figure 1.7. It is derived from current feedback amplifier (CFA).

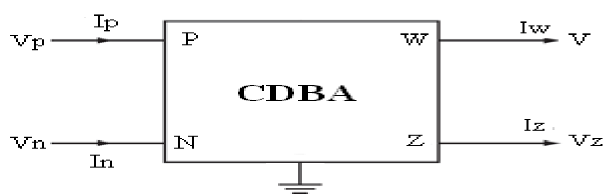


Figure 1.7 Block Diagram for CDBA

The characteristic equation of this element can be given as:

1. $$V_P = V_N = 0 \tag{1.5}$$

2. $$I_Z = I_P - I_N \tag{1.6}$$

3. $$V_W = V_Z \tag{1.7}$$

Here, current through z-terminal follows the difference of the currents through p-terminal and n-terminal. Input terminals p and n are internally grounded. The difference of the input currents are converted into the output voltage V_W, therefore CDBA element can be considered as a special type of current feedback amplifier with differential current input and grounded y input.

The CDBA is simplifying the implementation, free from parasitic capacitances, able to operate in the frequency range of more than hundreds of MHz (even GHz), and suitable for current mode operation while, it also provides a voltage output. Several voltage and current mode continuous-time filters, oscillators, analog multipliers, inductance simulators and a PID controller have been developed using this active element.

The CDBA offers several advantageous features viz., high slew rate, improved bandwidth, and accurate port tracking characteristics when configured with a pair of matched current feedback amplifier (AD-844-CFA) devices which leads to extremely low active circuit sensitivity.

- Current Differencing Transconductance Amplifier

Current differencing transconductance amplifier (CDTA) is a new active circuit element. The CDTA is free from parasitic input capacitances and it can operate in a wide frequency range due to current mode operation. Some voltage and current mode applications using this element have already been reported in literature, particularly from the area of frequency filtering: general higher-order filters, biquad circuits, all-pass sections, gyrators, simulation of grounded and floating inductances and LCR ladder structures. Other studies propose CDTA-based high-frequency oscillators. Nonlinear CDTA applications are also expected, particularly precise rectifiers, current-mode Schmitt triggers for measuring purposes and signal generation, current-mode multipliers, etc.

The CDTA element with its schematic symbol in Fig 1.8 has a pair of low-impedance current inputs p, n and an auxiliary terminal z, whose outgoing current is the difference of input currents.

Here, output terminal currents are equal in magnitude, but flow in opposite directions, and the product of transconductance (g_m) and the voltage at the z terminal gives their magnitudes.

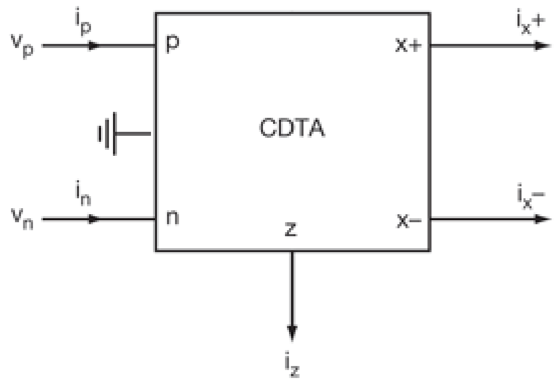


Figure 1.8 Block diagram of CDTA

Therefore, this active element can be characterized with the following equations:

$$V_P = V_N = 0 \tag{1.8}$$

$$I_Z = I_P - I_N \tag{1.9}$$

$$I_{X+} = g_m V_Z \tag{1.10}$$

$$I_{X-} = -g_m V_Z. \tag{1.11}$$

Where $V_{Z-} = I_Z Z_Z$ and Z_Z is the external impedance connected to z terminal of the CDTA. CDTA can be thought as a combination of a current differencing unit followed by a dual-output operational transconductance amplifier, DO-OTA. Ideally, the OTA is assumed as an ideal voltage-controlled current source and can be described by $I_X = g_m(V_+ - V_-)$, where I_X is output current, V_+ and V_- denote non-inverting and inverting input voltage of the OTA, respectively.

Note that g_m is a function of the bias current. When this element is used in CDTA, one of its input terminals is grounded (e.g., $V_- = 0$V). With dual output availability, $I_{X+} = -I_{X-}$ condition is assumed.

- CURRENT FEEDBACK OPERATIONAL AMPLIFIER

The current feedback operational amplifier (CFOA) is a type of electronic amplifier whose inverting input is sensitive to current, rather than to voltage as in a conventional voltage-feedback operational amplifier (VFA). The CFA was invented by David Nelson at Comlinear Corporation, and first sold in 1982 as a hybrid amplifier, the CLC103. The first integrated circuits CFAs were introduced in 1987 by both Comlinear and Elantec (designer Bill Gross). They are usually produced with the same pin arrangements as VFAs, allowing the two types to be interchanged without rewiring when the circuit design allows. In simple configurations, such as linear amplifiers, a CFA can be used in place of a VFA with no circuit modifications, but in other cases, such as integrators, a different circuit design is required. The circuit symbol of CFOA is as shown in fig.1.9. Its port relations can be characterized by the following matrix form:

$$\begin{bmatrix} I_y \\ V_x \\ I_z \\ V_o \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} V_y \\ I_x \\ V_z \\ I_o \end{bmatrix}$$

Therefore, this active element can be characterized with the following equations:

$$i_y = 0,\ v_x = v_y,\ i_z = i_x,\ v_o = v_z \qquad (1.12)$$

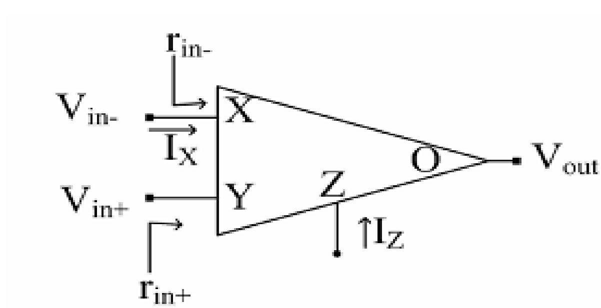


Figure 1.9: Circuit Symbol of CFOA

Current-Feedback Operational Amplifiers (CFOAs) are employed as an alternative to conventional voltage Opamp's because of their inherent advantages:

- ❖ The CFOA closed-loop bandwidth is independent of its close-loop gain, provided that the feedback resistance is kept constant
- ❖ The CFOA input and output stages work both in class AB and give high slew-rate values AD844A is CFOA chip which is commercially available in the market.

1.2 Related Work

In the early 1990s, RF integrated circuits (RFICs) were dominated by bipolar and GaAs technologies, while CMOS technologies were mainly used for baseband signal applications. As the gate lengths of CMOS devices were scaled down to one micron in the middle of 1990s, CMOS RFICs over GHz became possible. Since then, continuous scaling-down of CMOS devices to today's deep sub-micron devices has opened a new era for CMOS RFIC designs, including OTA circuits. Previous OTAs were developed in CMOS technologies due to the unique properties that CMOS can offer [5-7]:

1) Ease to implement a transconductors – MOSFETs are naturally voltage-controlled current devices.

2) High cutoff frequency – the latest record was beyond 400 GHz.

3) Well commercialized and low-cost processes.

4) Ability to integrate both digital and analog circuits, i.e., system-on-chip (SoC).

Currently, CMOS OTAs are developed in three trends: high frequency, high linearity and low power [4].

1.3 Scope of Project

Our approach is to use the Tunable OTA in the field of signal processing by implementing subsystems like filters, oscillators, etc. with it. Tunable OTA increases the bandwidth, linearity, gain etc. without much increasing the complexity of the circuit. The beauty of the circuit is its simple approach to overcome the limitations of conventional Tunable OTA and its applications.

1.4 Thesis Structure

The dissertation is divided into six chapters, each having ample information for comprehending the concept of this project:

Chapter 1: Introduction, need of ASP using different active blocks and its advantages & disadvantages, objectives of the project and outline of the thesis.
Chapter 2: This chapter describes the complete details of OTA, its internal structure and their individual characteristic. It also discusses the Advantages of OTA. Terminal characteristics of these circuits have been verified through PSPICE simulations.
Chapter 3: This chapter deals with different CMOS Realization of OTA and their proprieties and advantages & disadvantages.
Chapter 4: Discusses about the different filters topologies using OTA.
Chapter 5: This chapter describes the OTA based Voltage mode third order Quadrature Oscillator In this chapter two topologies of OTA based third order Quadrature oscillators (QO) are proposed. The proposed oscillator is designed using combination of a second order low pass filter followed by an integrator. Workability of the proposed QOs is verified through PSPICE simulations using 0.5µm AGILENT CMOS process parameters. The total harmonic distortion (THD) for both the QO designs is found to be less than 1%.
Chapter 6: This chapter will conclude on the results from all the simulations. Discussions and analysis are included in this section. There is, also, a discussion on the suggestion for future work.

CHAPTER 2

Literature Review

2.1 Introduction

An Ideal Operational transconductance amplifier (OTA) is a voltage controlled current source with constant transconductance and infinite input/output impedances, that takes the differential voltage as input and provide current as the output, on the other hand conventional Op-Amp, which is a voltage-controlled voltage source [7-9]. However, they become inaccurate when a practical OTA at high frequency or with large input signal is concerned. Effects of high frequency parasitic and active device nonlinearity have to be considered in such cases. The high-frequency and linearity issues of OTAs will be addressed in the following chapters.

OTA is a transconductors type device, which means that the input voltage controls an output current by controlling the device transconductance. The output current is related to the differential input voltage as:

$$I_O = g_m\,(V_+ - V_-) \qquad (2.1)$$

What is important and useful about the OTA's transconductance parameter g_m is that it is controlled by an external current, the amplifier bias current, I_{bias}, and is given as follows:

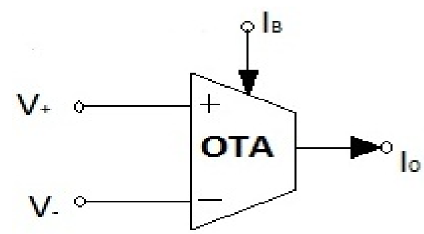
$$g_m = \sqrt{\mu_o C_{ox} \frac{W}{L} I_{Ref}} \qquad (2.2)$$

Thus its transconductance can be varied by varying just the bias current. OTA is basically an op-amp without an output buffer and hence can drive only capacitive load.

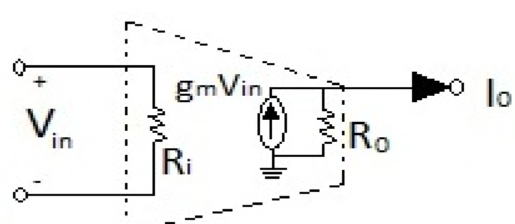
It has been used as a replacement for the conventional op-amp in both first and second-order active filters because of its several advantages over the conventional op-amp. Conventional op-amp required lots of complex feedback networks to attain stable configuration. Figure 2.1 shows the basic schematic and equivalent circuit of OTA.

Depending on the input and output configurations, OTAs can be categorized into three types: single input/output, differential-input single-output and differential input/output (fully differential). The other possible type, single-input differential-output, is not practical and has not been used in previous applications.

In these three types of OTAs, the transconductance g_m can be tuned via their DC current bias I_{bias}. The single input/output transconductors is the simplest to implement, e.g. a single-NMOS common-source transconductors [6-7]. The simplicity of this type of OTA makes it interesting for high frequency implementation, while most previous works preferred the differential configurations due to their common-mode rejection and their flexibility to engage feedback configurations. The latter two types have nearly the same performance and can be replaced by each other in most circuits using OTAs. The implementations of these three types will be further described in Chapter 3.



(a)



(b)

Figure 2.1: OTA (a) Symbol (b) Equivalent Circuit

Characteristics of Ideal OTA can be summarized as follows [8]

Input impedance (Z_{in}) = ∞

Output Impedance (Z_O) = ∞

Bandwidth (BW) = ∞

Perfect balance, i.e., $I_o = 0$ when $V_1 = V_2$

Transconductance (g_m) is finite and controllable with the amplifier bias current I_{Bias}.

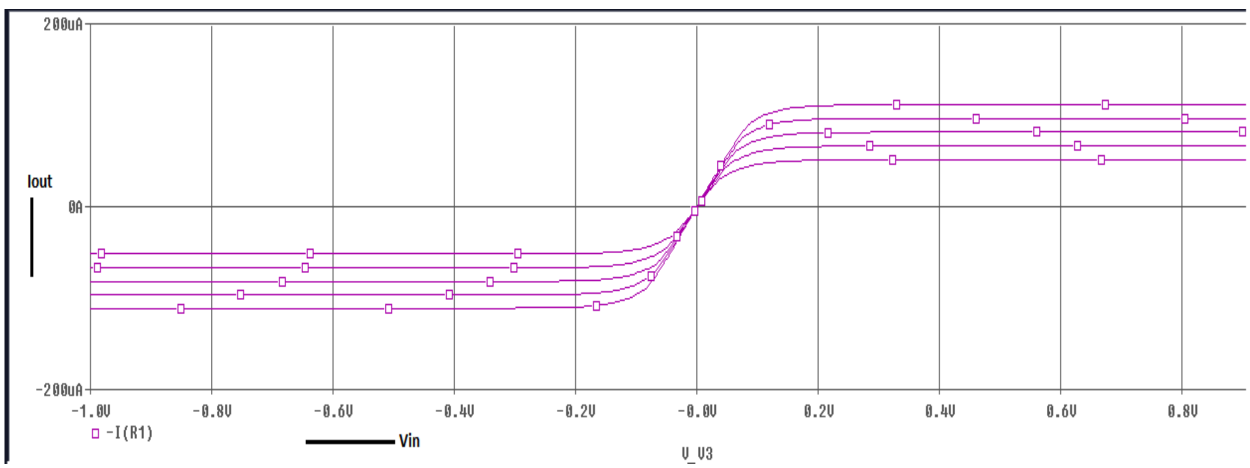


Figure.2.2: DC transfer characteristics of OTA

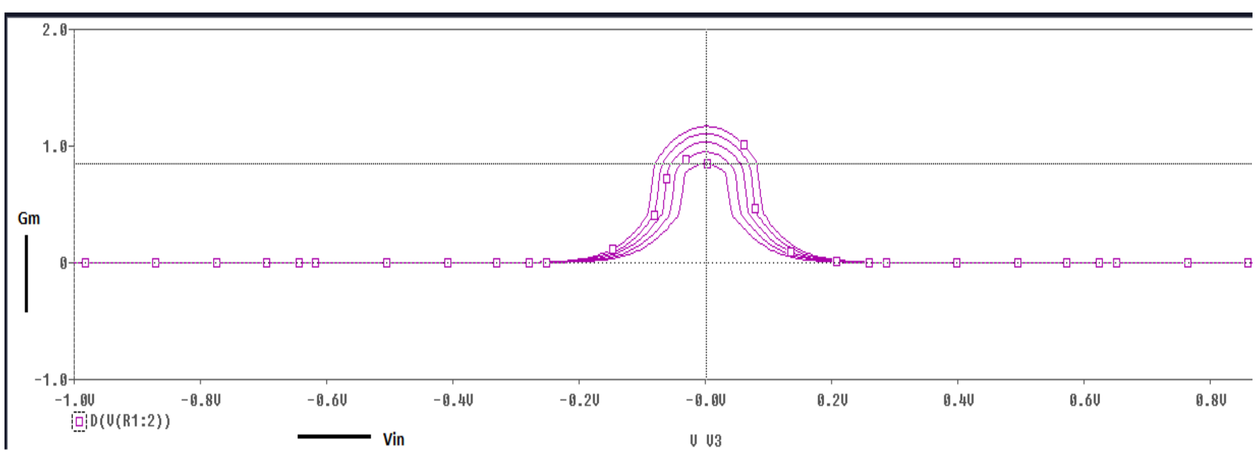


Figure 2.3: Transconductance of OTA.

2.2 Internal Circuit of OTA

It basically consists of three kinds of sub-circuit:

1. Differential pair

2. Current source

3. Current mirrors

The internal circuit of OTA consists of a differential input stage while the remaining part of the OTA is composed of current mirrors. The geometry of these mirrors is kept such that the current gain is unity. Thus the output current is precisely defined by the differential input amplifier. The output transfer characteristic of the amplifier is shown in the figure 3. The shape of this characteristic remains constant and is independent of supply voltage. Only the maximum current is modified by the bias current.

2.2.1. Differential pair

Differential amplifiers are well-known electronic devices for amplifying a voltage difference between two input signals. Differential amplifiers are utilized to amplify, and produce an output signal which is a function of the difference between two differential, or complementary, input signals and to thereby enable the detection of relatively weak signal levels while inherently rejecting noise common on the differential input lines. A differential amplifier usually comprises two electrical paths that are independently coupled to a voltage source at one end, and are together coupled to a voltage or current source at an opposite end. Each electrical path usually comprises a transistor element and a resistance element. They are used to amplify analog, as well as digital signals, and can be used in various implementations to provide an amplified output in response to differential inputs. Many electronic devices use them internally [7].

The output of an ideal differential amplifier is given by:

$$V_O = A_d (V_+ - V_-) \qquad (2.3)$$

Where V+ and V- are the input voltages and A_d is the differential gain.

In practice, however, the gain is not quite equal for the two inputs. This means, that if V_+ and V_- are equal, the output will not be zero, as it would be in the ideal case.

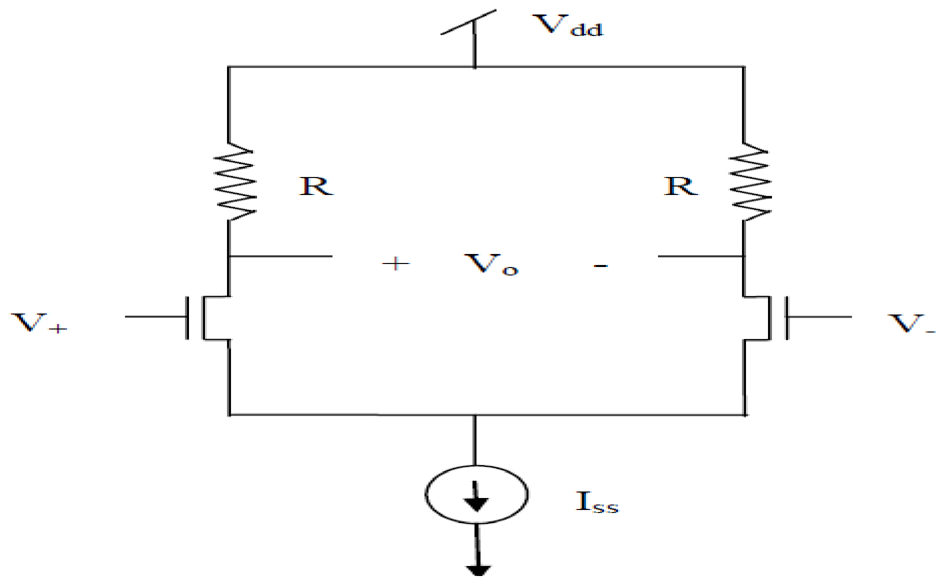


Figure 2.4: Differential Amplifier

A more realistic expression for the output of a differential amplifier thus includes a second term.

$$V_o = \frac{A_d(V+ - V-) + A_c\,(V+ - V-)}{2} \qquad (2.4)$$

Where, V+ and V- are the input voltages and A_d is the differential gain, A_c is the common-mode gain of the amplifier.

As differential amplifiers are often used when it is desired to null out noise or bias-voltages that appear at both inputs, a low common-mode gain is usually considered good.

Common-mode rejection ratio: It is usually defined as the ratio between differential-mode gain and common-mode gain, indicates the ability of the amplifier to accurately cancel voltages that are common to both inputs. Common-mode rejection ratio (CMRR):

$$\mathrm{CMRR} \approx \frac{\mathbf{A_d}}{\mathbf{A_c}} \tag{2.5}$$

Differential amplifier circuits are widely used in the electronics industry and are generally preferred over their single-ended counterparts because of the following advantages:

1. Better common-mode noise rejection.
2. Reduced harmonic distortion.
3. Increased output voltage swing.

2.2.2 Current mirror

A current mirror is a circuit designed to copy a current through one active device by controlling the current in another active device of a circuit, keeping the output current constant regardless of loading [7]. The current being 'copied' can be, and sometimes is, a varying signal current. Conceptually, an ideal current mirror is simply an ideal current amplifier. The current mirror is used to provide bias currents and active loads to circuits.

Operation M1 in Figure 6, will operate in saturation region as its source and drain are shorted, hence its drain current I_{D1} is neglecting the channel length modulation will be

$$I_{D1} = I_{Ref} = \frac{\beta_1\,(V_{GS} - V_{Th})^2}{2} \tag{2.6}$$

$$I_{D2} = \frac{\beta_2(V_{GS} - V_{Th})^2}{2} \tag{2.7}$$

From eqn (2.6) and (2.7), we can get that

$$I_{out} = \frac{\beta_1}{\beta_2} * I_{Ref} = \frac{\frac{W}{L}_1}{\frac{W}{L}_2} * I_{Ref} \tag{2.8}$$

Thus the topology presented allows precise copying of the current with no dependence on the process and temperature. The ratio of I_{OUT} and I_{REF} is given by the ratio of device dimensions, a quantity that can be controlled with reasonable accuracy.

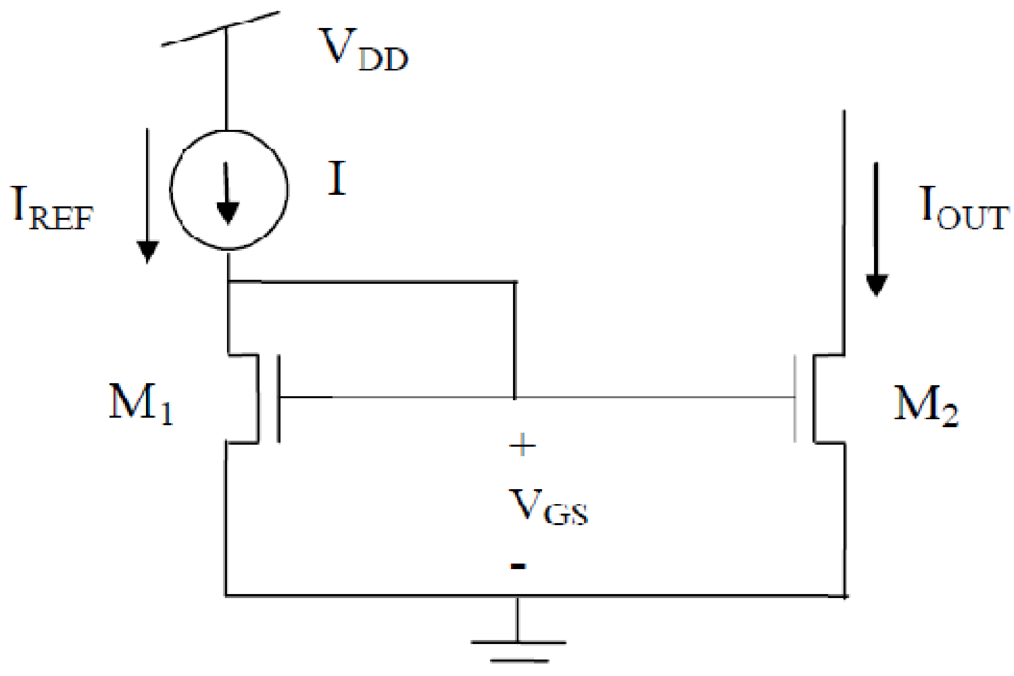


Figure 2.5: Current Mirror

Desired features

1. Generate an output current equal to input current multiplied by desired current gain factor

2. Current gain is independent of input frequency

3. Output current independent of output voltage to common node

2.3 OTA Applications

Together with the developments of CMOS OTAs, OTA applications were also explored in a variety of interesting analog circuits, from simple components such as variable resistors and active inductors, to more complicated circuits, e.g. filters and oscillators. Some of the applications either have been demonstrated or are potential at RF/microwave frequencies. These applications and their theories are briefly reviewed below.

2.3.1 Variable Resistors

Variable resistors are the simplest applications of the OTAs. There are two types of OTA variable resistors, positive and negative, depending on their feedback polarities. Figure 2.6 presents positive resistor realized using negative feedback on the OTAs. One is a single-ended version and the other is a floating version. The input impedance of these two versions can be derived by assuming an input voltage V_i and an input current I_i :

$$Z_i = \frac{V_i}{I_i} = \frac{1}{g_m} \tag{2.9}$$

Note from (2.9) that this resistor can be variable by changing g_m. OTA variable resistors with capacitors can compose OTA-C filters, where the variable resistors can be used to tune the filter frequencies. More details will be described in a later section.

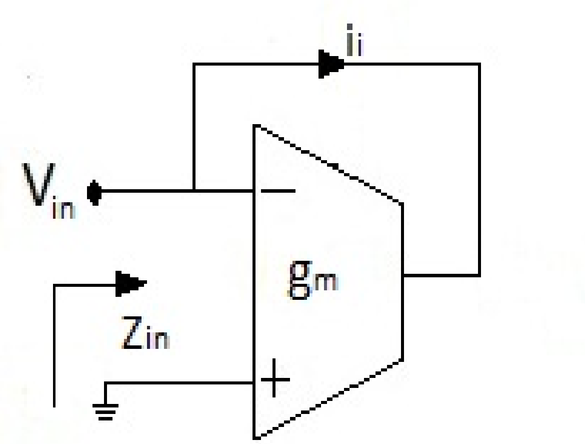


Figure 2.6: OTA variable resistors

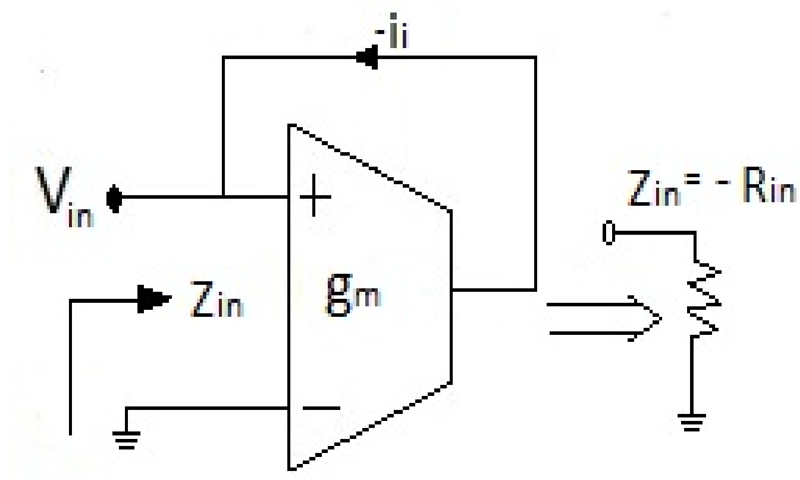


Figure 2.7 OTA Variable Negative Resistors

$$Z_i = V_i / I_i = -1/g_m \tag{2.10}$$

Because a negative feedback on an OTA in Fig. 2.7 creates a positive resistor, it is easily understood that positive feedback on an OTA can result in a negative resistor. Note that the positive feedback inverts the input current's direction, creating the minus signs in (2.10), i.e., the negative resistor. These negative resistors can also be tuned by changing *gm*. A negative resistor can be used either to compensate a filter's loss or to provide gain for an oscillator.

2.3.2 Active Inductors

Silicon CMOS technologies are widely used for analog and RF/microwave designs due to the previously-mentioned unique properties for their active devices. In contrast to these active devices, their passive devices, especially on-chip inductors, suffer problems from their lossy silicon substrate. The on-chip inductors are so close to the low-resistance silicon substrate that they generate eddy currents in the substrate beneath them. These eddy currents, in return, cause the loss and thereby reduce the Q factors of the inductors. Low Q factors degrade circuit performance, such as phase noise and gain. On-chi inductors also consume much larger IC areas compared to the active devices.

Multiple techniques were developed to reduce the loss with on-chip inductors by using some substrate modifications, such as silicon-on-insulator (SOI)

Compared to the on-chip spiral inductors, an active inductor has a higher Q factor and occupies a smaller IC area due to its composition of only active devices and capacitors, and thus is attractive to researchers. An active inductor contains an impedance inverter that "inverts" a real capacitor into a virtual inductor. Its input emulates a real inductor's voltage and current. Fig. 2.9 & Fig.2.10 illustrates diagrams for a single-ended active inductor and a floating active inductor using OTAs [2].

In Fig. 2.9, the two OTAs in a negative feedback loop construct the impedance inverter. The voltage and the current on its right are related to each other through the capacitor *C*:

$$I_o = \frac{V_o}{Z_L} = V_o sC \tag{2.11}$$

Where $I_o = g_{m1}V_i$, $V_O = \frac{I_i}{g_{m2}}$ = according to the two OTAs. Substituting them into (2.11) yields

$$Z_i = \frac{V_i}{I_i} = \frac{SC}{g_{m1}\, g_{m2}} \tag{2.12}$$

It is clear that (2-12) is an expression of an inductor, with an inductance value of:

$$L = \frac{SC}{g_{m1}\, g_{m2}} \tag{2.13}$$

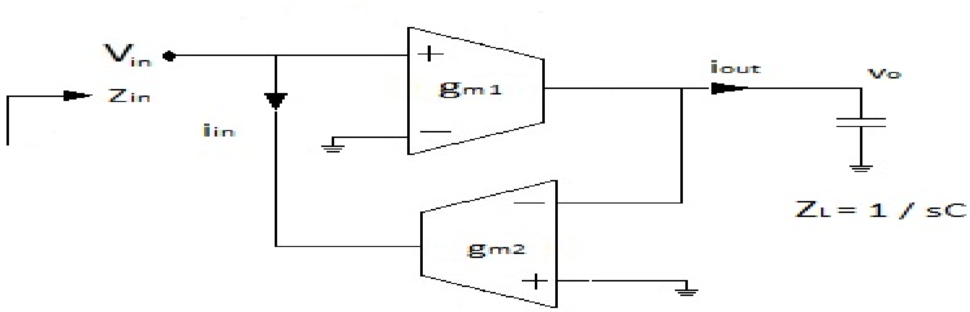


Figure 2.8: Single Ended Active Inductor

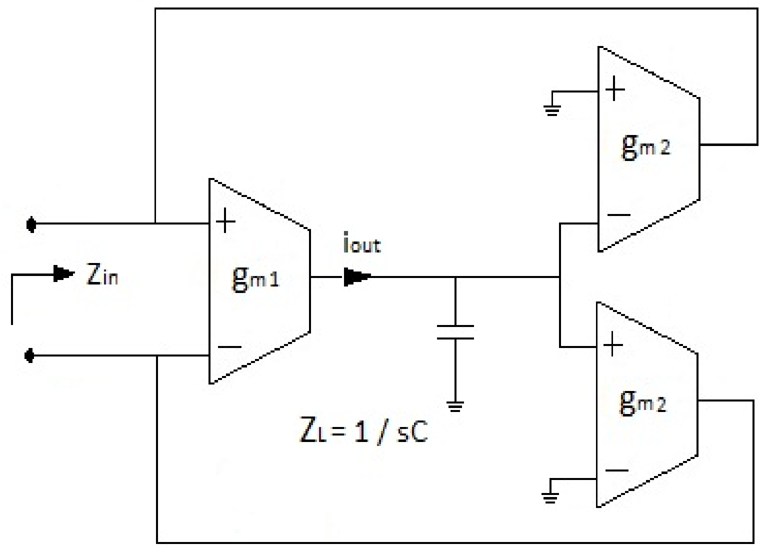


Figure 2.9: Floating Active Inductor

The derivation for the floating active inductor in Fig. 2.10 results in the same expression. The virtual inductor can be electronically tuned by changing the two OTAs' transconductance, g_{m1} and g_{m2}. This tunable characteristic offers designers another choice to tune circuit frequencies.

2.4 Advantages of OTA

1. Transconductance can be controlled electronically by adjusting the biasing current. Hence, provides the applications that their features can be tuned electronically

2. The OTA-C filter uses transconductors and capacitors as integrators which work in open loop, thus potentially having the best high frequency characteristics compared to active-RC..

3. The wide continuous tuning ability of the OTA also allows the filter to maintain precise filtering characteristics against process variation and temperature drift.

CHAPTER 3

Different OTA Cells

3.1 CMOS realization of OTA

CMOS technologies are very convenient for implementing OTAs because their MOSFETs are inherently voltage-controlled current devices. A variety of CMOS OTAs with different topologies have been developed for different purposes so far [10-17]. According to their input/output topologies though, they can be categorized into the previously mentioned three types, i.e., single input/output, differential-input single-output, and differential input/output. These three types of CMOS OTAs as well as their advantages and disadvantages are generally described below.

3.1.1 Simple Transconductance Amplifier

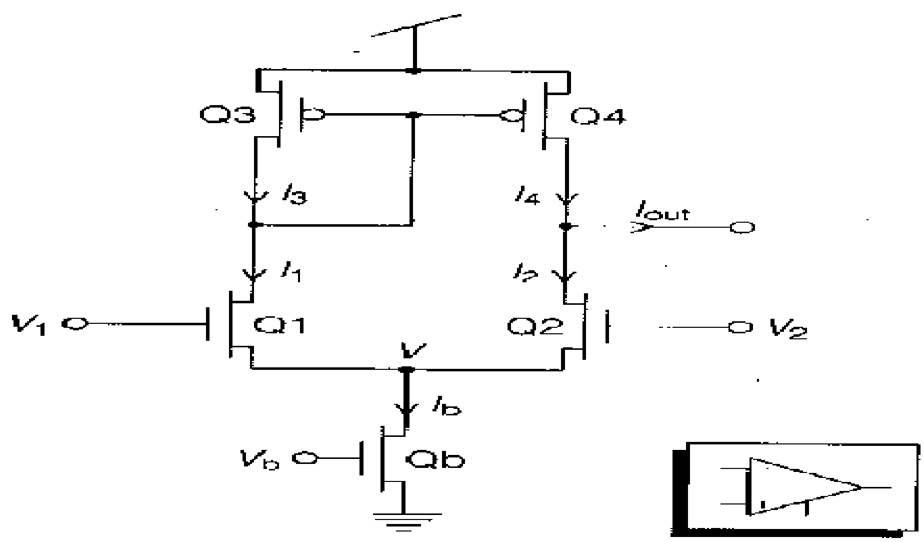


Figure 3.1: Simple Single Stage OTA

This circuit consists of a differential pair and a single current mirror. Differential Pair subtract the drain current I_1 & I_2. The current I_1 drawn out of Q3 is reflected as an equal current out of Q4[15].

This single stage OTA is less complex compare to other types of OTA topology. Because of its less complex property its speed is higher compare to other topology. Low impedance also leads to high unity gain bandwidth and high speed

LIMITATIONS:

- Mismatch between Transistors
- Deviation of Transistors from perfect current source behavior.
- Lower gain due to the fact that output impedance of this type configuration is relatively low.

3.1.2 Balanced OTA (BOTA)

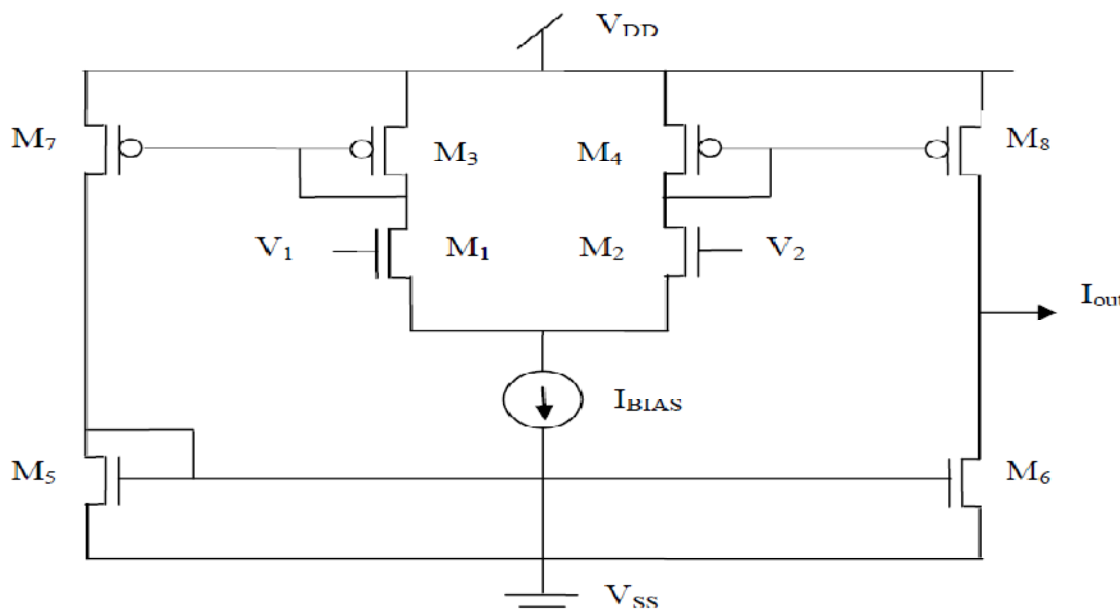


Figure 3.2: Schematic of Balanced OTA

It is the simplest OTA structure implemented using a differential amplifier and three pairs of current mirrors [15, 18, 20]. Balancing is achieved by coupling between the device sources and current mirrors. All the MOS devices are operated in Saturation region, the transistor Drain Current I_D is characterized by a square law model given by:

$$I_D = \beta(V_{GS} - V_{TH})^2, \text{ for } V_{GS} > V_{TH} \tag{3.1}$$

And $I_D = 0$, for $V_{GS} \leq V_{TH}$

β is the transconductance parameter of the transistor,

$$\beta = \mu_n C_{ox} \frac{W}{2L} \tag{3.2}$$

Where μ_n= carrier mobility (electrons)

Cox is the gate oxide capacitance per unit area, W is the effective channel width; L is the effective length of the channel. The Balanced OTA in above figure is formed by coupling the device sources and current mirrors, where,

$$V_{in}=V_2-V_1 \tag{3.3}$$

i_o is the differential output current and I_{BIAS} is the bias current. Let us assume that all the transistors have unity gain (i.e. W/L=1).

$$i_o=i_2-i_1 \tag{3.4}$$

From the differential amplifier, the output current in terms of input voltage, bias current and process parameters can be derived and the expression for the output current is given by:

$$i_o=g_m V_{in}=\sqrt{2\beta I_{BIAS}} \tag{3.5}$$

3.1.3 Folded cascode OTA

Because of their high gain and speed, folded cascode (FC) operational transconductance amplifiers (OTAs) play a central role in many analog systems.

Fig. 3.3 shows the architecture of Folded Cascode op-amp. It is cascade of a differential transconductance stage with a current stage followed by a cascade current-mirror load

This op-amp uses cascading in output stage combined with an unusual implementation of the differential amplifier to achieve good input common mode range. Use of a cascade mirror leads to achieve the gain of two-stage and allows for self- compensation. The basic form of an n-channel input, folded cascade op-amp is shown in fig. 3.3

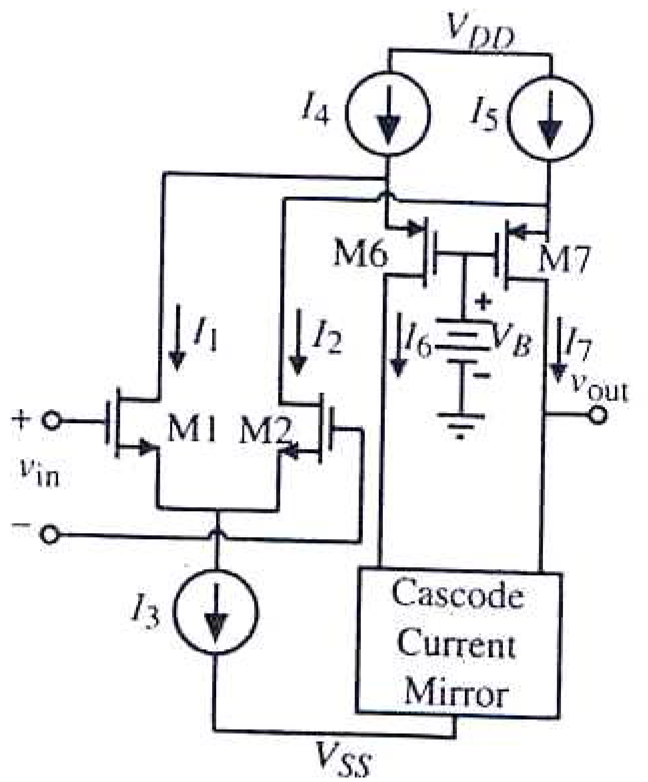


Figure 3.3: Schematic of conventional folded cascode amplifier.

The folded cascade does not require perfect balance of currents in differential amplifier because excess DC current can flow into or out of current mirror. Because the drains of M1 and M2 are connected to drains of M4 and M5, a positive input common mode voltage that can be achieved by using current source loads is achieved. The bias currents I3, I4 and I5 of folded cascade op-amp should be designed so that the DC current in cascade mirror never goes to zero. If the current should go to zero, this requires a delay in tuning the mirror back on because of parasitic capacitances that must be charged.

For example, suppose Vin is large enough so that M1 is on and M2 is off. Then, all of I3 flows through M1 and none through M2, resulting in I1 = I3 and I2 = 0. If I4 and I5 and not greater than I3, then the current I6 will be zero. To avoid this, the values of I4 and I5 are normally between the values of I3 and I3.

The gain that can be achieved by a single stage is around 40 dB. Thus, in order to achieve 80 dB or so it is necessary to use a cascade of two stages. However, two stages bring about two poles one close to the other and this requires compensation network, besides increasing the global complexity, reduces the design flexibility. A cascade with cascade load permits us to achieve high gain without the disadvantage of having two poles one close to each other. Therefore the use of cascode based OTA is an interesting solution alternative to the two stages OTA.

Thus two options have been left, one is Telescopic configuration and the other one is folded cascode configuration. The primary advantage of folded structure lies in the choice of voltage levels because it does not “stack” the cascode transistor on the top of the input device. Further Telescopic OTA suffers with limited output swing, and there is a difficulty in shorting the input and output (which is the foremost requirement of my filter).

3.1.4 Telescopic OTA

The Telescopic Cascode OTA configuration is as shown in figure 3.4. The OTA is designed to be fully-differential for a number of reasons. This doubles the effective output swing and the amount of current available for slewing. Because the signal power quadruples while the noise power only doubles, the dynamic range is also doubled. Moreover, fully-differential circuits have been shown to effectively attenuate even-order harmonic distortion, substrate noise, supply noise, and common mode disturbances [11].

The single-stage telescopic OTA employs differential input pairs (M1 and M2), cascode devices (M3 and M4) and current source loads (M5, M6, M7 and M8). An input device (M1 and M2) generates drain current proportional to input voltages (V_{inp} and V_{inn}). The cascode devices simply routes the current to current loads. Furthermore, cascode devices will increase the output resistance. Therefore, the OTA gain increases. OTA is loaded with PMOS cascode current sources (M5, M6, M7 and M8) to increase output impedance

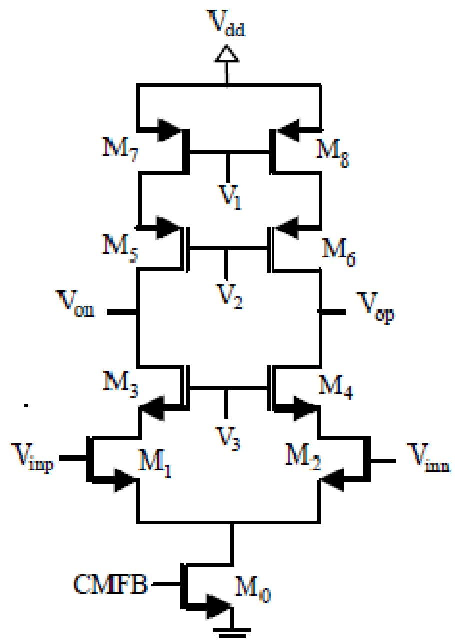


Figure 3.4: Schematic of Telescopic OTA

Although telescopic operational amplifier has smaller swing, which means reduced dynamic range, which means that the Telescopic OTA is a better candidate for low power, low noise single stage OTA. The single stage architecture normally suggests low power consumption [18].

Advantages:

1. The telescopic cascode achieves a gain similar to the one of the two stages architecture, without having two poles one close to each other.

2. All the nodes, excluding the output, show a pretty low small signal resistance.

3. Its lower noise factor makes it a better candidate for low power, low noise single stage operational transconductance amplifier.

Disadvantages:

1. Telescopic op-amp has severely limited output swing and hence the dynamic range. It is smaller than that of Folded Cascode because the tail transistor directly cuts into output swing from both side of op-amp.

2. Higher gain at the cost of output swing and additional poles

3.1.5 Dual Output OTA

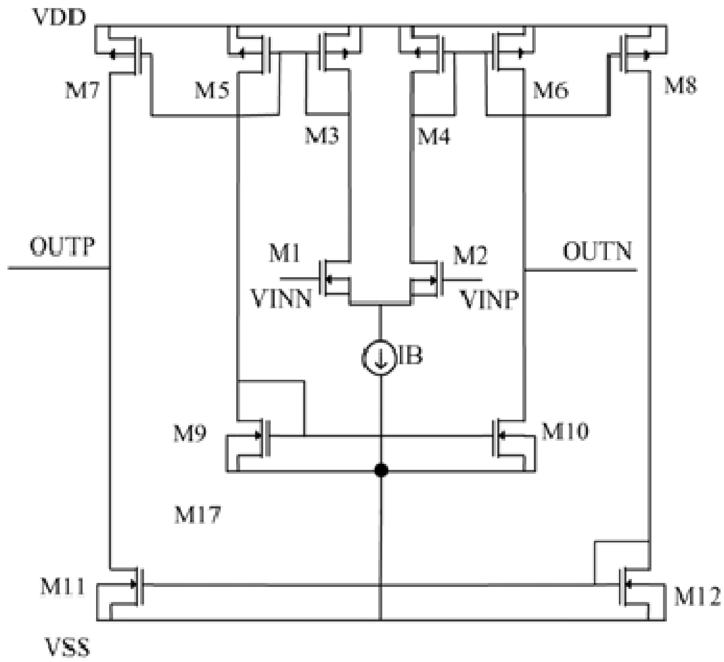


Figure 3.5: CMOS realization of symmetrical DO- OTA

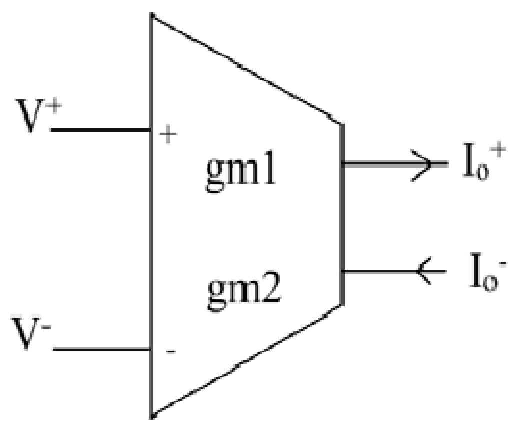


Figure 3.6: Circuit symbol of the DO-OTA

Ideally, DO-OTA is assumed as an ideal voltage controlled current source and can be described by following equations [18].

$$I_{O+} = g_{m1}\,(V_{+} - V_{-}) \quad (3.6)$$

$$I_{O-} = g_{m2}\,(V_{-} - V_{+}) \quad (3.7)$$

Generally, the transconductance of the DOOTAs are chosen as gm1=gm2 similar to the

The Two Output of the DO-OTA belongs to direct current and inverted current. The inverted current is the current going into the DO-OTA and can be used for feedback. The direct current is the current going out of the device and can be used as an output or as an input to another block.

An important drawback of these active devices is the finite output resistance, which is parallel with the load capacitor causing a lossy integration, thus generating filtering errors. Therefore very high output impedance output stages are required both to enable filtering at low frequencies and to reduce filtering errors.

CHAPTER 4

Different FILTER Topologies

4.1 Introduction

A filter is defined as an electric network, which passes or allows unattenuated transmission of electric signal within certain frequency range and stops or disallows transmission of electric signal outside this range. Filters are signal conditioners, each function by accepting an input signal, blocking pre-specified frequency components, and passing the original signal minus those components to the output. For example, a typical phone line acts as a filter that limits frequencies to a range considerably smaller than the range of frequencies human beings can hear, that’s why listening to CD-quality music over the phone is not as pleasing to the ear as listening to it directly.

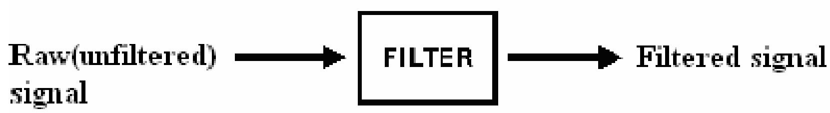


Figure 4.1: Simple block diagram of filter

Filters are the indispensable parts, used virtually in every modern electronic system. A filter is an electronic device used to select a particular pass band range. Signals within that range are allowed to pass while the signals outside that range are disallowed. According to circuit theory a more general definition is, a filter is an electrical network that alters the amplitude and/or phase characteristics of a signal with respect to frequency. Ideally, a filter will not add new frequencies to the input signal, nor will it change the component frequencies of that signal, but it will change the relative amplitudes of the various frequency components and/or their phase relationships. Filters are often used in electronic systems to emphasize signals in certain frequency ranges and reject signals in other frequency ranges. Most A/D converters (ADCs) are preceded by a filter, which removes frequency components that are beyond the ADC's range. Such a filter has a gain, which is dependent on signal frequency. Since their frequency domain effects on signals define filters, it makes sense that the most useful analytical and graphical descriptions of filters also fall into the frequency domain.

Thus, curves of gain vs. frequency and phase vs. frequency are commonly used to illustrate filter characteristics, and the most widely used mathematical tools are based in the frequency domain. The frequency-domain behavior of a filter is described mathematically in terms of its transfer function or network function. This is the ratio of the Laplace transforms of its output and input signals. The voltage transfer function H(s) of a filter can therefore be written as [18-22]:

$$H(s) = \frac{V_{out}(S)}{V_{in}(S)} \tag{4.1}$$

where $V_{in}(s)$ and $V_{out}(s)$ are the input and output signal voltages and s is the complex frequency variable..

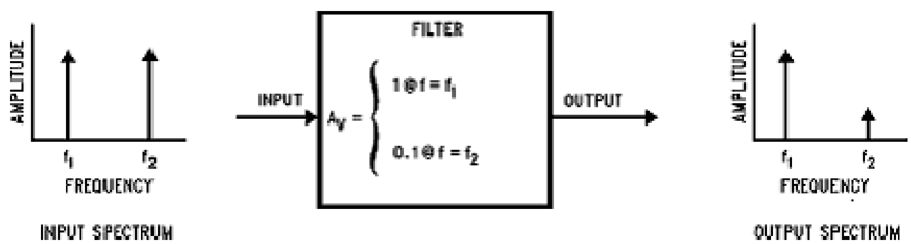


Figure 4.2 Frequency spectrum of filter

4.2 Frequency Response

Simple filters are usually defined by their responses to the individual frequency components that constitute the input signal. There are three different types of responses. A filter's response to different frequencies is characterized as pass band, transition band, or stop band. The pass band response is the filter's effect on frequency components that are passed through (mostly) unchanged. Frequencies within a filter's stop band are, by contrast, highly attenuated. The transition band represents frequencies in the middle, which may receive some attenuation but are not removed completely from the output signal [23-29].

In fig. 4.3, which shows the frequency response of a low pass filter, ω_p is the pass band ending frequency, ω_s is the stop band beginning frequency, and As is the amount of attenuation in the stop band. Frequencies between ω_p and ω_s fall within the transition band and are attenuated to some lesser degree.

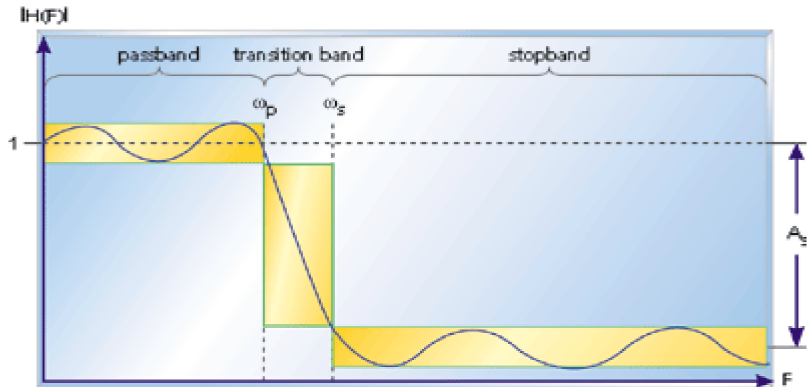


Figure 4.3: Response of a low pass filter to various input frequencies

Figure 4.3 Response of a low pass filter to various input frequencies Given these individual filter parameters, one can generate the required signal processing equations and coefficients for implementation on a DSP. There are some additional terms, which are needed to be introduced. Ripple is usually specified as a peak-to-peak level in decibels. It describes how little or how much the filter's amplitude varies within a band. Smaller amounts of ripple represent more consistent response and are generally preferable.

Transition bandwidth describes how quickly a filter transitions from a pass band to a stopband, or vice versa. The more rapid this transition, the higher the transition bandwidth; and the more difficult the filter is to achieve. Though an almost instantaneous transition to full attenuation is typically desired, real-world filters don't often have such ideal frequency response curves. There is, however, a tradeoff between ripple and transition bandwidths, so that decreasing either will only serve to increase the other.

4.3 Classification of Filters

Digital Filters and Analog Filter:

A digital filter takes a digital input, gives a digital output, and consists of digital components. In a typical digital filtering application, software running on a digital signal processor (DSP) reads input samples from an A/D converter, performs the mathematical manipulations dictated by theory for the required filter type, and outputs the result via a D/A converter [20-21].

An analog filter, by contrast, operates directly on the analog inputs and is built entirely with analog components, such as resistors, capacitors, and inductors. The performance of analog filters is directly related to the quality of the components used and the circuit design. Operational amplifiers are commonly used to increase the performance of these filters..

Passive Filters:

Passive filters are made up of passive components like resistors, capacitors, and inductors, so they are referred to as passive filters. A passive filter uses no amplifying elements (transistors, operational amplifiers, etc.).

Advantages

- Provide simplest implementation of a given transfer function.
- Require no power supplies.
- Can work well at very high frequencies.
- Used in applications involving large current or voltage levels.
- Generate little noise when compared with active filters.

Disadvantages

- Cannot provide signal gain.
- Input impedances can be lower than desirable, and output impedances can be higher for some applications, so buffer amplifiers may be needed.
- Require inductors for the synthesis of most useful passive filter characteristics and, can be prohibitively expensive if high accuracy (1% or 2%) is required.
- Tuning these to the required values is time-consuming and expensive.
- Complex passive filters (higher than 2nd-order) can be difficult and time-consuming to design.

Active Filters:

Active filters use amplifying elements, especially op amps, with resistors and capacitors in their feedback loops, to synthesize the desired filter characteristics. Active filters can have high input impedance, low output impedance, and virtually any arbitrary gain. They are easier to design than passive filters. They do not require any inductor. Still, the problems of accuracy and value spacing also affect capacitors. At high frequencies their performance is limited by the gain-bandwidth product of the amplifying elements. Active filters generate noise due to the amplifying circuitry, but this can be minimized by the use of low-noise amplifiers and by designing the circuit carefully.

The Switched-Capacitor Filter:

Switched-capacitor filters are clocked, sampled-data systems; the input signal is sampled at a high rate and is processed on a discrete-time, rather than continuous, basis. Switched capacitors can, due to their architecture, be very flexible. If used properly, they can be an excellent alternative to both discrete and integrated continuous-time filters. The operation of switched-capacitor filters is based on the ability of on-chip capacitors and MOS switches to simulate resistors The primary weakness of switched-capacitor filters is that they have more noise than active filter at their outputs.

Mathematical Representation:

Mathematically, filters can be categorized into first order, second order and higher order filters [8]. This categorization is on the basis of the transfer function obtained. The generic transfer function of any order filter can be represented as under:

$$T(s) = \frac{a_M s^M + a_{M-1} s^{M-1} + a_{M-2} s^{M-2} + \cdots + a_0}{b_N s^N + b_{N-1} s^{N-1} + b_{N-2} s^{N-2} + \cdots + b_0}$$

T(s) is a rational function of complex variable‘s’ with real coefficients. The numerator and the denominator have a degree of *M* and *N* respectively. For a stable filter operation, the degree of the numerator should be less than or equal to that of the denominator i.e. *M* £ *N*. The coefficients in the numerator and the denominator i.e. a_0, a_1,...,a_M and b_0, b_1, b_2,..., b_N are real numbers. *N* is considered to be the order of the filter. Solving for the roots of the equation determines the poles (denominator) and zeros (numerator) of the circuit. Each pole will provide a –6 dB/octave or –20dB/decade response. Each zero will provide a +6 dB/octave or +20 dB/decade response. These roots can be real or complex.

When they are complex, they occur in conjugate pairs .These roots are plotted on the s plane (complex plane) where the horizontal axis is real axis and the vertical axis is imaginary axis. The distribution of these roots on the s plane indicates a number of aspects regarding the circuit as indicated below. In order to ensure stable operation, all poles must be on the left side of the plane. If zero is located on the origin *i.e.* a zero in the numerator, the filter corresponds to a high-pass or a band-pass filter. In a high-pass filter the transfer function shows that both zeros are at s = 0, while in a band-pass filter, one transmission zero is present at s = 0 while the other is present at s = ¥.

First-Order Filters:

The general first-order transfer function is expressed as :

$$T(s) = \frac{a_1 s + a_0}{s + \omega_o}$$

which characterizes a first-order filter with a pole at $s = -\omega_0$ and a transmission zero at $s = -a0/a1$.

A first-order section can be built in a variety of ways. The simplest design uses a passive RC configuration. The centre frequency of this filter is 1/(2_RC). Correspondingly, an active RC configuration can also be used. Active RC realizations are considerably more versatile than their passive counterparts.

Second-Order Filters:

Second-order filters are also known as biquadratic filters. The standard transfer function of such filters can be expressed as [8]:

$$T(s) = \frac{a_2 s^2 + a_1 s + a_0}{s^2 + \left({}^{\omega_0}/_{Q}\right) s + \omega_0{}^2}$$

Depending on the constants a2, a1 and a0, filters are also categorised as low-pass (a2 = a1 = 0), high-pass (a1 = a0 = 0), band-pass (a2 = a0 = 0) and band-stop (a1 = 0).

Filters Based on Frequency Response:

Filters are also classified according to the band of frequencies that they pass: low-pass, high pass, band-pass, band-stop and all-pass.

A low pass filter allows only low frequency signals (below some specified cutoff) through to its output and attenuates all signal components higher than the frequency cut-off. A low pass filter is handy, in that regard, for limiting the uppermost range of frequencies in an audio signal. It is the type of filter that a phone line resembles. Low-Pass filters. This filter type is useful in improving signal to noise ratio by also reducing system intrinsic noise.

A high pass filter does just the opposite, attenuate all low frequency components below the cut-off frequency and remove the dc component (0 Hz) from the signal. This is useful in removing the dc offset that may be causing an overload condition to occur. An example of high pass application is cutting out the audible 60 Hz AC power "hum", which can be picked up as noise accompanying almost any signal in the U.S.

Band pass filter allows certain band of frequency to pass. Band-Pass filters can be designed for broadband or narrow-band applications and are essentially the combination of a High-Pass and Low-Pass filter pair. The designer of a cell phone or any other sort of wireless transmitter would typically place an analog band pass filter in its output RF stage, to ensure that only output signals within its narrow, government-authorized range of the frequency spectrum are transmitted.

All pass or a phase-shift filter has no effect on the amplitude of the signal at different frequencies. Instead, its function is to change the phase of the signal without affecting its amplitude. This type of filter is particularly useful in dealing with group-delay problems or shaping the phase response of a transfer function.

Band stop filters sometimes called a Notch Filters pass both low and high frequencies and block a predefined range of frequencies in the middle; just reverse of band pass filters. This filter offers high attenuation over a narrow range of frequencies.

Filters Based on Input and Output Applied and Obtained:

Filters are further classified as single-ended type or fully-differential type depending upon the way the signal is applied and obtained. If one of the terminals at the input and output is grounded, a single ended-type filter results. If none of the signal terminals is grounded, the filter obtained is a fully differential type. Thus, a fully-differential filter has a differential input and output. A differential output permits the placement of the filter before another differential stage such as a differential amplifier. A fully-differential filter structure has several advantages over its single ended counterpart. It includes features such as greater dynamic range of the processed signal, greater attenuation of common mode signal and lower harmonic distortion. However it has a few disadvantages such as larger on-chip area, greater power consumption and a more complicated design.

4.4 Filter Structures

Very simple kind of filter architecture available is G_m-C filters, which consist of only OTA and capacitor in the network. They are easy to construct and cascadable and they are very fast.

Then there are different Biquad filters which use OTA's and different passive elements to construct a filter network.

4.4.1 Gm.-C Filter

In the modern sub-micron technology the Gm-C filters can be operated over the entire VHF/UHF range. Generally the Gm-C integrators are very fast, because generally the Gm-C integrators are very fast, because an open-loop transconductance amplifier drives the capacitive load [18]. However, unfortunately, they are not very linear. Linearization techniques exist, but often they affect the transconductors bandwidth and a linearity-bandwidth compromise must be done. The noise is another problem of Gm-C integrators. In this case, in the synthesis process the dynamic range must be optimized. Taking into account all these advantages and disadvantages, the most natural applications for the Gm-C filters are those where speed is vital, and the amplitude of signal can be kept relatively low.

The filter designed in this paper is a low pass filter. The ideal response of a low pass filter is shown in figure. All frequencies below the cut off frequency pass through the filter without obstruction. The band of these frequencies is the filter pass band. Frequencies above cut off are prevented from passing through the filter and they constitute the filter stop band. The ideal low pass filter response cannot be realized by a physical circuit.

An OTA integrator is the simplest OTA-C filter, which can be regarded as a first order low pass filter, as shown in Fig. 4.4. It consists of only an OTA and a capacitor, simpler than an OPAMP integrator. The transfer function for this integrator can be derived as

$$\frac{V_o}{V_{i1} - V_{i2}} = \frac{g_m}{sC} \tag{4.2}$$

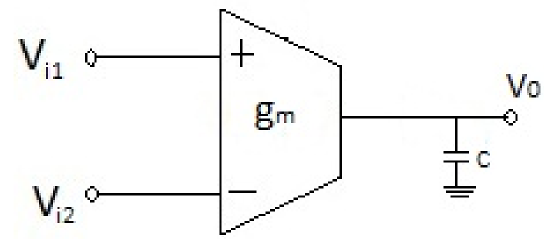


Figure 4.4: OTA Integrator

The first order low pass section plays an important role in the realization of odd order and higher order filters. The circuit for realization of first-order OTA-C low pass filter is as shown in figure.

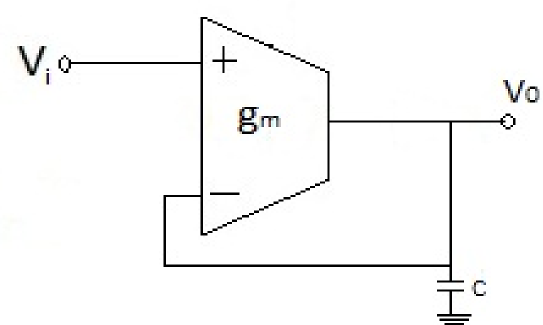


Figure 4.5: First Order Gm-C Low Pass Filter

The routine analysis of the circuit in figure yields the transfer function as:

$$T_{LP}(S) = \frac{\frac{g_m}{c}}{S+\frac{g_m}{c}} \tag{4.3}$$

$$T_{LP}(S) = \frac{H_1}{S+w_o} \tag{4.4}$$

The high pass structure of Fig. 4.6 also has a 3dB cut off frequency given by

$$f_{3dB} = \frac{g_m}{2\pi C} \tag{4.5}$$

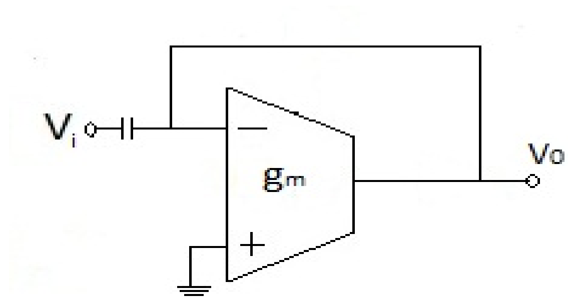


Figure 4.6: First Order Gm-C High Pass Filter

It can be observed that the characteristic networks for the low pass and high pass structures of Figs. 4.5 and 4.6 are identical, and thus they have the same pole structures. They differ only in where the excitation is applied.

4.4.2 Biquad Filters

Fig. 4.7 gives structure of second-order OTA-C filters based on the OTA integrators and capacitors. Their output expressions are shown on the right. Note that each structure contains a global negative feedback. The low-pass, band-pass, high-pass, and band-stop functions can be flexibly realized using this structure by selecting the proper input from V_A, V_B, and V_C. For example, a second order low-pass function can be obtained by choosing V_A as the input and grounding V_B and V_C in Fig. 4.4. Its transfer function becomes:

$$H(s)_{\text{Low pass}}=\frac{g_{m1}\,g_{m2}}{S^2C_1C_2+SC_2g_{m2}+g_{m1}g_{m2}} \tag{4.6}$$

which have a second-order low-pass property. The other input configurations of to realize band-pass, high-pass and band-stop functions and their transfer functions are listed below respectively

V_C=Input, V_A=V_B=0 $$H(s)_{\text{High Pass}}=\frac{S^2C_1\,C_2}{S^2C_1C_2+SC_2g_{m2}+g_{m1}g_{m2}} \tag{4.7}$$

V_B=Input, V_A=V_C=0 $$H(s)_{\text{Band Pass}}=\frac{SC_1\,g_{m2}}{S^2C_1C_2+SC_2g_{m2}+g_{m1}g_{m2}} \tag{4.8}$$

$V_A=V_C=$Input, $V_B=0$ $\qquad H(s)_{\text{Band Stop}}=\dfrac{S^2C_1\,C_2+g_{m1}\,g_{m2}}{S^2C_1C_2+SC_2g_{m2}+g_{m1}g_{m2}}$ (4.9)

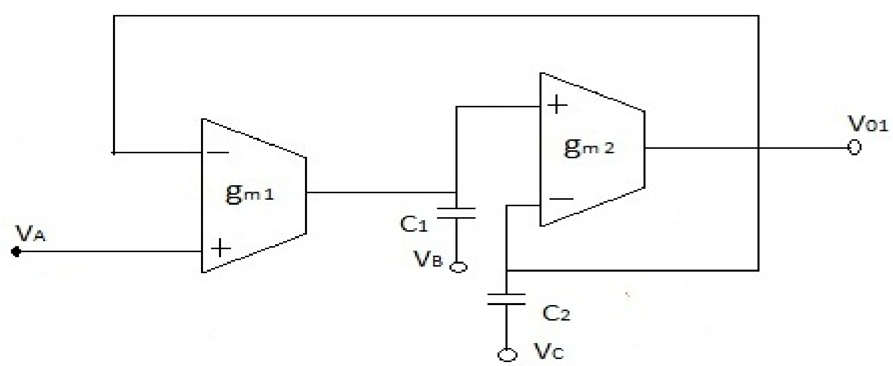


Figure 4.7: OTA-C 2nd Order Biquad Filter Structure

Tunable Voltage-Mode, multifunction biquadratic filter with three inputs single output, employing three single-ended OTAs and two Capacitors.

Active filters are widely applied in the field of electrical engineering. They can be found in crossover network used in a three-way high-fidelity loud-speaker, portable ECG detection used in front-end circuits and touch-tone telephone used for tone decoding. For integrated circuits, active filters with electronic control are interested where the natural frequency and/or the quality factor can easily be adjusted by the currents or voltages [23]. It is well known that an OTA provides an electronic tunability, a wide tunable range of its transconductance gain and simple circuitry. Furthermore, OTA-based circuits require no resistors hence they are highly suitable for IC implementation. Therefore, several biquadratic filters using OTAs have been reported.

Considering the number of input and output terminals, these filters can simply be classified into two categories: (i) a single-input multiple-output (SIMO) type and (ii) a multiple-input type. Generally, the SIMO filters can simultaneously realize three basic filter functions, i.e., low-pass (LP), band-pass (BP), and high-pass (HP), at a time without altering the connection way of the circuits and without input signal matching. However, for the realizations of all-pass (AP) and band-stop (BS) functions, additional addition and subtraction circuits are usually required. The multiple-input filter can realize multifunction outputs by altering the way in which the input signals are connected. Compared with the SIMO filter, the multiple-input filter provides a variety of circuit characteristics with different input voltage and usually does not require any parameter matching conditions. In addition, multiple-input filter may lead to a reduction in the number of active elements used. Moreover, to realize a larger variety of filter functions such as inverting and/or non-inverting-type functions, the MISO configuration seems to be more suitable than the SIMO configuration. On the other hand, the active filters using low active components are more suitable for IC implementation and also reduced the power consumption and the area of chip when they build in form IC.

Here, an electronically tunable voltage-mode universal biquadratic filter with three inputs and one output using three single-ended OTAs and two capacitors is presented. The employment of single-ended OTAs makes the circuit more suitable for IC implementation. By appropriately connecting the input terminals, the proposed circuit can provide low-pass, band-pass, high-pass, band-stop and all-pass voltage responses without changing the circuit topology. No component-matching conditions are required for realizing five standard filter responses. The natural frequency and the quality factor can be controlled orthogonally and electronically. PSPICE simulation results are performed to confirm the theoretical analysis.

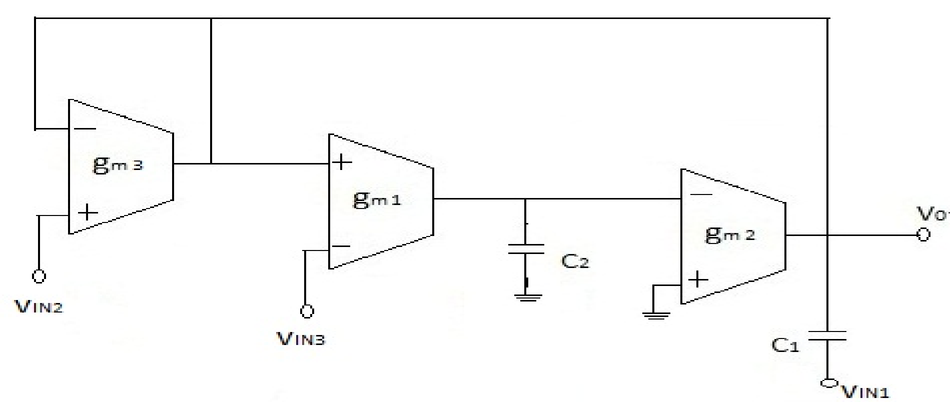


Figure 4.8: Voltage Mode Universal Biquadratic Filter

The voltage-mode universal filter is shown in Fig.4.8 It should be noted that the circuit employs only three single-ended OTAs and two capacitors. If V_{in1}, V_{in2} and V_{in3} are the input voltage signals, the output voltage Vo can be expressed as

$$V_o = \frac{S^2 C_1 C_2 V_{in1} + S C_2 g_{m3} V_{in2} + g_{m1} g_{m2} V_{in3}}{S^2 C_1 C_2 + S C_2 g_{m3} + g_{m1} g_{m2}} \tag{4.10}$$

From above equation, it can be seen that LP, BP, HP, BS and AP filters can be obtained as

LP: $V_{in3}=V_{in}$, $V_{in1}=V_{in2}=0$.

BP: $V_{in2}=V_{in}$, $V_{in1}=V_{in3}=0$.

HP: $V_{in1}=V_{in}$, $V_{in2}=V_{in3}=0$.

BS: $V_{in1}=V_{in3}=V_{in}$, $V_{in2}=0$.

AP: $V_{in1}=-V_{in2}=V_{in3}=V_{in}$.

The natural frequency (ω_o) and the quality factor (Q) can be given by

$$\omega_o = \sqrt{\frac{g_{m1} g_{m2}}{C_1 C_2}} \tag{4.11}$$

$$Q = \frac{1}{g_{m3}} \sqrt{\frac{C_1 g_{m1} g_{m2}}{C_2}} \tag{4.12}$$

Sensitivity:

Analog filters are realized by interconnecting passive components. These components may deviate from nominal design values due to some environmental effects such as temperature, humidity or chemical change occur in the network. As a result of this, filter performance will deviate from the desired values. Sensitivity is the most important criteria to compare different filter architecture [21].

Given a component Y, then in general any performance criteria P, such as quality factor or pole/zero frequency will depend on Y such that P= P(Y)

The sensitivity is defined as the deviation P caused by an errorΔY, can be expressed as:

$$S_Y^P = \frac{\frac{dP}{P}}{\frac{dY}{Y}} \tag{4.13}$$

Above equation indicates that the relative change of a performance measure P, S_Y^P , times large as the relative change of circuit parameter Yon which P depends.

Thus,
$$\frac{\Delta P}{P} = S_Y^P\left(\frac{\Delta P}{P}\right) \tag{4.14}$$

Therefore, good circuits should have low sensitivities to variation of their components.

Deviation in performance P caused by tolerance of one single component Y is measured by single parameter sensitivity. In reality, a filter consists of many components which will all contribute to the performance P variation. This is studied by multi parameter sensitivity.

X	$S_X^{\omega_o}$	S_X^Q
g_{m1}	0.5	0.5
g_{m2}	0.5	0.5
g_{m3}	0.0	-1
C_1	-0.5	0.5
C_2	-0.5	-0.5

Table 4.1: SENSITIVITIES OF CIRCUIT COMPONENTS FOR FIG.4.9

Voltage-Mode LP, HP, and BP, Biquad Filter Using Simple CMOS OTAs

In this paper, a voltage-mode universal biquadratic filter with single input and three outputs using four simple CMOS OTAs and two grounded capacitors, which is advantage in view of integrated circuit implementation, is presented. This filter can realize high input impedance voltage-mode LP, BP and HP filters from the same configuration. Critical component matching conditions are not required in the design, which provides the advantage of electronic tuning capability and is especially interest from the IC implementation point of view [24].

Also, the natural frequency (ωo) and the quality factor (Q) can be set orthogonally by adjusting the circuit components.

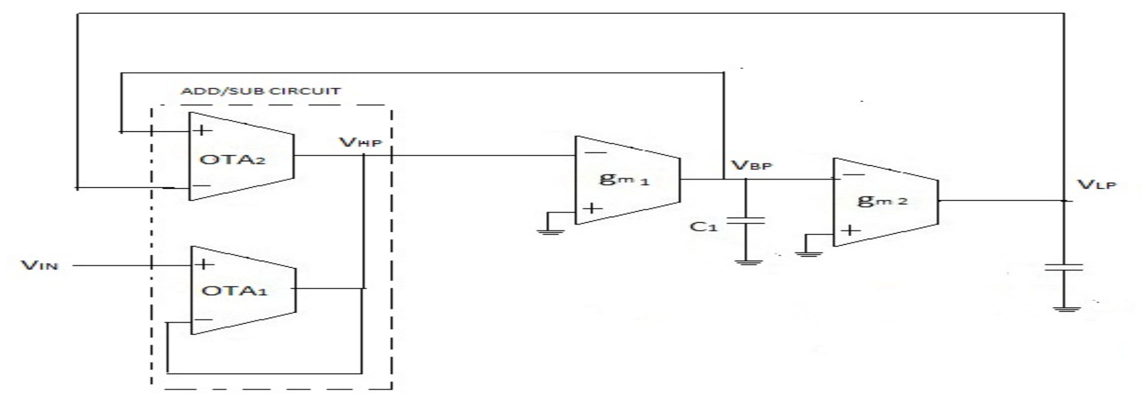


Figure 4.9: Voltage Mode SIMO Biquad Filter

Based on the use of simple CMOS OTA, the addition/subtraction voltage signal can be shown in Fig.4.7

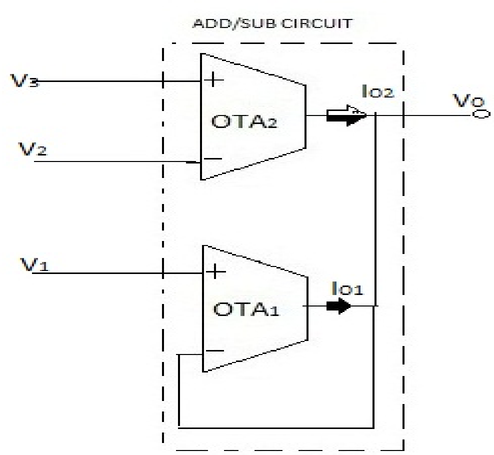


Figure 4.10: Addition/Subtraction Circuit using OTA.

From above circuit,

$$I_{o1} = g_m (V_1 - V_o) \tag{4.14}$$

$$I_{o2} = g_m (V_3 - V_2) \tag{4.15}$$

Now this gives us,

$$V_o = V_1 - V_2 + V_3 \tag{4.16}$$

By routine circuit analysis, the voltage transfer function of Fig. 4.6 can be given by

$$\frac{V_{HP}}{V_{IN}} = \frac{S^2 C_1 C_2}{S^2 C_1 C_2 + S C_2 g_{m1} + g_{m1} g_{m2}} \tag{4.17}$$

$$\frac{V_{BP}}{V_{IN}} = \frac{-S C_2 g_{m1}}{S^2 C_1 C_2 + S C_2 g_{m1} + g_{m1} g_{m2}} \tag{4.18}$$

$$\frac{V_{LP}}{V_{IN}} = \frac{g_{m1} g_{m2}}{S^2 C_1 C_2 + S C_2 g_{m1} + g_{m1} g_{m2}} \tag{4.19}$$

The natural frequency (ω_o) and the quality factor (Q) can be given by

$$\omega_o = \sqrt{\frac{g_{m1} g_{m2}}{C_1 C_2}} \tag{4.20}$$

$$Q = \sqrt{\frac{g_{m2} C1}{g_{m1} C2}} \tag{4.21}$$

X	$S_X^{\omega_o}$	S_X^{Q}
g_{m1}	0.5	-0.5
g_{m2}	0.5	0.5
C_1	-0.5	0.5
C_2	-0.5	-0.5

Table 4.2: SENSITIVITIES OF CIRCUIT COMPONENTS FOR FIG.4.10

CHAPTER 5

Voltage Mode Third Order Quadrature Oscillator

5.1 Introduction

Oscillators are of many types and come with different type of circuit configurations. Some of oscillators produce sinusoidal signals and non sinusoidal signals. Non sinusoidal oscillators, such as pulse and ramp (or saw tooth) oscillators, find use in timing and control applications. Pulse oscillators are commonly found in digital-systems clocks, and ramp oscillators are found in the horizontal sweep circuit of oscilloscopes and television sets. Sinusoidal oscillators are used in many applications, for example, in consumer electronic equipment (such as radios, TVs, and VCRs), in test equipment (such as network analyzers and signal generators), and in wireless systems [20, 30-32].

The reduction of the minimum feature size of an MOS transistor for digital VLSI circuits has been ongoing for the past few decades. As the channel length is scaled down into deep sub micrometer dimensions, the lower power supply voltage is required to ensure the device reliability. To be compatible with digital VLSI technologies, analogue integrated circuits, which can operate at low supply voltages, are also receiving significant attention.

Oscillators are used in various types of applications for these applications low value of total harmonic distortion (THD) is an essential requirement as higher harmonics have detrimental effects on electrical equipment. These higher order harmonics can also interfere with communication transmission lines since they oscillate at the same frequencies as the transmit frequency. If left unchecked, increased temperature and interference can greatly shorten the life of electronic equipment and cause damage to power systems.

Op amp oscillators are restricted to the lower end of the frequency spectrum because op amps do not have the required bandwidth to achieve low phase shift at high frequencies. Voltage-feedback op amps are limited to low kHz range since their dominant; open loop pole may be as low as 10 Hz. The new current-feedback op amps have a much wider bandwidth, but they are very hard to use in oscillator circuits because they are sensitive to feedback capacitance. Oscillators are useful for creating uniform signals that are used as a reference in applications such as audio, function generators, digital systems, and communication systems [20-21]. For these applications low value of total harmonic distortion (THD) is an essential requirement as higher harmonics have detrimental effects on electrical equipment. These higher order harmonics can also interfere with communication transmission lines since they oscillate at the same frequencies as the transmit frequency. If left unchecked, increased temperature and interference can greatly shorten the life of electronic equipment and cause damage to power systems.

Quadrature oscillators (QO) produce outputs having a phase difference of 90°. The phase-locked sine-cosine relationship of QO has useful applications in the field of telecommunications where the modulation scheme utilizes both in-phase and Quadrature components, such as in single-sideband generators, and Quadrature mixers. The QOs are also used extensively in the field of instrumentation and power electronics. For these applications low value of total harmonic distortion (THD) is an essential requirement as higher harmonics have detrimental effects on electrical equipment. These higher order harmonics can also interfere with communication transmission lines since they oscillate at the same frequencies as the transmit frequency. If left unchecked, increased temperature and interference can greatly shorten the life of electronic equipment and cause damage to power systems.

It is well known that a higher order networks as compared to lower order circuits, provide better accuracy, frequency response and distortion performance. However, it has been observed that higher order QO designs have not been explored much, as only a few third order QOs have appeared in literature in recent years. A careful observation suggests that the reported QO designs are based on forming closed loop using, (i) a second order low pass filter followed by an integrator [31,33-34,35-33]

An extensive literature review suggests though a large number of OTA based second order QOs are available in literature, yet only a few third order QO topology using OTA is reported. This QO design and is based on forming a closed loop using a second order low pass filter followed by an integrator.

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Oscillators

- ***The Oscillator Feedback Loop***

The basic structure of a sinusoidal oscillator consists of an amplifier and a RC or LC frequency selective network connected in a positive-feedback loop, such as that shown in block diagram form in Fig.5.1. Although in an actual oscillator circuit, no input signal will be present, we include an input signal here to help explain the principle of operation. It is important to note that unlike the negative-feedback loop here the feedback signal is summed with a *positive* sign. The amplitude of the generated sine waves is limited, or set, using a nonlinear mechanism, implemented either with a separate circuit or using the nonlinearities of the amplifying device itself. In spite of this, these circuits, which generate sine waves utilizing resonance phenomena, are known as linear oscillators. Circuits that generate square, triangular, pulse (etc.) waveforms, called nonlinear oscillators or function generators, employ circuit building blocks known as multivibrators**.**

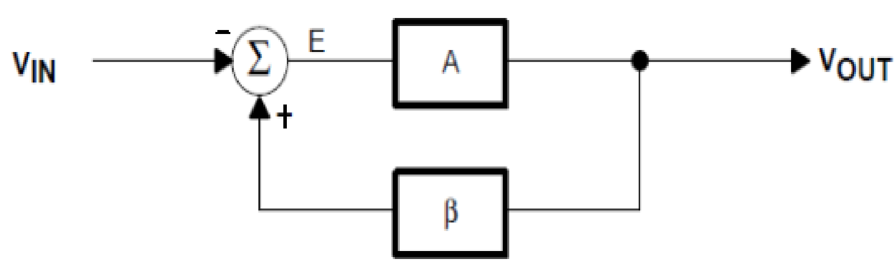

Fig.5.1: Basic structure of oscillator

Barkhausen Criterion

Barkhausen criterion is that in order to produce continuous undamped oscillations at the output of an amplifier, the positive feedback should be such that

$$m_v A_v = 1$$

Once this condition is set in the positive feedback amplifier, continuous undamped oscillations can be obtained at the output immediately after connecting the necessary power supplied.

Mathematical explanation: The voltage gain of a positive feedback amplifier is given by;

$$A_v(f) = = \frac{Av}{1 - mvAv}$$

If $m_v A_v = 1$ then $Av(f) = \rightarrow$ *infinite*

$A_v(f)$ = voltage gain of amplifier with feedback

Where A_v = voltage gain of amplifier without feedback

β *(or)* m_v = feedback fraction

This condition is called the Barkhausen criterion. We know that we cannot achieve infinite gain in an amplifier. So what does this result infer in physical terms? It means that a vanishing small input voltage would give rise to finite (*i.e.*, a definite amount of) output voltage even when the input signal is zero. Thus once the circuit receives the input trigger, it would become an oscillator, generating oscillations with no external signal source.

The loop gain is given by

$$L(S) = A(S).\beta(S) \tag{5.1}$$

Characteristics equation gives rise to

$$1 - L(S) = 0 \tag{5.2}$$

We know that S=jw_0 where w_0 is the frequency of oscillation

An alternative approach to the study of oscillator circuits consists of examining the circuit poles, which are the roots of the characteristic equation eq 5.1. For the circuit to produce sustained oscillations at a frequency (w_0) the characteristic equation has to have roots at $s = \pm jw_0$. There are different oscillator structures available in literature based on bipolar transistor and also using Operational amplifier.

Total Harmonic Distortion:

The term harmonics referred to Power quality in ideal world would mean how pure the voltage is, how pure the current waveform is in its sinusoidal form. Power quality is very important to commercial and industrial power system designs. Ideally, the electrical supply should be a perfect sinusoidal waveform without any kind of distortion. If the current or voltage waveforms are distorted from its ideal form it will be termed as harmonic distortion. This harmonic distortion could result because of many reasons. In today's world, prime importance is given by the engineers to derive a method to reduce the harmonic distortion. Harmonic distortion was very less in the past when the designs of power systems were very simple and conservative. But, nowadays with the use of complex designs in the industry harmonic distortion has increased as well.

The harmful and damaging effects of harmonic distortion can be evident in many different ways such as electronics miss-timings, increased heating effect in electrical equipments, capacitor overloads, etc

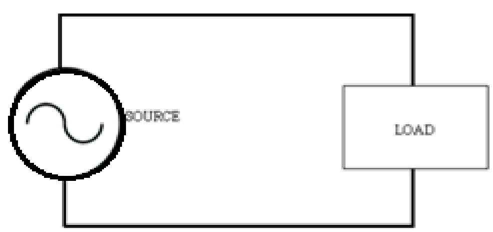

Fig. 5.2 AC source and an electrical load

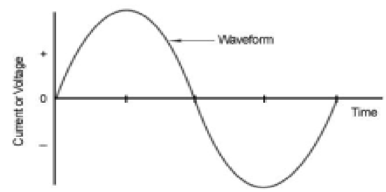

Fig 5.3: Ideal Sine wave

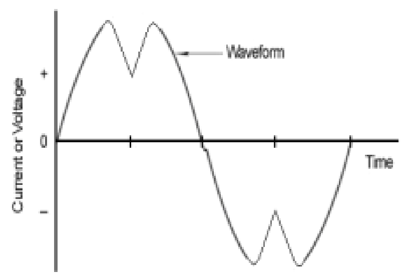

Fig 5.4: Distorted Waveform

As can be observed from the waveform in Figure 5.4, waveform distortions can drastically alter the shape of the sinusoid [20]. However, no matter the level of complexity of the fundamental wave, it is actually just a composite of multiple waveforms called harmonics. Harmonics have frequencies that are integer multiples of the waveform's fundamental frequency. For example, given a 60Hz fundamental waveform, the 2nd, 3rd, 4th and 5th harmonic components will be at 120Hz, 180Hz, 240Hz and 300Hz respectively. Thus, harmonic distortion is the degree to which a waveform deviates from its pure sinusoidal values as a result of the summation of all these harmonic elements. The ideal sine wave has zero harmonic components. In that case, there is nothing to distort this perfect wave.

Total harmonic distortion, or THD, is the summation of all harmonic components of the voltage or current waveform compared against the fundamental component of the voltage or current wave:

$$\text{THD} = \frac{\sqrt{{V_2}^2 + {V_3}^2 + \cdots + {V_n}^2}}{V_1} * 100 \tag{5.3}$$

The formula above shows the calculation for THD on a voltage signal. The end result is a percentage comparing the harmonic components to the fundamental component of a signal. The higher the percentage, the more distortion that is present on the mains signal

5.2 Circuit Description

CIRCUIT I

The first design topology is shown in Fig.5.5. It makes use of two lossy integrator, connected in non inverting mode and inverting mode respectively using single OTA, in the feedback forming a closed loop resulting in loop gain of the system as A(s)β(s) where A(s) is forward path gain and β(s) is feedback gain involving OTA.

The criterion for oscillations [1] to occur, is given by

$$1- A(s)\ \beta(s) =0 \tag{5.4}$$

If we choose R1=R2 and C1=C2, then Routine analysis of the circuit of Fig. 5.5 results in following characteristic equation as

$$S^2C^2 + 2SCG + G^2 + {g_m}^2 = 0 \tag{5.5}$$

From (5.5) it is clear that the circuit Fig.5.5 that it will always produce oscillations, the FO of oscillations can be adjusted by proper selection of resister and capacitor values. R should be high as possible.

$$\text{FO:} \qquad f = \frac{1}{2\pi}\frac{g_m}{C} \tag{5.6}$$

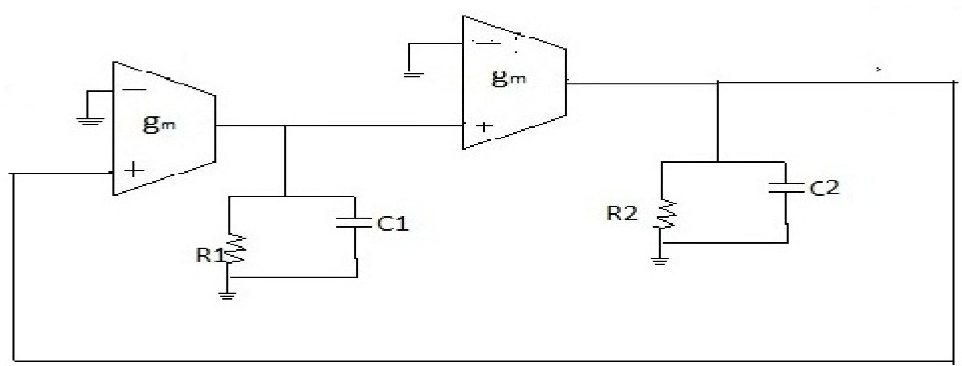

Figure 5.5: Circuit I

Since the resistors are bulky and take larger area, it needs to be replaced by the impedance inverter implemented by OTA and this resistor can be variable by changing g_m .

OTA variable resistors with capacitors can compose OTA-C filters, where the variable resistors can be used to tune the filter frequencies.

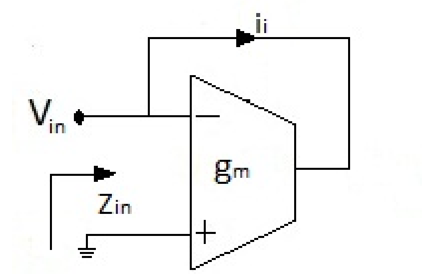

Figure 5.6: Positive Resistor Implementation from OTA

$$Z_i = V_i / I_i = 1/g_m \tag{5.7}$$

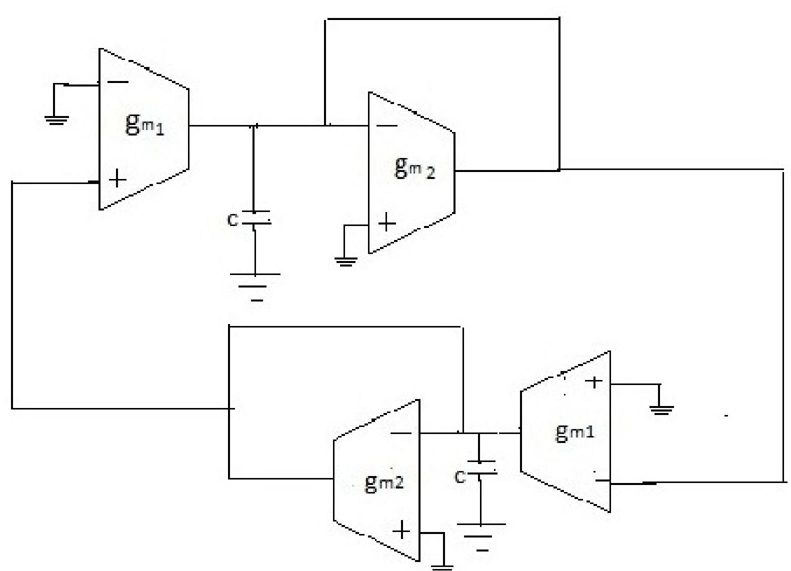

Figure 5.7: OTA implementation of Circuit I

New characteristic equation will be :

$$S^2C^2 + 2SCg_{m2} + {g_{m2}}^2 + {g_{m1}}^2 = 0 \tag{5.8}$$

$$f = \frac{1}{2\pi}\frac{g_{m1}}{C} \tag{5.9}$$

CIRCUIT II

The second design topology is shown in Fig.5.8. It makes use of three lossy integrator, out of which first two are connected in non inverting mode and the third one is connected in inverting mode respectively using single OTA, in the feedback forming a closed loop resulting in loop gain of the system as A(s) β(s) where A(s) is forward path gain and β(s) is feedback gain involving OTA.

The criterion for oscillations [1] to occur, is given by

$$1- A(s)\,\beta(s) =0 \tag{5.10}$$

If we choose R1=R2=R3=R and C1=C2=C3=C, then Routine analysis of the circuit of Fig. 5.8 results in following characteristic equation as

$$S^3C^3 + 3S^2C^2G + 3G^2SC + G^3 + {g_m}^3 = 0 \tag{5.11}$$

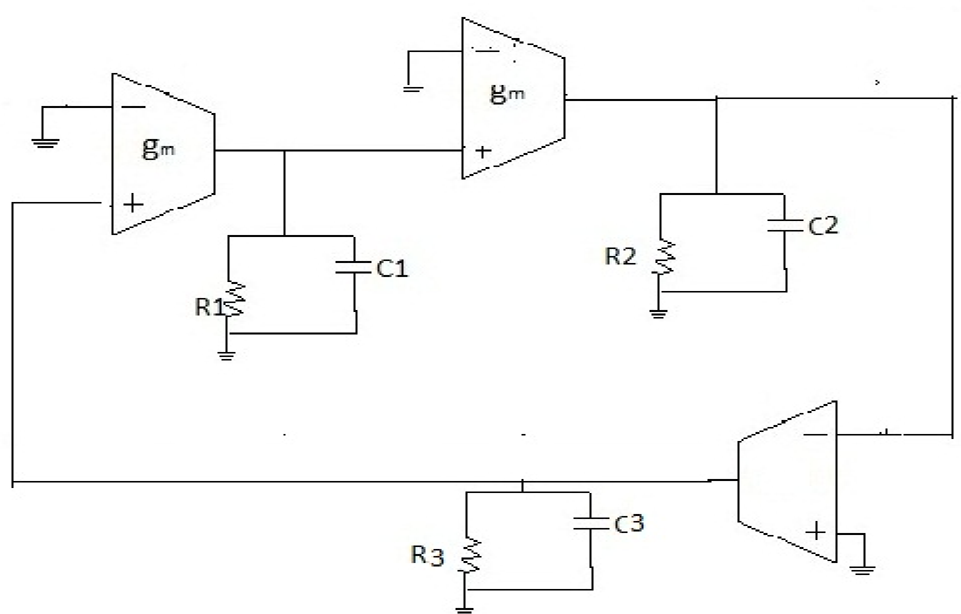

Figure 5.8: Circuit II

From this characteristic equation the frequency of oscillation (FO) and the condition of oscillation (CO) can be found to be

$$\text{FO:} \quad f = \frac{1}{2\pi}\frac{\sqrt{3}.G}{C} \tag{5.12}$$

$$\text{CO:} \quad g_m = 2G \tag{5.13}$$

OTA implementation of Resistor is shown in figure 5.8

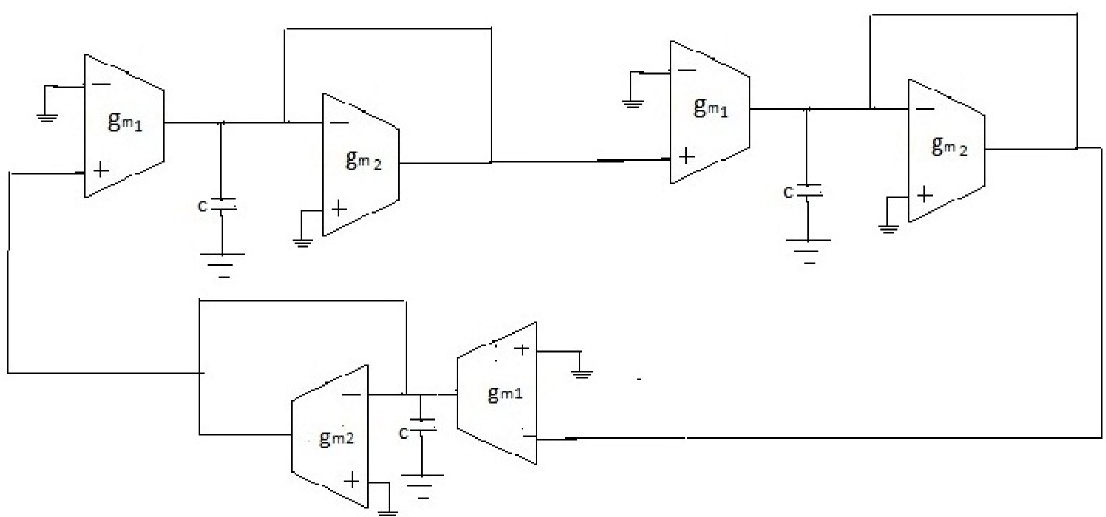

Figure 5.9: OTA implementation of Circuit II

New characteristic equation will be:

$$S^3C^3 + 3S^2C^2g_{m2} + 3g_{m2}{}^2SC + g_{m1}{}^3 + g_{m2}{}^3 = 0 \tag{5.14}$$

From this characteristic equation the frequency of oscillation (FO) and the condition of oscillation (CO) can be found to be

$$\text{FO:} \quad f = \frac{1}{2\pi}\frac{\sqrt{3}.g_{m2}}{C} \tag{5.15}$$

$$\text{CO:} \quad g_{m1} = 2g_{m2} \tag{5.16}$$

CHAPTER 6

Simulation Results

6.1 Introduction

This chapter is organized as follows: The simulations results of second and third order Quadrature oscillator are discussed in section 6.2. Section 6.3 presents the simulation results of different biquad filters.

6.2 Simulation Results of Quadrature Oscillators

The sinusoidal undamped oscillators are verified through simulations using the CMOS implementation of the OTA [16] which is shown in Fig.6.1.The PMOS and NMOS transistors have been simulated by respectively using the parameters of a 0.25µm TSMC CMOS technology [16].

The transistor aspect ratios of PMOS and NMOS transistor are indicated in Table 6.1. Supply voltages taken are ±1.5V. Value of I_{Bias} taken is 108u. For both the circuits the resistor and capacitor values are taken as R_1=R_2=90K, C_1=C_2=10p and for circuit II R_1=R_2=2K, C_1=C_2=100p. Topology 1 is designed for an FO of 17.8 MHz and the simulated value was observed to be 18 MHz and for Topology II is designed for an FO of 160 KHz and the simulated value was observed to be 144 KHz. The simulated transient output and corresponding frequency spectrum for Circuit I is shown in Fig.6.2 and Fig.6.3 respectively and those for Circuit II are depicted in Fig.6.4 and Fig6.5. The percentage total harmonic distortion (THD) Table 6.4 is 1.47% for Circuit I and that for Circuit II Table 6.5 is observed to be 6.55%.

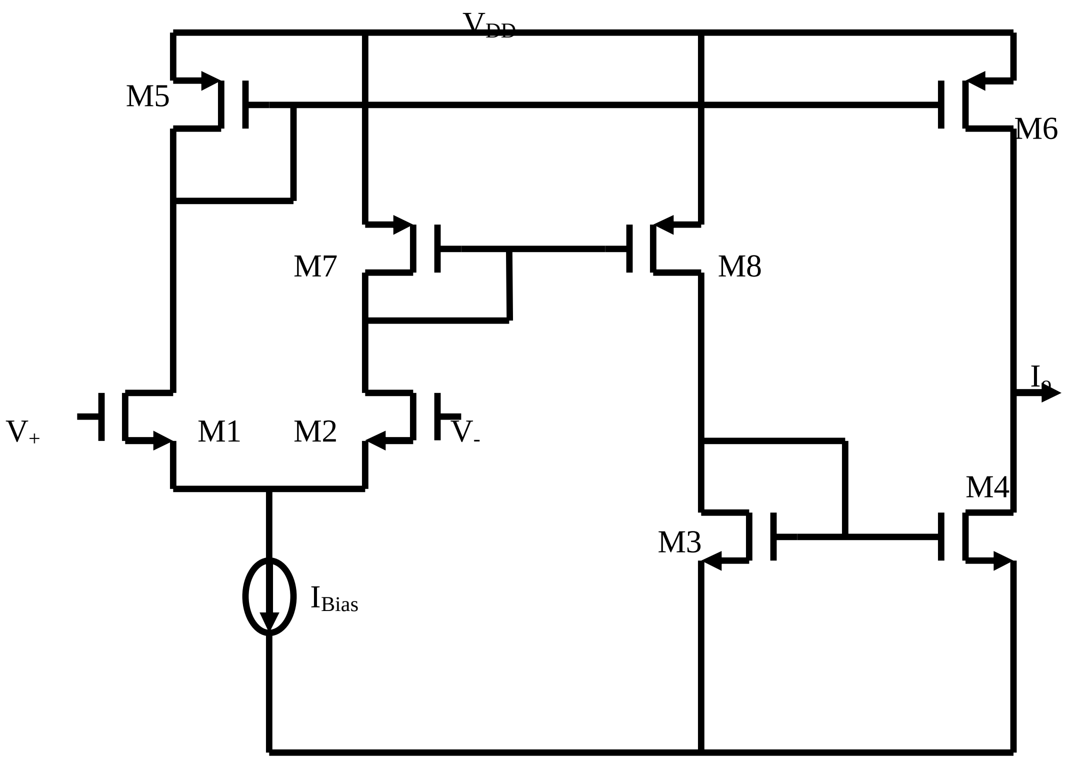

Figure 6.1:CMOS Implementation of OTA

Transistor	W(**µm)**	L(**µm)**
$M_1 - M_2$	16	0.25
$M_3 - M_4$	3	0.25
$M_5 - M_6 - M_7$	5	0.25
M_8	4.45	0.25

Table 6.1: Dimension of the Transistors

Circuit	Active Element	Number of C Used	Number of R used	C.O	F.O
Figure 5.1	2 OTA	2	2	-	$f = \frac{1}{2\pi}\frac{g_m}{C}$
Figure 5.4	3 OTA	3	3	$g_m = 2G$	$f = \frac{1}{2\pi}\frac{\sqrt{3}.G}{C}$

Table 6.2: Summary of the circuits

Circuit	Active Element	C.O	F.O
Figure 5.3	4	-	$f = \frac{1}{2\pi}\frac{g_{m1}}{C}$
Figure 5.7	6	$g_{m1} = 2g_{m2}$	$f = \frac{1}{2\pi}\frac{\sqrt{3}.g_{m2}}{C}$

Table 6.3: Summary of the circuits after use of OTA Impedance Inverter

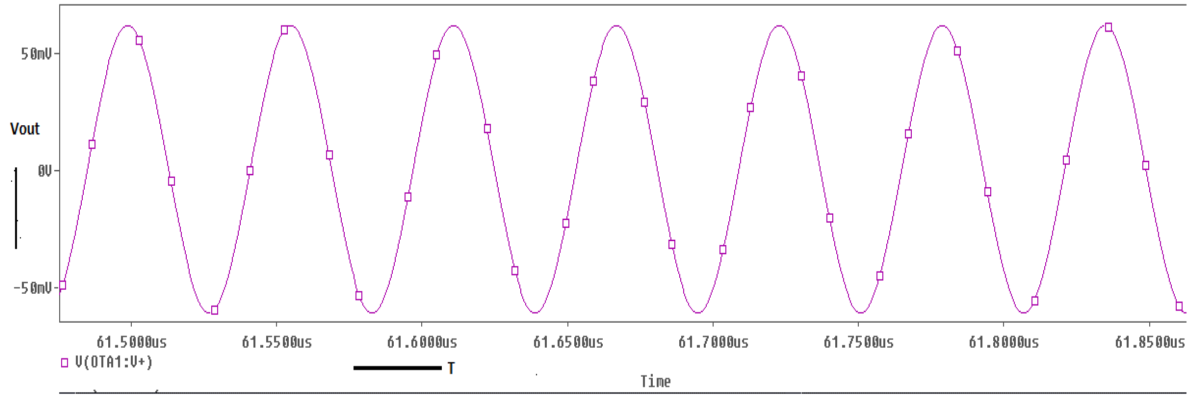

Figure 6.2: Transient Output of Circuit I

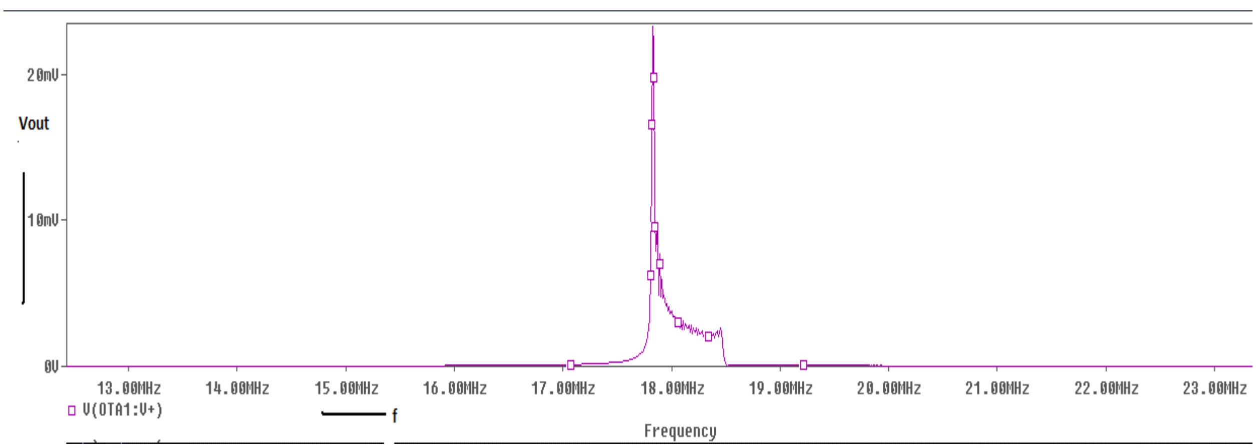

Figure 6.3: Frequency spectrum of output of Circuit I

HARMONIC NO	FREQUENCY (HZ)	FOURIER COMPONENT	NORMALIZED COMPONENT	PHASE (DEG)	NORMALIZED PHASE(DEG)
1	1.780E+07	6.212E-02	1.000E+00	-6.810E+01	0.000E+00
2	3.560E+07	3.104E-04	4.997E-03	1.313E+02	1.298E+01
3	5.340E+07	1.854E-04	2.984E-03	8.201E+00	2.505E+02
4	7.120E+07	1.180E-04	1.899E-03	1.011E+02	1.327E+02
5	8.900E+07	9.791E-05	1.576E-03	8.698E+01	2.449E+02
6	1.068E+08	9.711E-05	1.563E-03	7.839E+01	4.761E+02
7	1.246E+08	8.061E-05	1.298E-03	6.313E+01	2.315E+02
8	1.424E+08	9.702E-05	1.562E-03	1.093E+01	1.798E+02
9	1.602E+08	1.648E-04	2.654E-03	4.257E+00	4.198E+02
10	1.780E+08	3.135E-04	5.047E-03	4.576E+00	1.426E+02
TOTAL HARMONIC DISTORTION = 8.890463E-01 PERCENT					

TABLE 6.4: TOTAL HARMONIC DISTORTION FOR FIGURE 5.5

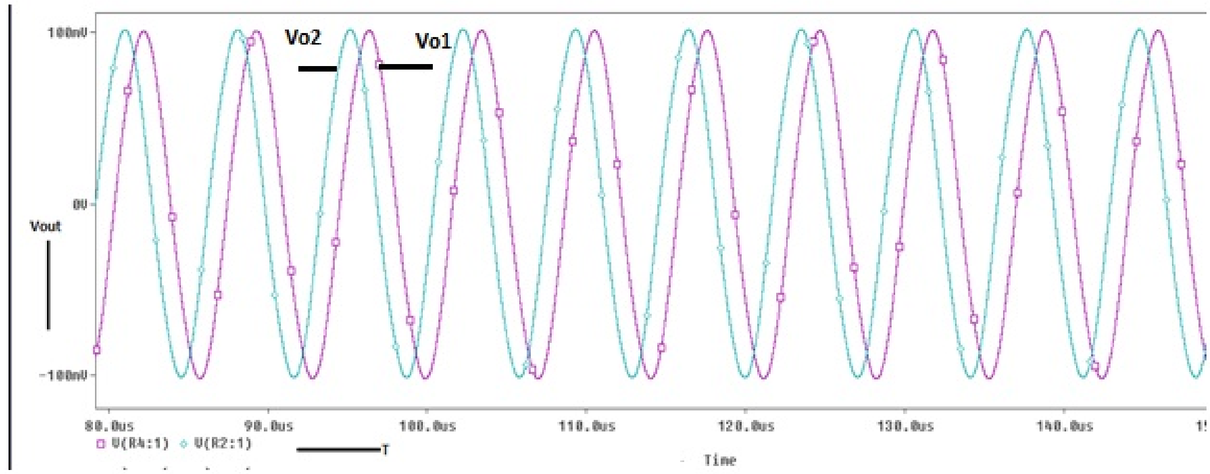

Figure 6.4: Transient Output of Circuit II

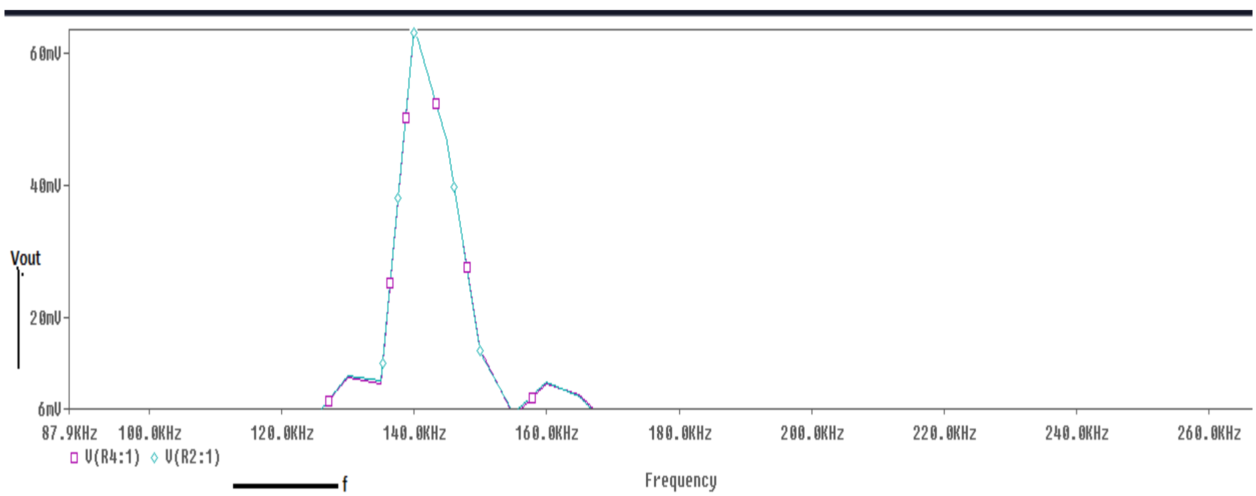

Figure 6.5: Frequency spectrum of output of Circuit II

HARMONIC NO	FREQUENCY (HZ)	FOURIER COMPONENT	NORMALIZED COMPONENT	PHASE (DEG)	NORMALIZED PHASE(DEG)
1	1.600E+05	9.445E-02	1.000E+00	9.997E+01	0.000E+00
2	3.200E+05	4.466E-03	1.616E-02	1.149E+02	-2.211E+02
3	4.800E+05	3.722E-03	1.347E-02	8.653E+01	-3.943E+02
4	6.400E+05	3.330E-03	1.205E-02	1.047E+02	-5.605E+02
5	8.000E+05	2.856E-03	1.033E-02	1.219E+02	-7.249E+02
6	9.600E+05	2.266E-03	8.199E-03	1.242E+02	-8.904E+02
7	1.120E+06	1.653E-03	5.980E-03	1.303E+02	-1.420E+02
8	1.280E+06	1.180E-03	4.267E-03	1.319E+02	-1.600E+02
9	1.440E+06	9.843E-04	3.561E-03	1.385E+02	-1.427E+02
10	1.600E+06	9.813E-04	3.550E-03	1.445E+02	-1.609E+02
TOTAL HARMONIC DISTORTION = 6.558155E+00 PERCENT					

TABLE 6.5: TOTAL HARMONIC DISTORTION FOR FIGURE 5.8

6.3 Simulation Results of Biquad Filters

- **Voltage Mode Universal Filter using OTA**

To show the performance of the circuit shown in chapter 4, fig.4.5, PSPICE simulators are used. The PMOS and NMOS transistors have been simulated by respectively using the parameters of a 0.25μm TSMC CMOS technology. The power supplies are selected as

VDD = -VSS = 1.25 V. As an example design, the parameters C1 = C2 = 1nF, I_{ref1} = I_{ref2} = 108μA (g_{m1}=g_{m2} = 1.12mS) and I_{ref3} = 50μA (g_{m3} = 181.9μS) are given.

Since $g_{m1}=g_{m2}$, then $\omega_o = \frac{g_m}{C}$ and $Q = \frac{g_{m1}}{g_{m2}}$

This gives us Q to be 6.15

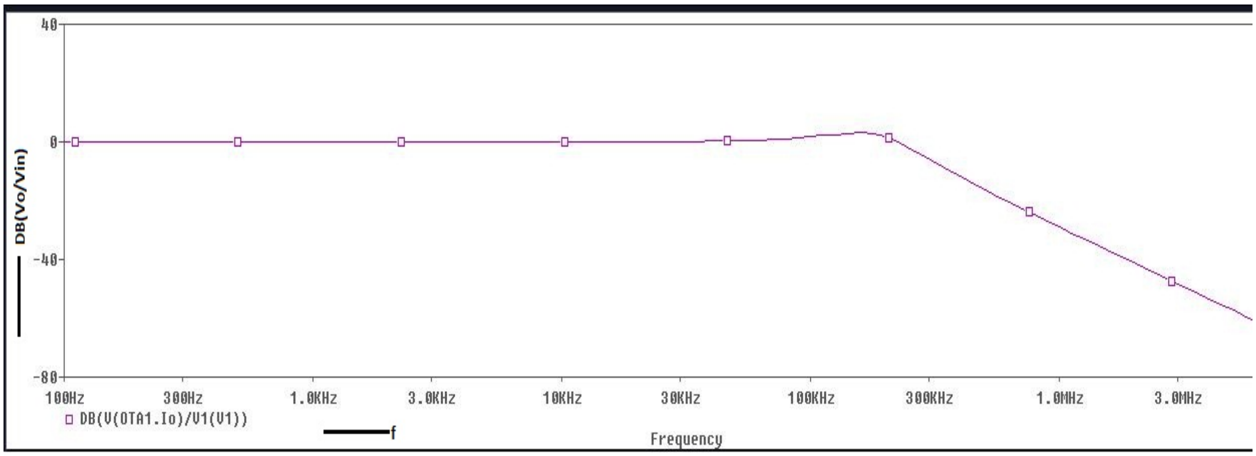

Figure 6.6: Low Pass Response

Cut-off frequency obtained here during simulations: f_{3dB}=260.2 kHz

Theoretically Obtained frequency = 180 kHz

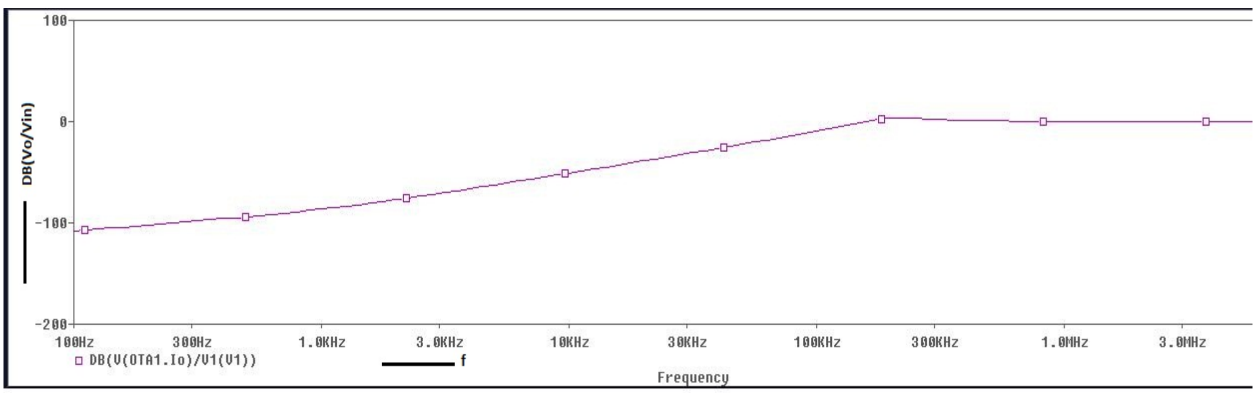

Figure 6.7: High Pass Response

Cut-off frequency obtained here during simulations: f_{3dB}=133 kHz

Theoretically Obtained frequency = 180 kHz

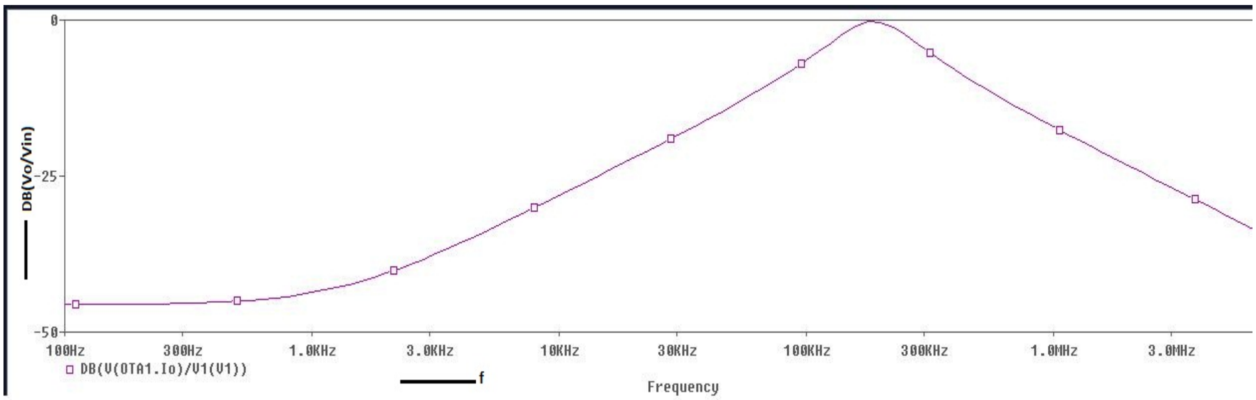

Figure 6.8: Band Pass Response

Centre cut-off frequency obtained here during simulations: f_{3dB}=186.2 kHz

Theoretically Obtained frequency = 180 kHz

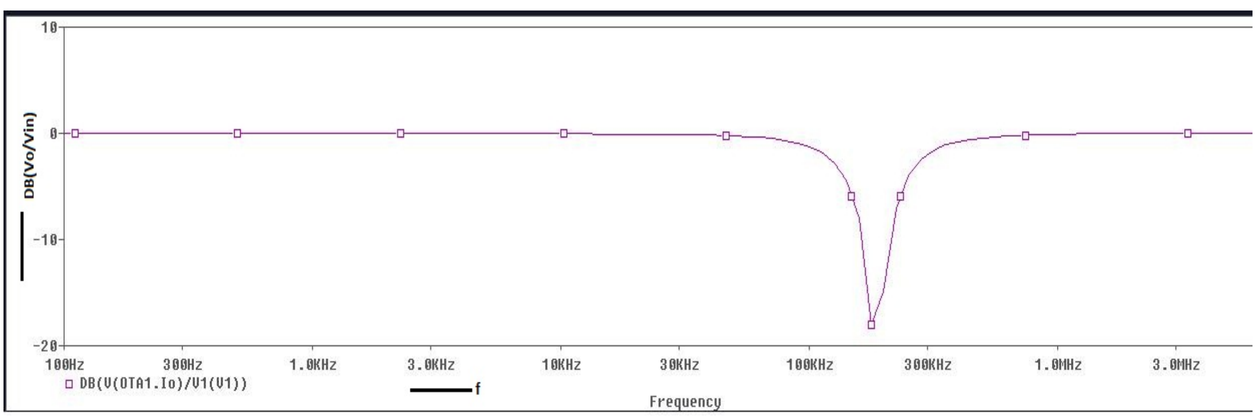

Figure 6.9: Band Stop Response

Centre cut-off frequency obtained here during simulations: f_{3dB}=179.2 kHz

Theoretically Obtained frequency = 180 kHz

- **Voltage Mode SIMO Biquad Filter**

To show the performance of the circuit shown in chapter 4, figure 4.6, PSPICE simulators are used. The PMOS and NMOS transistors have been simulated by respectively using the parameters of a 0.25µm TSMC CMOS technology. The power supplies are selected as

VDD = -VSS = 1.25 V. As an example design, the parameters C1 = C2 = 100pF, I_{ref1} = I_{ref2} = 20µA (g_{m1}=g_{m2} = 266.8µS) and I_{ref3} = 10µA (g_{m3} = 137µS) are given.

Since g_{m1}=g_{m2} then $\omega_o = \frac{g_m}{C}$

And as Q = $\sqrt{\frac{g_{m2}C_1}{g_{m1}C2}}$, this gives us Q=1 as g_{m1}=g_{m2} and C1=C2

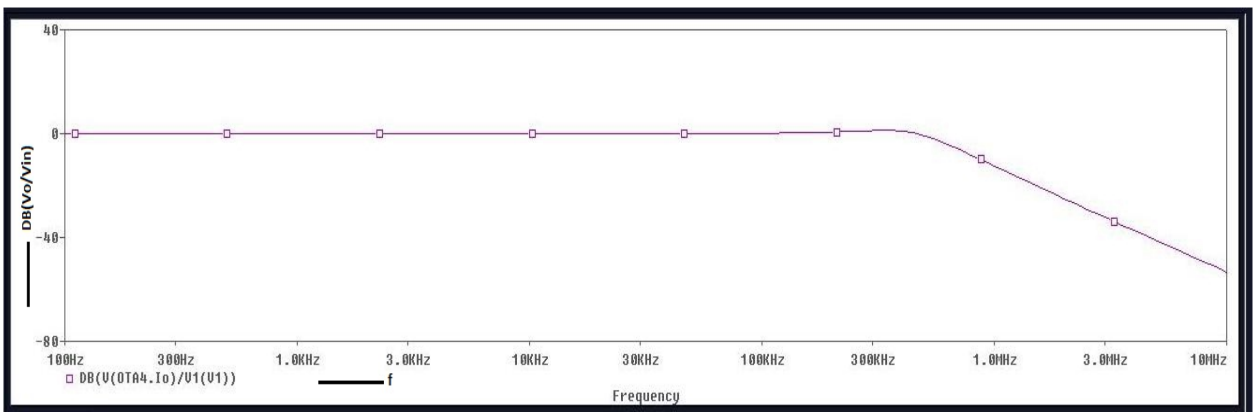

Figure 6.10: Low Pass Response for fig. 4.6

Cut-off frequency obtained here during simulations for low pass: f_{3dB}=590 kHz

Theoretically Obtained frequency = 430 kHz

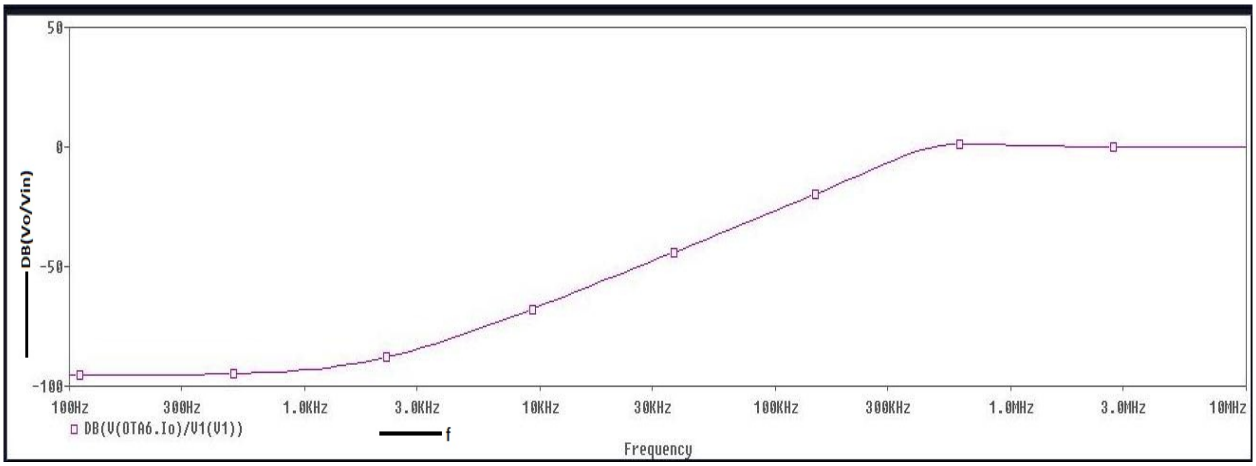

Figure 6.11: High Pass Response for fig. 4.6

Cut-off frequency obtained here during simulations: f_{3dB}=365 kHz

Theoretically Obtained frequency = 430 kHz

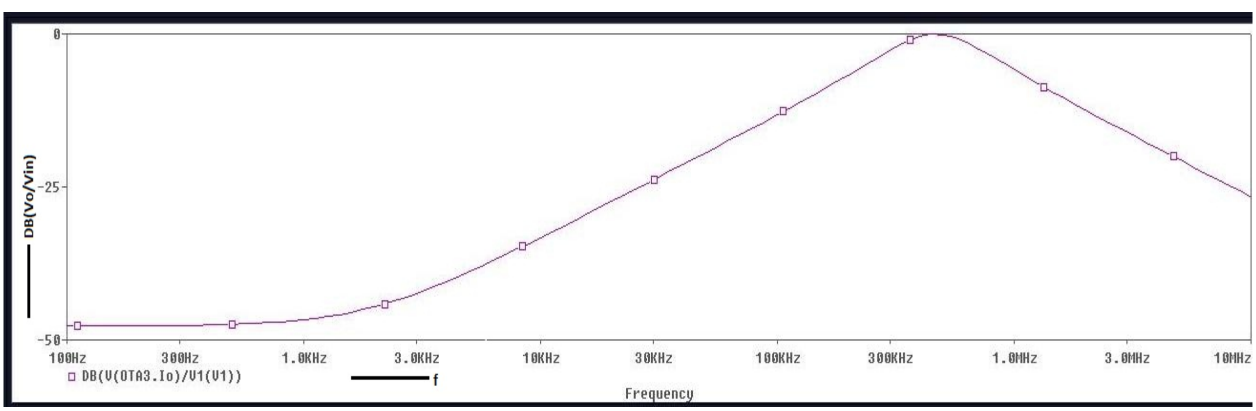

Figure 6.12: Band Pass Response for fig. 4.6

Cut-off frequency obtained here during simulations: f_{3dB}= 467 kHz

Theoretically Obtained frequency = 430 kHz

CHAPTER 7

CONCLUSION & FUTURE SCOPE

7.1 *THE OUTCOME*

The working of third order Quadrature Oscillator and its analysis is performed and the Resistors used in the topologies are replaced from the variable resistor (An Application of the OTA). All the simulations have been done using PSPICE. Different filters topologies also studied and simulated in thesis, several single input multiple outputs and multiple inputs single output filters have been studied and few of them have been implemented. Both single input multiple out and multiple input single output circuits have been simulated and output has been observed. Theoretical and simulation results have been verified.

We also discussed the fundamentals of the Operational Transconductance Amplifier (OTA) and its properties. Several different models of OTA and their advantages and disadvantages also have been discussed.

It has following advantages:

- Provides an electronic tunability
- Because of the linear proportionality of the transconductance it can be tuned linearly using a control voltage.
- It has very high input impedance so it’s easy to cascade other stages.

In this report efforts are made to study the scope of OTA as an active building block in analog circuits.

7.2 *SCOPE FOR FUTURE WORK*

In this thesis efforts are made to study the scope of OTA as an active building block in analog circuits. Various CMOS realization of OTA present in the literature are studied and these circuits are used to realize various signal processing and generating circuits having applications in communication. All the circuits were simulated using PSpice program and 0.25um process parameters. Simulation results show that the various characteristics are in good agreement with the theory. Slight variations in the results arise due to the non ideal behavior of the OTA used.

In this thesis some of the Quadrature oscillators using OTA are discussed while non linear applications are not discussed and the harmonic analysis is not carried out for the oscillators discussed in chapter 5 .The phase analysis of the circuits is not included in this thesis.

Due to lack of time, non ideal analysis of above structures could not be performed. OTA has been used to implement filters and oscillators in the presented work, but the area of its utility is very broad and hence a lot of possible work is yet to be done. OTA can be combined with other concepts like local feedback, bootstrapping etc., to further improve the performance.

- Simulation results show that the current conveyors have current transfer bandwidth greater than 1.8 MHz this bandwidth can be further increased by using the bandwidth enhancement techniques.
- The present work is designed at a supply voltage of ±1.25V in order to have satisfactory dc voltage following action. If the simple current mirrors are replaced by low voltage FGMOS current mirrors then the present work can be designed at low supply voltages with low biasing currents and static power dissipation is also low in this case.
- The output offset current errors are strongly dependent on the matching accuracy of the current mirrors. This offset can be minimized by matching the current mirrors by careful layout techniques like common centroid and unit cell layout techniques.

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APPENDIX

The PMOS and NMOS transistors have been simulated by respectively using the parameters of a 0.25µm TSMC CMOS technology.

.MODEL MbreakN NMOS (LEVEL = 3

+ TOX = 5.7E-9 NSUB = 1E17 GAMMA = 0.4317311

+ PHI = 0.7 VTO = 0.4238252 DELTA = 0

+ UO = 425.6466519 ETA = 0 THETA = 0.1754054

+ KP= 2.501048E-4 VMAX= 8.287851E4 KAPPA= 0.1686779

+ RSH = 4.062439E-3 NFS = 1E12 TPG = 1

+ XJ= 3E-7 LD = 3.162278E-11 WD = 1.232881E-8

+ CGDO = 6.2E-10 CGSO = 6.2E-10 CGBO = 1E-10

+ CJ = 1.81211E-3 PB = 0.5 MJ = 0.3282553

+ CJSW = 5.341337E-10 MJSW = 0.5)

.MODEL MbreakP PMOS (LEVEL = 3

+ TOX=5.7E-9 NSUB = 1E17 GAMMA = 0.6348369

+ PHI = 0.7 VTO = -0.5536085 DELTA = 0

+ UO = 250 ETA = 0 THETA = 0.1573195

+ KP=5.194153E-5 VMAX=2.295325E5 KAPPA = 0.7448494

+ RSH = 30.0776952 NFS = 1E12 TPG = -1

+ XJ= 2E-7 LD = 9.968346E-13 WD = 5.475113E-9

+ CGDO= 6.66E-10 CGSO = 6.66E-10 CGBO = 1E-10

+ CJ 1.893569E-3 PB = 0.9906013 MJ = 0.4664287

+ CJSW = 3.625544E-10 MJSW = 0.5)