The Four-Terminal Floating Nullor (FTFN) and its Applications

by

Hussain Abdullah Al-Zaher

A Thesis Presented to the

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

In Partial Fulfillment of the Requirements for the Degree of

MASTER OF SCIENCE

In

ELECTRICAL ENGINEERING

May, 1997
INFORMATION TO USERS

This manuscript has been reproduced from the microfilm master. UMI films the text directly from the original or copy submitted. Thus, some thesis and dissertation copies are in typewriter face, while others may be from any type of computer printer.

The quality of this reproduction is dependent upon the quality of the copy submitted. Broken or indistinct print, colored or poor quality illustrations and photographs, print bleedthrough, substandard margins, and improper alignment can adversely affect reproduction.

In the unlikely event that the author did not send UMI a complete manuscript and there are missing pages, these will be noted. Also, if unauthorized copyright material had to be removed, a note will indicate the deletion.

Oversize materials (e.g., maps, drawings, charts) are reproduced by sectioning the original, beginning at the upper left-hand corner and continuing from left to right in equal sections with small overlaps. Each original is also photographed in one exposure and is included in reduced form at the back of the book.

Photographs included in the original manuscript have been reproduced xerographically in this copy. Higher quality 6” x 9” black and white photographic prints are available for any photographs or illustrations appearing in this copy for an additional charge. Contact UMI directly to order.

UMI
A Bell & Howell Information Company
300 North Zeeb Road, Ann Arbor MI 48106-1346 USA
313/761-4700 800/521-0600
THE FOUR-TERMINAL FLOATING NULLOR (FTFN)
AND ITS APPLICATIONS

BY
HUSSAIN ABDULLAH AL-ZAHER

A Thesis Presented to the
FACULTY OF THE COLLEGE OF GRADUATE STUDIES
KING FAHD UNIVERSITY OF PETROLEUM & MINERALS
DHAHRAIN, SAUDI ARABIA

In Partial Fulfillment of the
Requirements for the Degree of

MASTER OF SCIENCE
In
ELECTRICAL ENGINEERING

MAY, 1997
KING FAHD UNIVERSITY OF PETROLEUM & MINERALS
DHAHRAN, SAUDI ARABIA

This theses is written by

Hussain Abdullah Al-Zaher

under the direction of his Thesis Advisor, and approved by his Thesis committee, has been
presented to and accepted by the Dean of the College of Graduate Studies, in partial
fulfillment of the requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

Thesis Committee

Dr. M. T Abuelma'atti (Chairman)

Dr. A.R. K. Al-Ali (Member)

Dr. C. B. Yahya (Member)

Dr. A.K. Hamid (Member)

Dr. Samir Al-Baiyat
Chairman, Electrical Engineering Department

Dr. Abdullah M. Al-Shehri
Dean, College of Graduate Studies
Date: 17-5-97
To
My Father,
Mother,
Wife,
Brothers
&
Teachers

For their patience, moral support and encouragement
Acknowledgment

First and foremost all praise to Almighty Allah who gave me the ability, the power and the patience to complete this work successfully.

I remain grateful to King Fahd University of Petroleum and Minerals and particularly the Electrical Engineering Department for supporting this work.

I am extremely grateful to my thesis advisor Dr. M. T. Abuelma’atti for his continuous support and guidance throughout the work.

Also, I thank my committee members Dr. Al-Ali, A., Dr. Yahyai and Dr. Hamed for their advise and cooperation.
TABLE OF CONTENTS

Acknowledgment i
Table of Contents ii
List of Figures vi
List of Tables ix
Abstract (English) x
Abstract (Arabic) xi

CHAPTER 1

INTRODUCTION

1.1 The Four-Terminal Floating Nullor (FTFN) 1
1.2 The FTFN Realizations 4
1.3 Literature Review of FTFN Applications 14
1.4 The FTFN Versatility and Flexibility 21
1.5 Approaching the Universal Element 25
1.6 Problem Definition 28
1.7 Thesis Organizations 29
CHAPTER 2

APPLICATIONS OF THE FTFN:

I. ACTIVE FILTERS

2.1 Introduction 30

2.2 Evaluating an Active Filter 32

2.3 Universal Second-Order Filters 34

2.4 Universal Multiple Input Single Output (MISO)
   Current Mode Filters Using FTFN 35
      2.4.1 Universal (MISO) Current Mode Filter (I) 35
      2.4.1 Universal (MISO) Current Mode Filter (II) 43

2.5 Comparison with CCII+- Counterparts 51
   2.5.1 New CCII+ based (MISO) Current Mode Filter 53

2.6 Comparison with OTA-Op Amp Counterparts 59

2.7 Universal Single Input Multiple Output (SIMO)
   Current Mode Filters 67
      2.7.1 Universal (SIMO) Current Mode Filters (I) 68
      2.7.2 Universal (SIMO) Current Mode Filter (II) 73

2.8 Comparison with CCII+ Counterparts 79

2.9 Universal (MISO) Voltage Mode Filters Using FTFN 87

2.10 Comparison with CCII+ Counterparts 96

2.11 Comparison with CFA Counterparts 97

2.12 Summary 104
CHAPTER 3

APPLICATIONS OF THE FTFN:

II. SINUSOIDAL OSCILATORS

3.1 Introduction 107

3.2 Assessing a Sinusoidal Oscillator 109

3.3 Current-Mode Sinusoidal Oscillators Using Single FTFN 110

3.3.1 Experimental Results 118

3.3.2 Summary 124

3.4 SINUSODAL OSCILLATORS USING TWO FTFNs 127

3.4.1 Sinusoidal oscillators derived from the proposed oscillator based on two FTFNs 133

CHAPTER 4

NONIDEAL ANALYSIS OF THE FTFN FRONT-END

4.1 Introduction 137

4.2 Nonlinear Analysis of the FTFN Based on the (CCII+) 138

4.3 Harmonic and Intermodulation Products 142

4.4 Simulation Results 146

4.5 Conclusion 147
CHAPTER 5

CONCLUSION AND FUTURE WORK

5.1 Introduction 148
5.2 Conclusions 149
5.3 Future Work 152

References 153

Vita 162
LIST OF FIGURES

1.1 The Four terminal Floating Nullor FTFN 3
1.2 The Four Types of Amplifiers Based on FTFN 3
1.3 Wilson’s Current Mirror and Its Symbol 5
1.4 The FTFN or OFA Based on Current Sensing Technique 6
1.5 FTFN or OFA Based on Current Source Power Supply 7
1.6 Four CCII- based FTFN Realizations 9
1.7 CCII- Based on CCII+ 12
1.8 FTFN based on CCII− 13
1.9 Fundamental Applications of the FTFN 14
1.10 Voltage to Current Conveyor Based on Op Amp 15
1.11 Integrator and Differentiator using FTFN 16
1.12 Multiple Outputs OTA Based on FTFNs 17
1.13 Obtaining the Basic Filters (LP, BP and HP) From Each Others 24
1.14 CCII+ and CCII- Obtained From the Universal Element 26
1.15 CCI- and CCI+ From UE 26
1.16 First -order Filters Using the UE 27
1.17 The Universal Element Equivalent (UE) 28
2.1 Proposed (SIMO) Current Mode Universal Filter (I) 38
2.2 Simulation Results of the Circuit of Fig.2.1 40
3.4 Experimental and Simulation Results Obtained from the Circuit of Fig. 3.2 120
3.5 Svoboda Model of Non-Ideal Current Conveyor CCII+ 122
3.6 Experimental Results of the Circuit of Fig. 3.3 123
3.7 A typical Current Quadrature signals obtained from the Circuit of Fig. 3.2 126
3.8 Variable Phase Proposed Oscillator Using Two FTFN 127
3.9 Adding Resistors to the Circuit of Fig. 3.8 for Controlling the Phase Shift 129
3.10 Simulation results of the Oscillator of Fig. 3.8 131
3.11 Simulation results of Phase Shift of the Oscillator of Fig. 3.9 131
3.12 Simulation results of Phase Shift of the Oscillator of Fig. 3.9 132
3.13 Quadrature Current Signals Obtained from the Oscillator of Fig. 3.9 132
3.14 Proposed Oscillators Derived From the Circuit of Fig. 3.8 134-135
3.15 Adding a Resistor for Controlling the Amplitude of the Circuit of Fig. 3.14(g) 136
4.1 Mixed Translinear Loops 139
4.2 Calculated and Simulated THD of the Translinear loops of Fig. 4.1 146
LIST OF TABLES

2.1 Comparison Between (MISO) Current Mode Filters Based on FTFN and CCII+ 58
2.2 Comparison Between (MISO) Current Mode Filters Based on FTFN and OTA-Op Amp 66
2.3 Comparison Between (SIMO) Current Mode Filters Based on FTFN and CCII+ 86
2.4 Comparison Between (MISO) Voltage Mode Filters Based on FTFN and CCII+ 96
2.5 Comparison Between (SIMO) Voltage Mode Filters Based on FTFN and CFA 103
3.1 Frequency and condition of oscillation of the oscillator circuits of Case I 112
3.2 Frequency and condition of oscillation of the oscillator circuits of Case II-1 113
3.3 Frequency and condition of oscillation of the oscillator circuits of Case II-2 114
3.4 Frequency and condition of oscillation of the oscillator circuits of Case II-3 114
3.5 Frequency and condition of oscillation of the oscillator circuits of Case II-4 115
3.6 Frequency and condition of oscillation of the oscillator circuits of Case III 116
ABSTRACT

Name: Hussain Abdullah Al-Zaher

Title: The Four-Terminal Floating Nullor (FTFN) and Its Applications

Major Field: Electrical Engineering

Date of Degree: May, 1997

The Thesis discusses the Four-Terminal Floating Nullor (FTFN) and its applications. Different possible realizations of the FTFN are evaluated. Its wide range of applications is investigated. The versatility and the flexibility of the FTFN is demonstrated and exemplified. Universal filters and sinusoidal oscillators based on the FTFN are designed. The advantages and the disadvantages of each circuit are indicated. The proposed filters are compared with new CCII+, Op Amp-OTA, CFA counterparts and the proposed oscillator circuits are compared with the previously published circuits. Experimental and simulation results confirming the theoretical designs are included.

Master of Science Degree

King Fahd University of Petroleum and Minerals

Dhahran, Saudi Arabia

May, 1995
خلاصة الرسالة

الاسم: حسن عبد الله احمد الزاهر

عنوان الرسالة: النويل ذو الأطراف الأربعة العائمة وتطبيقاته

التخصص: الهندسة كهربائية

تاريخ الشهادة: مايو 1997م

هذه الرسالة تناقش النويل ذو الأطراف الأربعة العائمة وتطبيقاته. وقد قمت طرق تحقيق النويل المختلفة وتطبيقاته المتعددة. وكذلك عرضت بعض الأمثلة للدلالات على سعة النويل وسماحته. وفي هذا الإطار صممت أعداد كثيرة من المرشحات ومولدات الدفعات وتمكنت من تجربة كل منها. وقررنا استخدام مكونات جديدة من المشاواة باستخدام ناقل التيار (الجذب الثانية الموجه)، المضخم العادي مبع مكيرات التحويل الجهد إلى تيار وكذلك المضخم ذو التغذية التيارية الراجعة. أما دوائر مولدات الدفعات فقد قررنا استخدام مشاواة سبيس تنشرها. وقد حققت من عمل الدوائر المفترضة باختبارها جميعا عدة عمليات تمثلها باستخدام برامج الحاسب المتخصصة.

درجة الماجستير في العلوم
جامعة الملك فهد للبترول والمعادن
الظهران – المملكة العربية السعودية

مايو 1997م
CHAPTER 1

INTRODUCTION

1.1 The Four-Terminal Floating Nullor (FTFN)

The current mode approach of designing analog integrated circuits is becoming increasingly important with growing opportunities. The nearly true complementary transistors are being developed opening the way for designing efficient current mode devices. The current mode processing has potential advantages over the voltage mode such as wider bandwidth, almost independent of the closed loop gain, wider dynamic range with a very high slew rate, lower power consumption and simpler circuitry [1,2].
As a result, several old current mode techniques are being reinstated and a new generation of current mode building blocks and systems are being developed aiming to overcome the conventional Op Amps limitations. Which one of them will replace the Op Amps is not clear at present. Among these, however, the Four-Terminal Floating Nullor (FTFN) seems to be a very promising, and extremely powerful device combining both voltage and current mode capabilities. In fact, most of the devices proposed in both academic and industry can be configured using the FTFN. For example, the minus-type second generation current conveyor CCII- which proved to be very promising device is only a special case of FTFN. This implies that the most versatile and flexible active device has to be a practical implementation of FTFN.

The FTFN, described as the elemental active device or the ideal amplifier, was first implicitly introduced in 1954 by Tellegen [3]. It was demonstrated that any active circuit can be realized using only such element and passive components. This universal active network element was called Nullor in 1964 by Carlin [3]. It comprises an input Nullator (V₁=0, I₁=0) and a Norator at the output port with arbitrary (V₂ & I₂) as illustrated in Fig.1.1. It is known as Nullor shorting a Nullator-Norator pair. Also, since the four terminals of the Nullor are isolated from ground (floating) and to be distinguished from the three terminal Nullor of [4] which is associated with a grounded terminal, this Nullor is commonly referred to as Four-Terminal Floating Nullor (FTFN). According to the definition of FTFN, an ideal FTFN exhibits these terminal characteristics: Iₓ=Iᵧ=0, Vₓ=Vᵧ, and I₂=Iₓ.
Fig. 1.1: The Four terminal Floating Nullor FTFN

One important feature of the FTFN is that it can be configured as any type of the four basic amplifiers: Voltage Amplifier, Current Amplifier, Transresistance Amplifier and Transconductance Amplifier using the minimum number of active and passive elements as shown in Fig 1.2 and was, therefore, called the ideal (universal) amplifier.

Fig. 1.2: The Four Types of Amplifiers Based on FTFN
1.2 The FTFN Realizations

Since its introduction in the sixties, the FTFN has remained a theoretical element used only in circuit theory books. Recently, however, after the improvement of complementary transistors fabrication and the current mode circuits evolution, the FTFN implementation has become feasible. Thus, it is moving from being a theoretical element to the wide range of possible applications.

At present, the FTFN is not yet commercially available. However, several different techniques of realizing the FTFN utilizing the available devices have been suggested. These are possible candidates for the implementation of the FTFN as a discrete element. Moreover, the full integration of the entire circuit of the FTFN provides even higher freedom [5]. For example, Huijsing and Korte described the design of a monolithic FTFN [6]. These different approaches of realizing the FTFN are evaluated in this section. A suitable realization is selected to be used in simulating and experimenting the circuits to be proposed.

Nowadays, the Operational Floating Amplifier (OFA) is the most popular form of realizing the FTFN. This technique extends the performance of the available Op Amps from voltage mode to current mode by providing an additional current-output terminal. This additional floating terminal can be achieved in, mainly, two ways.
The first, is to use current mirrors, such as Wilson improved current mirror, shown in Fig.1.3, to sense the output current of an Op Amp and convey it at an additional current port as illustrated in Fig.1.4. [7]. The idea behind this technique is simple. Since the sum of the currents in the supply leads of the Op Amp is equal to the sum of its input and output currents, current mirrors are incorporated to sense the split currents of the Op Amp supply rails and combine them at single high impedance output. This approach is valid provided that there is no other connections with ground carrying large current rather than supply leads and the load and that the input currents to the Op Amp are almost zero. Fortunately, this condition is satisfied in most Op Amps. An additional current mirror is added to invert the output current direction to exhibit the current polarity required to simulate the FTFN.

Fig.1.3: Wilson Current Mirror and Its Symbol
Fig.1.4: The FTFN or OFA Based on Current Sensing Technique

The second technique of providing the floating output is achieved by feeding DC power to the Op Amp from the balanced current source and sink proposed by Nordholt [8]. As shown in Fig.1.5, two series zener diodes are used to fix the supply voltages. The signal current at the negative output terminal (W) cannot flow through the fixed current source or current sink and hence has to be picked by the Op Amp rails and it will be reflected at the positive output terminal (Z). Similarly, a signal current at the positive terminal (Z) will be reproduced at the negative terminal (W).
According to nullator and norator definition the ideal FTFN exhibits undefined input and output impedances. However, the practical FTFN conciliates the impedance levels of the circuits used to implement it. According to the above discussion, it is clear that the OFA approximates the FTFN adapting high input impedances and low and high impedances at the output ports. However, realizing the FTFN using the OFA suffers from all the Op Amp limitations and hence loosing all the inherent advantages of the current mode processing such as higher bandwidth and slew rate. Furthermore, the practical
implementation of the OFA is extremely difficult. The OFA using the current mirroring technique requires matched transistors for accurate current transfer properties and the OFA using the second approach relies on exactly equal supply currents. Several alternative techniques of realizing the FTFN were proposed. For example, a two Op Amp realization of the OFA was suggested in [9]. This approach, however, requires six matched resistors. Moreover, an Op-Amp-OTA (operational transconductance amplifier) based implementation of the FTFN was described by Senani [10]. Although it seems to be a simple technique, a temperature dependent realization constraint must be satisfied. Such requirement is very difficult to be met in practice. In addition, this method combines both the Op Amp and the OTA limitations.

All of the above techniques comprise Op Amps to realize the FTFN which will limit its frequency performance. Alternatively, high frequency performance FTFN could be obtained utilizing the available current mode devices. Recently, several current mode devices trying to approximate the characteristics of the FTFN were proposed. For example, the Operational floating conveyor (OFC) comprises a voltage buffer at the input port and a current follower at the output port [11]. However, similar to a CCII a current is allowed to pass through terminal X of the OFC which determines the voltage at one of the current follower output. Thus, I_X is not zero and the voltages at the two terminal of the current follower are not arbitrary. Also, the current feedback operational amplifier (CFA) with additional output exhibits similar characteristics as the OFC [12]. Moreover, the current conveyor based on voltage mode operational amplifier proposed by Bruun can be
generalized to be equivalent to an FTFN [13]. In this case, two current conveyors and an output voltage buffer are used to implement the Op Amp. If the grounded output of the new Op Amp is utilized and the output buffer is removed then this Op Amp reduces to an FTFN.

Four different FTFN implementations using CCII- were proposed in [14] and are shown in Fig. 1.6. Three of them are derived from previously mentioned current circuit candidates for approximating the FTFN: Fig. 1.6(a) represents the general case of Op Amp based current conveyor suggested by Bruun. Fig. 1.6(c), is the OFC proposed by Toumazou et al [11]. Fig. 1.6(d) is the CFOA with current output and Fig 1.6(b) is a two CCII- based FTFN. Actually, the two current conveyors CCII based FTFN was first indicated in [10].

Fig.1.6: Four FTFN Realizations Based on CCII-
As mentioned previously Fig. 1.6(c) and Fig. 1.6(d) do not fit the FTFN definition since the current associated with the nullator are not zero. That is a current signal can flow through the low impedance terminal X of the CCII- at the input port even when the current of its terminal Z is zero. However, Fig. 1.6(a) and Fig. 1.6(b) satisfy the definition of the FTFN. Both realizations conciliate a nullator with zero currents at the input stage. The associated impedance at both terminals of the nullator is high and ideally infinite. The Norator port definition is met by the two realizations, however, with different characteristics. The circuit of Fig. 1.6(b) generates two equal and opposite currents through the two terminal of the norator. This current is produced from the voltage difference between the two terminals of the nullator due to the small internal resistance of terminal X of the two current conveyors. Thus, the relation between the currents and the input voltage is \( I_x = \frac{V_d}{2R_x} \) where \( V_d \) is the differential input voltage. In fact this realization of the FTFN is no more than a fully differential transconductance amplifier. A transconductance amplifier with two equal and opposite output currents.

However, the circuit of Fig. 1.6(a) incorporates additional Current Conveyor to reproduce the current signal flowing through one terminal of the norator to the other (Terminal (X) and terminal (Z) of the current conveyor respectively). In fact, the currents passing through the two terminals of the norator are not identical. Actually, the currents relation is \( I_z = I_x + I_c \) where \( I_c \) is equivalent to the current produced from the differential input contribution \( I_c = \frac{V_d}{2R_x} \). This current error can be easily removed by disconnecting terminal Z of the input stage current conveyor.
It is clear that the two realizations use different mechanisms to approximate the definition of the FTFN. They exhibit different inherent internal parasitic impedances and output terminals impedances. Consequently, a wide statement of which is the best remains ambiguous. However, depending on a particular application the optimum realization among these candidates can be selected. For example, an FTFN exhibiting high and low impedances at the output ports Fig.1.6(a) would be suitable for implementing the four different types of amplifiers. Since the desired impedance level at each stage can be achieved by applying the proper feedback.

However, for applications incorporating all terminals of FTFN in the internal circuit and associated with a complex circuitry, it becomes more difficult to decide the best choice. For instance, the design of filters and oscillators may result in a multi-feedback circuit. Hence in such a case the advantageous candidate among these becomes unclear. In such cases, any nonzero equal and opposite currents generated at the two terminal of the norator regardless of their values will provide approximately the same results. This is supported by Carlossena el at. [14] who demonstrated that all realizations perform approximately the same when they are used to implement a floating frequency dependent negative resistor (FDNR) of a notch filter. Note that all the FTFN terminals are internally connected. Thus all of them proved to be eligible for the implementation of the FTFN in such applications.
It is clear that the optimum realization among the two ways (Fig.1.6(a) and (b)) is dependent on the specific application. In applications such as sinusoidal oscillators based on a single FTFN where an output current is required, the two realizations are expected to perform similarly. So minimum number of CCII would be used to realize the FTFN. However, in applications where it is required to convey a well determined current to a high impedance output, the FTFN of Fig.1.6(a) is advantageous and that of Fig.1.6(b) can not be used.

There is no commercially available CCII-. However, there are several ways by which the available CCII+ (AD844) can be configured as CCII- as shown in Fig.1.7. This will result in several different equivalent circuit of the two possible implementations. Fig.1.8 shows two possible arrangements of CCII+ to realize FTFN.

Fig.1.7: CCII- Based on CCII+
In summary, there are mainly two approaches to realize the FTFN using the available devices. The first is using Op Amp with additional output current terminal (OFA) and the second is using CCII+. The Op Amp exhibits more precise characteristics than the CCII+. That is closer to unity voltage gain between Y and X terminals and less values of offset voltages and input error currents \((I_x\) and \(I_y\)) [15]. However, CCII+ frequency performance is superior to that of the Op Amp. Also, FTFN based on CCII+ is easier for practical implementation. Thus, whenever it is applicable and if the number of CCII+ required to realize FTFN is reasonable, then the two-CCII+ realization of Fig. 1.8(b) will be used for experimenting the circuits to be proposed. On the other hand, if the number is high or CCII+ realization is not suitable, then SPICE simulation using the OFA of Fig. 1.4 will be used to verify the theoretical analysis. Actually, if a large number of active elements such as the CCII+ is used in implementing FTFN based circuits, it may results in unpredictable practical performance of the circuit. For example, this may create undesirable active-R oscillation. This is attributed to the parasitics of the FTFN.
1.3 Literature Review of FTFN Applications

The FTFN is a universal active device having a wide range of possible applications. The most fundamental applications are the voltage and current followers Fig.1.9(a) and Fig.1.9(b) respectively. The voltage follower exhibits the relations \( V_i = V_o \), \( I_i = 0 \) and the output current is determined by the load. The current follower exhibits the relations \( I_i = I_o \), \( V_i = 0 \) and the output voltage is governed by the load.

![Fig.1.9: Fundamental Applications of The FTFN](image)

A typical voltage to current converter (\( I_o = V_i / R \)), capable of driving a grounded load with signal current, is shown in Fig.1.9(c). This is the optimum voltage to current converter using the minimum number of components. Such a voltage to current converter, built around the standard Op Amp requires three Op Amps, four matched resistors in addition
to a transfer determining resistor $R_3$, as shown in Fig. 1.10 [16], and even though its output current is not associated with high impedance. This illustrates the superiority of FTFN over the standard operational amplifier.

Fig. 1.10: Voltage to Current Conveyor Based on Op Amp

Several applications evolve from the basic circuits previously mentioned. Current mode ideal integrator with $I_o=I_{in}/(sRC)$ and differentiator with $I_o=(sCR)I_{in}$ can be obtained directly as illustrated in Fig. 1.11(a) and 1.11(b) respectively.
Fig.1.11: (a) Integrator (b) Differentiator using FTFN

A multiple outputs OTA can be obtained by using two voltage-to-current converters and a resistor as shown in Fig.1.12. However, the transconductance of this OTA is not electronically controllable by the biasing current. Also, an instrumentation amplifier comprising balanced voltage-to-current converter and current-to-voltage converter is shown in [Fig.6 of [16]]. Note that the popular instrumentation amplifier built around the conventional Op Amp requires matched resistors to obtain a high Common Mode Rejection Ratio (CMRR).
Moreover, a fully floating gyrator can be obtained from two balanced voltage to current converters as shown in [Fig. 7 of [16]]. The gyrator plays a very important role in filter designs because of its capability of simulating inductors.

Furthermore, filters and sinusoidal oscillators play essential roles in electronic engineering. They are fundamental building blocks in control, signal processing and communication engineering. In response, filters and sinusoidal oscillators using the FTFN have been
receiving significant attention recently. A systematic method of converting a voltage mode filter using a single Op Amp to current mode filter using the FTFN was described in [17].

In addition, several cascadable universal current mode filters using the FTFN have been proposed. A single FTFN based current mode universal filter that was proposed in [18] suffers from many disadvantages. For example, the denominator of the transfer function associated with LP, HP and BP have a difference term. Thus, these realizations suffer from high passive sensitivity. The voltage tracking error was not considered in the analysis so the active sensitivity remains questionable. Also, many matching conditions have to be met to obtain a particular filter. In addition, it incorporates extensive number of passive components: eight for the notch and the AP, seven for the BP, six for the HP and LP.

Another cascadable universal current filter using a single FTFN was suggested in [19] to avoid some of the drawbacks of the previous filter. Similar to the previous filter, the different realizations are insensitive to the current tracking error except for the current gain. Although the proposed LP and HP are insensitive to the voltage tracking error except for current gain, notch and AP sensitivities are dependent on the passive component values. The BP realization relies severely on a frequency dependent condition to exhibit low sensitivity to the voltage tracking error limiting the maximum operating frequency of the filter. The required number of passive components is reduced to five for the BP, LP and HP and to six for BR and AP. However, the number of floating capacitors is increased to two for AP, BR and BP filters. All of these universal filters [18-19] using a
single FTFN require changes of the circuit topology to realize a specified filter function as well as they rely on cancellation and matching restrictions of the passive elements.

Moreover, insensitive current/voltage mode filters using two FTFNs were proposed in [20]. It is important to note that while the number of active components was increased to two, the number of passive elements was reduced to four. The current mode filter, comprising grounded capacitors, can provide only cascadable BP function at the high output impedance. The additional LP filter can not be cascaded without using extra FTFN or current conveyor. Although the active and the passive sensitivities of the BP filter were shown to be low (i.e. less than one), they remain unclear for the LP. The suggested voltage mode filter is universal and is capable of realizing all of the five functions [20]. This filter incorporates two floating capacitors and two floating resistors. It was claimed that the passive and the active sensitivities are low without showing supporting analysis. However, it is important to note that all of the proposed filters [17-20] suffer from a major disadvantage; that is, the interdependence of the main filter characteristic parameters (the center frequency and the bandwidth).

The sinusoidal oscillator is another essential building block for many applications. It has a wide range of applications in communication. Also, it is considered as a basic element of the signal generator since a square wave and a triangular wave can be easily obtained from a sine wave. Recently, several FTFN based sinusoidal oscillators have been proposed. Senani [21] showed that transferring single Op Amp oscillators, to equivalent oscillator
built around the FTFN, produces several new oscillators having the same characteristic equation. Some of these new oscillators may have interesting properties. The most important feature to note is that current mode oscillators with high output impedance can be obtained. The suggested current mode oscillator is single grounded-resistor controlled (SGRC) incorporating seven passive elements, two of them are floating capacitors.

Moreover, a current mode quadrature sinusoidal oscillator was proposed in [22]. This oscillator is (SGRC) comprising seven elements, among them two are grounded capacitors, the condition of oscillation is controlled from a floating resistor. Two additional current FTFN or current conveyor must be used to obtain the two quadrature currents. However, we discovered that the characteristic equation obtained by the authors does not correspond to the proposed circuit. The circuit was reanalyzed and it was found that the same features of the proposed oscillator can be produced by certain rearrangement of the components. It is interesting to note that the proposed oscillator was found to exhibit the same features as one of Senani oscillators proposed in [21].

More recently, six different single FTFN oscillators obtained from the same characteristic equation were proposed in [23]. Four of those are single element (resistor or capacitor) controlled oscillators using seven passive components. The remaining two are single frequency oscillators with six passive elements. Any of these oscillators uses at least one floating capacitor and none of them is capable of providing output current without adding
additional active elements. Furthermore, a voltage buffer is required to obtain an output voltage without disturbing the circuit oscillation.

1.4 The FTFN Versatility and Flexibility

At present, many current mode basic building blocks, for example, the OTA, CCI, CCII, CFA and FTFN are competing in replacing the well known Op Amp. Actually, the Op Amp is no more than a particular configuration of the FTFN with one terminal of the norator grounded. Thus, it follows that any circuit built around the Op Amp is achievable using the FTFN. Moreover, it was previously shown that the FTFN can be utilized in several applications where it is very difficult to use the standard Op Amp.

The second generation current conveyor (CCII) is one of the most well established current mode devices. Most of the researchers consider the CCII to be the most versatile current mode building block [2,24,25]. In fact, the CCII- is no more than a special case obtainable from the FTFN by connecting one terminal of the Norator to the Nullator. This implies that all of the CCII- based circuits are directly achievable using FTFN.

In this section, the advantages of the FTFN over the CCII will be highlighted. It will be shown that the FTFN is more versatile and flexible than all of the existing devices including CCII. For instance, a systematic approach of transferring a voltage mode circuit
to a current mode circuit and vice versa using the duality principle is very well known
[26]. A direct application of this theorem is to convert the well developed voltage mode
circuits based on Op Amp to current mode counterparts [27-30]. This will allow the
researchers to utilize the wealthy background of the Op Amp-based circuits in various
applications. The procedure of transferring voltage mode circuit using the Op Amp to
current mode using the duality principle consists of three steps. First, the Op Amp is
represented by the its nullor model. Then, the dual of the circuit is obtained utilizing the
fact that Norator and Nullator are dual pair. Finally, the Nullor is replaced by equivalent
active devices. In most applications the resultant circuit would comprise nullors that can
not by realized by any active device except the FTFN [10,17, 27-30]. For example, this
approach was used to convert the Sallen Key LP filter based on the Op Amp to a current
mode equivalent. It was found that the resulting circuit can not be implemented by any
active device other than the FTFN [17]. Also, it was mentioned previously that Senani
[21] described a systematic method of converting an Op Amp-based oscillator circuit to
FTFN counterparts and some of the new realizations would exhibit interesting features. In
fact it is worth mentioning that this approach can not be applied using active devices other
than the FTFN.

Furthermore, a novel approach to convert an arbitrary grounded driving point impedance
into a floating equivalent was suggested in [31]. This method also relies on the FTFN to
perform the transformation where all other active devices could not fit. This approach has
a wide range of applications. For example, it was used to transfer synthesized lossless grounded inductor (LGI) to an equivalent lossless floating inductor (LFI) [31]. Moreover, this technique was applied to convert a grounded lossy inductor (with series or parallel RL impedance) to associated floating inductor [10].

Another application demonstrating the versatility and flexibility of the FTFN over the CCII was described in [32]. A first-order AP filter implemented using FTFN was compared with an equivalent circuit realized by CCII. The CCII realization comprises a capacitor between terminal X and ground. It was found that such connection produces unstable operation because X exhibits a low input resistor. This parasitic resistor combined with the externally connected capacitor produces a pole which results in unstable operation. On the other hand, the FTFN based circuit was found to be stable. Actually, in this application the FTFN was configured as CCII- with high impedance at X. This implies that a CCII-derived from an FTFN is more flexible the standard CCII-. Furthermore, the restriction of grounding terminal X of CCII through a capacitor limits or complicates the design of several applications. Actually, a parasitic pole is generated due to externally connecting a capacitor to the small internal resistance of the X terminal. Even if such parasitic pole does not cause instability, the frequency performance will be limited. At high frequencies the internal resistance of terminal X will be dominant altering the transfer function obtained assuming ideal CCII. Moreover, there are several applications that require connecting a capacitor to terminal X such as ideal differentiator as shown in Fig.1.11(b). Furthermore,
starting from a particular filter (BP, HP, and LP) connecting a capacitor to terminal X would make it easy to obtain the other two functions as illustrated in Fig. 1.13.

Fig. 1.13: Obtaining the Basic Filters (LP, BP and HP) From Each Others
1.5 Approaching The Universal Element

In the light of the previous discussion, the FTFN appears to be a very versatile and flexible active element. It has been found that it is not only possible for the FTFN to replace the conventional Op Amp in all possible applications, but it extends applications range to much more. In addition, a special configuration of the FTFN reduces to CCII- with better characteristics expanding the possible range of applications and giving more freedom to the designers. One difficulty, however, of the FTFN is that it is impossible to configure it as a CCII+. CCI- or CCI+. Therefore, it is essential to solve this problem before considering the FTFN as a universal element.

The idea of designing multi-output devices instead of single output ones proved to be useful in many applications. Multiple output OTAs [33-35] and multiple output current conveyors [36,37] proved to be very powerful in many applications. It was demonstrated that multiple output devices would reduce the number of active components required for those particular applications tremendously. In current mode the price paid to obtain multiple output devices is very cheap. It requires no more than adding additional current mirrors. If this idea is applied to the FTFN, similar results are expected. New circuits would directly evolve. Most interesting to be noted is that both CCII- and CCII+ will be particular applications of the universal FTFN as illustrated in Fig.1.14. Moreover, both types of the first generation current conveyors (CCI- and CCI+) are obtainable as shown in Fig.1.15.
Also, new additional applications are possible. For example, first order current mode
LP\{I_o/I_{in}=1/(1+sRC)\}, HP\{I_o/I_{in}=sRC/(1+sRC)\} and AP\{I_o/I_{in}=(1-sCR)/(1+sRC)\} filters
using minimum number of components is shown in Fig.1.16. Furthermore, it will be
illustrated later that this additional current output will be very powerful in filters design.
Although the FTFN can provide output current associated with high impedance, similar to the current conveyor, an output voltage from low output impedance calls for a voltage buffer. It was found that adding such a voltage buffer to the current conveyor allows the extension of the Current Feedback Amplifier (CFA) applications to cover voltage mode in addition to current mode. Thus, for the FTFN to combine both voltage mode and current mode application a voltage buffer has to be added. Equivalent results of extending the
applications of the FTFN by incorporating the voltage buffer are expected similar to those of the CFA. Thus, after adding additional current output and voltage buffer to the FTFN the universal element (UE) is achieved as illustrated in Fig.1.17. The UE exhibits the following terminal characteristics: $V_x = V_y$ ($V_{xy} = 0$), $I_x = I_y = 0$, $I_z = I_{w1} = I_{w2}$ and $V_{w2} = Z_{w2}I_{w2}$ where $Z_{w2}$ is the impedance connected between terminal $W_2$ and the ground. This UE can realize the basic types of active elements: the conventional Op Amp, OTA, CCI-, CCI+, CFA, CCI-, CCI+, as well as the FTFN.

![Fig.1.17 The universal element (UE): (a) UE from FTFN (b) Symbol](image)

### 1.6 Problem Definition

According to the literature review the FTFN is becoming increasingly important with growing opportunities as the dream of realizing the practical FTFN becomes feasible. Based on the preceding discussion, it is believed that the FTFN is the most versatile and
flexible active device compared to other available devices. However, only a little work has been done in designing universal filters and sinusoidal oscillators using the FTFN. The main objective of this thesis is to propose new universal filters and sinusoidal oscillators using the FTFN. These circuits are evaluated and compared with those previously suggested in the literature.

Moreover, new and improved universal filters based on CCII+, OTA and CFA are proposed. The advantages and disadvantages of all circuits are studied. A comprehensive comparison between the proposed circuits based on different active devices is performed to clarify the position of the FTFN in this field.

1.7 Thesis Organizations

The thesis addresses the following issues in order. In Chapter 2 active filters based on the FTFN are proposed. Also, new and improved active filters using CCII+, OTA and CFA are presented. The filters based on these different active devices are evaluated and compared. New sinusoidal oscillators based on the FTFN are proposed in Chapter 3. The advantages and disadvantages of those oscillators are discussed. Chapter 4 discusses the nonlinear behavior of the FTFN as it is a possible source of errors in designing FTFN filters and oscillators. The thesis is concluded by some conclusions, recommendations and suggestions for the future work in Chapter 5.
CHAPTER 2

APPLICATIONS OF THE FTFN:

I. ACTIVE FILTERS

2.1 Introduction

A filter is a linear circuit that shapes the frequency spectrum of the input signal and its phase according to its magnitude and phase response respectively. Filters are mainly used to perform frequency selection actions. That is passing signals having frequencies within a particular range, and rejecting signals whose frequency spectrum lies outside this range.
The major types of such filters are: low-pass (LP), high-pass (HP), band-pass (BP) and band-reject (BR) which is also called band-stop (BS) or notch.

Furthermore, there are applications where the phase response of the filter is of major interest. This introduces the fifth type of the basic filters namely the all-pass (AP). The AP filter has ideally a constant amplitude response at all frequencies, but its phase response is frequency selective. The AP filters are used as phase shifters. They are utilized as phase shapers in some systems such as the delay equalizers to make the overall time delay of transmission constant with frequency.

Active filters are essential building blocks in communication and instrumentation systems. The earliest filters design uses only inductors and capacitors and therefore are called passive LC filters. These filters are suitable for high frequency operation. However, the required inductors become large and bulky with low quality factor in low frequency applications (DC to 100kHz)[38].

Moreover, such inductors are incompatible with the modern trend of monolithic integration. The active-RC filters are one of the most powerful and successful inductorless filters. Active filters utilize active elements together with resistors and capacitors. Such filters are fabricated using discrete, hybrid thick-film or hybrid thin-film technology[38].
2.2 Evaluating an Active Filter

The optimum active filter is the one capable of meeting the following engineering standards and requirements:

1. Independent control of the characteristic parameters of the filter such as the center frequency (ω₀) and the bandwidth (ω₀/Q₀) where Q₀ is the quality factor.

2. All capacitors are grounded to absorb the stray capacitance effect and to make it more compatible for integration purpose and high frequency operation.

3. It is advantageous to design a filter where the center frequency and the bandwidth are controllable via grounded resistors.

So these resistors can be replaced by programmable active resistors such as the resistance of JFET or MOSFET transistors. Also, if the control is via a grounded capacitor, electronic tuning is achievable using a varactor. The varactor is a diode specially fabricated to enhance the variable capacitance versus reverse bias characteristic.

4. Grounded resistors are easier to be implemented in the integration process than the floating resistors. Thus, preferably, all the resistors have to be grounded.
5. The filter has to canonical utilizing the minimum number of passive and active components to occupy less area and consume less power.

6. No matching/cancellation requirements for realizing the different filtering functions.

7. The active and passive sensitivities of the filter parameters ($\omega_o$ and $Q_o$) with respect to the active and passive components have to be low (less than unity).

The actual response of the practically implemented filter will deviate from the ideal response because of the FTFN non-ideal characteristics and the tolerances in component values. The concept of sensitivity is employed to predict such deviations. Particularly, for second-order filters a designer is usually interested in finding how their main parameters, namely the center frequency and the quality factor, will vary in response to variations (both original tolerances and future changes) in RC components values and the FTFN nonideality, namely the voltage and current tracking errors. The well known sensitivity function $S_x^y$ is defined as[38]:

$$S_x^y = \lim_{\Delta x \to 0} \frac{\Delta y / y}{\Delta x / x} = \frac{\partial y}{\partial x} \frac{x}{y}$$

Where, $x$ designates the value of a component (a passive element or voltage or current tracking error) and $y$ represents a filter parameter of interest such as $\omega_o$ and $\omega_o/Q_o$. Therefore, the value of sensitive $S_x^y$ determine the per-unit change in $y$ as a result of a
specified per-unit change in \( x \). For example, if the sensitivity of \( \omega_n \) relative to \( R_1 \) is 0.5, then a 1% increase in \( R_1 \) results in a 0.5% increase in the value of \( \omega_n \).

These standards will be used to evaluate the filter circuits which will be proposed.

2.3 Universal Second-Order Filters

The simplest filter transfer functions are those of the first and second order. These are useful by themselves in the design of simple filters. Moreover, high-order filters can be realized by cascading first-order and second-order filters. In fact, the cascade design is the most popular approach for the design of active filters since a high-order transfer function can be factored into the product of second order functions. If the transfer function is odd, an additional first order function is required.

Such active realizations provide several advantages over their passive counterparts. For instance, the output impedance of the active circuit is very low (for voltage mode) or very high (for current mode) making cascading achievable. Also, some of the transfer function parameters can be adjusted without disturbing others and the gain can be set to any desired value.
A universal voltage or current mode filter is capable of realizing the five different basic functions (LP, HP, BP, BR, AP filters). There are two main categories of universal filters: multiple inputs single output (MISO), and single input multiple outputs (SIMO). The former can realize each function separately by controlling the inputs while the later can provide all the five functions simultaneously.

2.4 Universal Multiple Input Single Output (MISO) Current Mode Filters Using FTFN

According to the literature survey [6,8-11,14,16-23,27-32], no universal current mode (MISO) filter using FTFN has been reported. Two such circuits are proposed and discussed in this section. The first uses only three FTFNs. The second uses four FTFNs. However, it overcomes some of the disadvantages of the first filter.

2.4.1 Universal (MISO) Current Mode Filter (I)

The first filter circuit comprising three FTFNs is shown in Fig.2.1. The proposed filter enjoys these attractive features: cascadability, since its output current is associated with high impedance, independent control of the center frequency and the bandwidth via grounded elements, fewer number of active elements compared with other devices, such as
CCs counterparts, as will be highlighted in the following section, and all the capacitors are grounded.

![Image of a circuit diagram]

**Fig. 2.1: Proposed (MISO) Current-Mode Universal Filter (I)**

Routine analysis of this circuit assuming ideal FTFN, characterized by \( I_x = I_y = 0 \), \( V_x = V_y \) and \( I_z = I_w \), yields the following current transfer function:

\[
I_2 = \frac{G_z}{G^*} \frac{s^2 C_1 C_4 I_{in1} - sC_4 G_2 I_{in2} + (sC_6 G_2 + sC_2 G_2 + G_2 G_3)I_{in3}}{s^2 C_4 C_6 + sC_6 G_z + sC_2 G_z + G_3 G_z}
\]  
(2.1)

It can be seen from (2.1) that:

1. LP response is realized with \( I_{in1} = 0, I_{in2} = I_{in3}, C_3 = 0, C_4 = C_6 \).
2. HP response is realized with \( I_{in2} = I_{in3} = 0 \).
3. BP response is realized with \( I_{in1} = I_{in3} = 0 \).
4. BR response is realized with \( I_{in1} = I_{in2} = I_{in3}, C_3 = 0, \) and \( C_4 = C_6 \).
5. AP response is realized with \( I_{in1} = I_{in2} = I_{in3}, C_3 = 0, \) and \( C_4 = 2C_6 \).
From (2.1), the center frequency ($\omega_c$), bandwidth ($\frac{\omega_c}{Q_o}$) and quality factor ($Q_o$) can be respectively, expressed as:

$$\omega_c = \sqrt[2]{\frac{G_3 G_s}{C_4 C_o}}$$  \hspace{1cm} (2.2)

$$\frac{\omega_c}{Q_o} = \frac{(C_3 + C_o)G_s}{C_4 C_o}$$  \hspace{1cm} (2.3)

$$Q_o = \frac{\sqrt{C_4 C_o G_3}}{\sqrt{G_3 (C_3 + C_o)}}$$  \hspace{1cm} (2.4)

Also, the LP, BP and HP gains are given by:

$$G_{LP} = 1$$  \hspace{1cm} (2.5)

$$G_{BP} = \frac{C_4}{C_3 + C_o}$$  \hspace{1cm} (2.6)

$$G_{HP} = \frac{C_1 G_3}{C_6 G_2}.$$  \hspace{1cm} (2.7)

It can be seen from (2.1)-(2.3) that this configuration is mostly suitable for realizing HP and BP where the center frequency and the bandwidth can be controlled independently via grounded elements $R_3 = 1/G_3$ and $C_3$ respectively. However, the bandwidth and the center frequency of the LP, BR and the AP can be orthogonally adjusted. That is the passive elements ($G_3$, $C_4$ and $C_6$) are first selected to meet the bandwidth requirement and then the center frequency requirement is satisfied by adjusting $R_3 = 1/G_3$. Moreover, the LP
realization requires two equal currents, and BR realizations require three identical input currents and some passive components matching requirements.

Nonideal analysis is required to find the active sensitivity of the filter. Assuming that the port relations of the FTFN can be expressed as $I_x = I_y = 0$, $V_x = \beta V_y$, and $I_z = \alpha_1 I_u$ where $\beta = 1 - \varepsilon$ ($|\varepsilon| << 1$) denotes the voltage tracking error of the FTFN and $\alpha = 1 - \delta$, ($|\delta| << 1$), represents the current-tracking error, reanalysis of the circuit shown in Fig. 2.1 yields the current-transfer function:

$$I_o = \alpha_1 \frac{G_1 \alpha_3 \beta \beta_s \hat{s}^2 C_1 C_4 I_m_1 - \alpha_2 \beta s G_2 I_m_2 + (\alpha_3 \beta \beta_s s G_2 + \alpha_1 \beta C_1 G_2 + \alpha_2 G_2 G_4) I_m_1}{\alpha_3 \beta \beta_s \hat{s}^2 C_4 C_6 + \alpha_1 \beta \beta_s s G_2 + \alpha_3 C_1 G_2 + \alpha_2 G_5 G_5}$$

(2.8)

where $\beta_i, \alpha_i, i = 1-3$, are the voltage- and the current-tracking errors of the $i$th FTFN. From (2.8) the center frequency, the bandwidth and quality factor can be, respectively, expressed as:

$$\omega_o = \sqrt{\frac{\alpha_3 G_3 G_5}{\alpha_1 \beta \beta_s C_4 C_6}}$$

(2.9)

$$\frac{\omega_o}{Q_o} = \frac{\alpha_3 C_1 G_5 - \alpha_1 \beta \beta_s C_6 G_5}{\alpha_1 \beta \beta_s C_4 C_6}$$

(2.10)

$$Q_o = \frac{\sqrt{\alpha_3 \beta \beta_s C_4 G_5}}{\sqrt{G_1 (\alpha_3 C_3 + \alpha_1 \beta \beta_s C_6)}}$$

(2.11)
It can be shown that the passive and active sensitivities of the parameters $\omega_n$ and $Q_n$ can be expressed as.

\[
S_{\omega_n} = S_{Q_n} = -S_{C_n} = S_{C_n} = -S_{\omega_i} = S_{\omega_i} = S_{\omega_n} = S_{\omega_n} = -\frac{1}{2}, S_{\omega_i} = S_{\omega_i} = 0
\]

\[
S_{C_n} = S_{C_n} = -S_{C_n} = S_{C_n} = -\frac{\alpha_1 C_3}{\alpha_2 C_3 + \alpha_1 \beta_2 \beta_3 C_3}, S_{\omega_i} = S_{\omega_i} = -\frac{\alpha_1 \beta_2 \beta_3 C_3}{\alpha_2 C_3 + \alpha_1 \beta_2 \beta_3 C_3}
\]

Thus all the active and passive sensitivities are no more than unity.

The circuit was simulated using PSPICE Simulation program. Each FTFN is realized by applying the power sensing technique to Op Amp as shown in Fig. 1.4. A PSPICE macromodel for Op Amp power supply current sensing was described in [39]. This macromodel may be incorporated in PSPICE with Op Amp macromodel to predict variations in power supply current with load current. The proposed macromodel was tested by simulating a precision full-wave rectifier, using the (uA741) Op Amp and Wilson current mirrors implemented using PSPICE default transistors model, and found to perform very well [39]. We have developed our own subcircuit of FTFN by applying this macromodel to the built-in PSPICE macromodel of LF156 Op Amp to simulate the power supply current sensing realization of the FTFN of Fig. 1.4. The Wilson current mirrors are built using the default transistor parameters of the PSPICE. The simulation results of LP, HP, BP, BR and AP are shown in Fig. (2.2).
Fig. 2.2: (a) Simulation results of the LP with: $C_1 = C_3 = C_6 = 100\,\text{pF}$, $C_4 = 700\,\text{pF}$, $R_2 = 40\,\text{k}\Omega$.

$R_3 = 1\,\text{k}\Omega$, $R_e = 40\,\text{k}\Omega$. '—' Theoretical results, 'x' Simulation results

Fig. 2.2: (b) Simulation results of the HP with: $C_1 = 5\,\text{nF}$, $C_3 = C_6 = 1\,\text{nF}$, $R_2 = 40\,\text{k}\Omega$.

$R_3 = 10\,\text{k}\Omega$, $R_e = 20\,\text{k}\Omega$. '—' Theoretical results, 'x' Simulation results
Fig. 2.2: (c) Simulation results of the BP with: $C_1=5\text{nF}$, $C_2=C_3=100\text{pF}$, $C_4=700\text{pF}$, $R_s=40\text{k}\Omega$, $R_1=1\text{k}\Omega$, $R_p=40\text{k}\Omega$. '—' Theoretical results, 'x' Simulation results.

Fig. 2.2 (d) Simulation results of the BP with: $C_1=5\text{nF}$, $C_2=1000\text{pF}$, $C_4=7000\text{pF}$, $R_s=40\text{k}\Omega$, $R_1=1\text{k}\Omega$, $R_p=40\text{k}\Omega$. '—' Theoretical results, 'x' Simulation results.
Fig. 2.2 (d) Simulation results of the BR with: $C_1=\dot{C}_1=\dot{C}_4=1\text{nF}$, $C_5=0$, $R_1=R_6=10\text{k}\Omega$.

- 'Theoretical results', 'x' Simulation results.

---

Fig. 2.2 (e) Simulation results of the AP Magnitude Response with: $C_1=\dot{C}_4=1\text{nF}$, $C_3=0$,

$C_4=2\text{nF}$, $R_2=R_3=R_6=10\text{k}\Omega$. '—'Theoretical results', 'x' Simulation results.
Fig. 2.2(f) Simulation results of the AP Phase Response with: $C_1=C_6=1\text{nF}$, $C_2=0$, $C_3=2\text{nF}$.

$R_1=R_3=R_c=10\text{kΩ}$. '—' Theoretical results, 'x' Simulation results

It is clear that the simulation results are in very good agreement with the theoretical analysis except for the BP response of Fig. 2.2(c) where there is some error. This deviation is mainly due to frequency limitation of the Op Amp. This is supported by the simulation results of the BP filter, with lower center frequency, of Fig. 2.2(d) which shows excellent accuracy.

2.4.2 Universal (MISO) Current Mode Filter (II)

A new current mode (MISO) universal filter is presented. This configuration overcomes some of the disadvantages of the previous proposed filter, namely the matching or
cancellation requirements for most of the filter functions. Also, it exhibits additional advantages. The structure of this biquad filter consists of four FTFNs, three grounded capacitors and four grounded resistors. The circuit is capable of realizing cascaddable second-order LP, HP, BP, BR and AP filters without changing its topology. The proposed circuit enjoys the attractive features of low active and passive sensitivities and independent control of the parameters $\omega_o$ and $\omega_o/Q_o$ using grounded elements. The proposed circuit is shown in Fig. 2.3.

Fig.2.3: Proposed (MISO) Current-Mode Universal Filter (II)

Using ideal port relations of the FTFN characterized by $I_x=I_y=0$, $V_x=V_y$ and $I_z=I_w$, the circuit’s current transfer function can be expressed as:

$$I_o = \frac{s^2C_2C_4G_5I_{in1} - sC_2G_5G_6I_{in2} + G_3G_5G_6I_{int}}{s^2C_2C_4G_6 + sC_1G_1G_5 + G_1G_3G_5}$$

(2.12)
It can be seen from (2.12) that:

1) LP function is obtained when $I_{m2}=I_{m3}=0$

2) HP function is obtained when $I_{m1}=I_{m3}=0$.

3) BP function is obtained when $I_{m1}=I_{m2}=0$.

4) BR function is obtained when $I_{m2}=0$ and $I_{m1}=I_{m3}$.

5) AP function is obtained when $I_{m1}=I_{m2}=I_{m3}$ and $C_1=C_2$, $G_1=G_5=G_7$.

Also, it can be seen from (2.12) that the center frequency, the bandwidth and the equality factor can, respectively, be expressed as:

$$\omega_o = \frac{G_s G_s G_s}{\sqrt{C_s C_s G_s}}$$

(2.13)

$$\frac{\omega_o}{Q_o} = \frac{C_s G_s G_s}{C_s C_s G_s}$$

(2.14)

$$Q_o = \frac{1}{C_s} \sqrt{\frac{C_s C_s G_s G_s}{G_s G_s}}$$

(2.15)

Furthermore, from (2.5) it can be seen that the LP, the HP and the BP gains are as follows.

$$G_{LP} = \frac{G_s}{G_1},$$

(2.16)

$$G_{HP} = 1,$$

(2.17)

$$G_{BP} = \frac{C_s G_s}{C_s G_1}$$

(2.18)
From (2.13) and (2.14) it can be seen that the parameter \( \omega_\alpha \) can be adjusted by controlling the resistor \( R_1 = 1/G_1 \) without disturbing the parameter \( \omega_\omega/Q_\omega \). Also, the parameter \( \omega_\omega/Q_\omega \) can be adjusted by controlling the capacitor \( C_1 \) without disturbing the parameters \( \omega_\alpha \).

Thus, the proposed circuit enjoys independent control of the parameters \( \omega_\alpha \) and \( \omega_\omega/Q_\omega \). Also, it can be seen from (2.13) and (2.15) that the circuit provides orthogonal control of the quality factor and the center frequency. That is the center frequency is tuned first by selecting \((G_1, G_3, G_5, G_6, C_2 \text{ and } C_4)\) then the quality factor requirement is adjusted via \( C_1 \).

Nonideal analysis is required to find the active sensitivity of the filter. Assuming that the port relations of the FTFN, can be expressed as \( I_x = I_y = 0, V_y = \beta V_x \) and \( I_w = \alpha I_z \) where \( \beta = 1-\varepsilon \) (\( |\varepsilon| < 1 \)), denote the voltage tracking error of the FTFN and \( \alpha = 1-\delta, (|\delta| < 1) \), represents the current-tracking error, routine analysis of the circuit shown in Fig 2.3 yields the current-transfer function:

\[
I_o = \frac{\beta_4 \alpha_4}{s^2 C_2 C_4 G_6 I_{n2} - s \frac{\beta_3}{\alpha_3} C_2 G_2 G_6 I_{n1} + s \frac{\beta_2 \beta_3}{\alpha_1 \alpha_2 \alpha_3} G_1 G_3 G_6 I_{m1}}
\]

\[
I_o = \frac{\beta_3}{s^2 C_2 C_4 G_6} + s \frac{\beta_1 \beta_2 \beta_3}{\alpha_1 \alpha_2 \alpha_3} C_1 G_3 G_5 + \frac{\beta_1 \beta_2 \beta_3}{\alpha_1 \alpha_2 \alpha_3} G_1 G_3 G_5
\]

\[\text{(2.19)}\]

where \( \beta_i, \alpha_i, i = 1-4 \), are the voltage- and the current-tracking errors of the \( i \)th FTFN. From (2.19) the center frequency, the bandwidth and quality factor can be, respectively, expressed as:

\[
\omega_\alpha = \sqrt{\frac{\beta_1 \beta_2 \beta_3 G_1 G_3 G_5}{\alpha_1 \alpha_2 \alpha_3 C_2 C_4 G_6}}
\]

\[\text{(2.20)}\]
\[
\frac{\omega_*}{Q_*} = \frac{\beta_1\beta_2\beta_3\alpha_1 G_1 G_*}{\alpha_3\alpha_2\alpha_1 C_1 C_2 G_*},
\]
(2.21)
\[
Q_* = \frac{1}{C_1} \sqrt{\frac{\alpha_3\alpha_2\alpha_1 C_1 C_2 G_1 G_*}{\beta_1\beta_2\beta_3 G_1 G_*}},
\]
(2.22)

From (2.20) and (2.22) it is easy to show that the active and passive sensitivities of the parameters \( \alpha_0 \) and \( Q_0 \) are given by:

\[
S_{\alpha_0}^{\omega_0} = S_{\alpha_0}^{\omega_0} = S_{\alpha_0}^{\omega_0} = -S_{\alpha_0}^{\omega_0} = -S_{\alpha_0}^{\omega_0} = \frac{1}{2}
\]
\[
S_{\alpha_1}^{\omega_0} = S_{\alpha_1}^{\omega_0} = S_{\alpha_1}^{\omega_0} = -S_{\alpha_1}^{\omega_0} = -S_{\alpha_1}^{\omega_0} = \frac{1}{2}
\]
\[
S_{\alpha_2}^{\omega_0} = S_{\alpha_2}^{\omega_0} = S_{\alpha_2}^{\omega_0} = -S_{\alpha_2}^{\omega_0} = -S_{\alpha_2}^{\omega_0} = \frac{1}{2}
\]
\[
S_{\alpha_3}^{\omega_0} = S_{\alpha_3}^{\omega_0} = S_{\alpha_3}^{\omega_0} = -S_{\alpha_3}^{\omega_0} = -S_{\alpha_3}^{\omega_0} = \frac{1}{2}
\]
\[
S_{\alpha_1}^{\omega_0} = -1, S_{\alpha_2}^{\omega_0} = 0
\]

All of which are small (no more than unity).

Experimental results to verify the theoretical analysis were performed. The FTFN was implemented using the two current conveyors (AD844) realization of Fig.18(b). The experimental results of LP, HP, BP, BR and AP shown in Fig.2.4 agree very well with the presented theory. The slight deviation of the experimental results from the theoretical is mainly due to the stray capacitance and the nonideal practical performance of the current conveyors used to realize the FTFN.
Fig. 2.4: (a). Experimental results of the LP with: $R_1=R_3=R_6=R_8=5k\Omega$, $C_1=C_2=C_4=1.2nF$.

'—' Theoretical results, '•' Experimental results

Fig. 2.4: (b). Experimental results of the HP with: $R_1=R_3=R_6=R_8=5k\Omega$, $C_1=C_2=C_4=1.2nF$.

'—' Theoretical results, '•' Experimental results
Fig. 2.4: (c). Experimental results of the LP with: $R_1=2k\Omega$, $R_3=R_5=R_6=5k\Omega$, $C_1=200\text{pF}$.

$C_1=C_4=1.2\text{nF}$. ‘—’ Theoretical results. ‘*’ Experimental results

Fig. 2.4: (d). Experimental results of the BR with: $R_1=R_3=R_5=R_6=5k\Omega$, $C_1=C_2=C_4=1.2\text{nF}$.

‘—’ Theoretical results. ‘*’ Experimental results
Fig. 2.4:(e). Experimental results of the AP Magnitude Response with: $R_1=R_2=R_4=R_5=5k\Omega$.

$C_1=C_2=C_4=1.2nF$. '—' Theoretical results, '*' Experimental results.

Fig. 2.4:(f). Experimental results of the AP Phase Response with: $R_1=R_2=R_5=R_4=5k\Omega$.

$C_1=C_2=C_4=1.2nF$. '—' Theoretical results, '*' Experimental results.
In summary, a universal current-mode filter has been presented. The circuit has three inputs and one output and can realize all the five basic filters without any changes in the circuit topology. The proposed filter enjoys the following advantages:

(a) All the passive elements are grounded.

(b) Independent tuning of the parameters \( \omega_n \) and \( \omega_n/Q_n \)

(c) Low active and passive sensitivities.

(d) No matching/cancellation requirements except for the AP

(e) Cascadable

2.5 Comparison with CCII+ Counterparts

As mentioned previously there is no FTFN (MISO) or (SIMO) filter suggested in the literature to be compared with the proposed filters. Thus, the proposed filters will be compared with current conveyor (CC) counterparts. A universal (MISO) current mode filters using five CCs was proposed in [40]. It comprises three CCII+, two CCII- and eight grounded passive elements and enjoys independent control of the filter parameters. Another such filter which uses less number of active and passive elements was proposed in [41]. This filter incorporates only two CCII-, two CCII+ and four grounded passive elements, however, its parameter can not be adjusted independently.
On the other hand, several (SIMO) CC based filters were proposed. A current mode biquad combines both the first and the second generations of CC was proposed by Senani [42]. It comprises one CCII+, two CCII-, four CCI- and six grounded passive elements and enjoys independent control of \( \omega_0 \) and \( \omega_0/Q_0 \). Another (SIMO) current mode filter using five CCs (one CCII+, one CCII-, one CCI+, two CCI-) and four grounded passive components was reported in [43]. However, the filter parameters can not be tuned independently.

Furthermore, Chang proposed a (SIMO) filter which comprises only the second generation of CC (four CCII+, three CCII-) and ten grounded passive elements [44]. The center frequency and the bandwidth of that filter can be controlled independently. Moreover, a similar filter using same number of passive and active components was suggested in [45]. The CCs used are five CCII+ and two CCI- and independent control of the filter parameters can be obtained.

It can be seen that none of the previously published filters incorporates only one type of CCs. For a fair comparison, the FTFN based filters should be compared with those built around only one type of CCs. Since CCII+ is commercially available, circuits based only on CCII+ are attractive. Also, the use of only CCII+ simplifies the configurations. Thus, new (MISO) and (SIMO) filters based on only CCII+ are proposed. Those novel filters exhibit most of the attractive features of the already known filters based on more than one type of CCs and offer additional advantages.
2.5.1 New CCII+ based (MISO) Current Mode Filter

A novel universal current-mode filter with three inputs and one high impedance output is presented. The proposed circuit, using only four plus-type second-generation current-conveyors, grounded resistors and capacitors, is shown in Fig.2.5.

Fig.2.5: Proposed (MISO) Current Mode Filter Based on CCII+

Using the standard notations, the CCII+ characteristics can be described by $i_c=\alpha i_v, v_c=\beta v_v$,

where $\alpha=1-\delta, (|\delta|<1)$, denotes the current-tracking error, $\beta=1-\varepsilon (|\varepsilon|<1)$ represents the voltage-tracking error. The single output current $I_o$ can be expresses as:

$$I_o = \frac{\alpha_4 \beta_4 G_1 \alpha_2 \beta_2 G_4 I_m - \alpha_4 \beta_4 \alpha_2 \beta_2 G_4 I_m}{G_4 s^2 C_1 C_5 + s C_1 C_5 + \alpha_1 \alpha_2 \alpha_3 \beta_1 G_2 G_6}$$  \hspace{1cm} (2.23)

From (2.23) it can be seen that:
(1) LP function is obtained when $l_{m2}=l_{m3}=0$

(2) HP function is obtained when $l_{m1}=l_{m3}=0$

(3) BP function is obtained when $l_{m1}=l_{m3}=0$.

(4) BR function is obtained when $l_{m2}=0$ and $l_{m1}=l_{m3}$

(5) AP function is obtained when $l_{r1}=l_{m2}=l_{m3}$ and $G_4=G_5$ and $C_3=C_6$

It can be seen from (2.23) that the center frequency, the bandwidth and the quality factor can be respectively expressed as:

$$\omega_c = \sqrt{\frac{\alpha_1 \alpha_2 \alpha_3 \beta_1 \beta_2 G_3 G_5}{C_4 C_6}}$$

(2.24)

$$\frac{\omega_c}{Q_c} = \frac{G_6}{C_6}$$

(2.25)

$$Q_c = \frac{1}{G_6} \sqrt{\frac{\alpha_1 \alpha_2 \alpha_3 \beta_1 \beta_2 C_4 G_3 G_5}{C_4}}$$

(2.26)

Also, from (2.23) it can be seen that the LP, HP and BP gains are approximately given by:

$$G_{LP} \equiv \frac{G_6}{G_3}$$

(2.27)

$$G_{HP} \equiv \frac{C_3 G_7}{C_4 G_4}$$

(2.28)

$$G_{BP} \equiv \frac{G_7}{G_6}$$

(2.29)

From (2.24) and (2.25) it can be seen that the parameter $\omega_c$ can be adjusted by controlling the resistor $R_3=1/G_2$ without disturbing the parameter $\omega_c/Q_c$ and the parameter $\omega_c/Q_c$ can
be adjusted by controlling $R_c=1/G_c$ without disturbing the parameters $\omega_o$. Thus the proposed circuit enjoys independent control of the parameters $\omega_o$ and $\omega_o/Q_o$. Also, from (2.24) and (2.26) it is easy to show that the active and passive sensitivities of the parameters $\omega_o$ and $Q_o$ are given by:

$$S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = 0$$

$$S_{\omega_o} = S_{\omega_o} = -S_{\omega_o} = -S_{\omega_o} = 0$$

$$S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = 0$$

$$S_{\omega_o} = S_{\omega_o} = S_{\omega_o} = 0$$

All of which are small (no more than unity).

In summary, the proposed circuit enjoys the following attractive features:

(a) Use of one type of second-generation current-conveyor (i.e. CCII+)

(b) Independent control of the parameters $\omega_o$ and $\omega_o/Q_o$.

(c) Use of grounded capacitors and grounded resistors.

(d) Enjoy low active and passive sensitivities.

The proposed circuit was used to realize LP, HP, BP, BR and AP. A sample of the experimental results obtained by using the AD844 current-conveyor, which confirms the presented theoretical analysis, is shown in Fig.2.6.
Fig. 2.6: (a) Experimental results of the LP with: $C_1 = C_3 = C_6 = 470 \, \text{pF}$, $R_1 = R_2 = R_4 = R_5 = 5 \, \text{k\Omega}$.

$R_3 = 4 \, \text{k\Omega}$. '—' Theoretical results, '*' Experimental results.

Fig. 2.6: (b) Experimental results of the HP with: $C_1 = C_3 = C_6 = 470 \, \text{pF}$, $R_1 = R_2 = R_4 = R_5 = 5 \, \text{k\Omega}$.

$R_3 = 4 \, \text{k\Omega}$. '—' Theoretical results, '*' Experimental results.
Fig. 2.6: (c) Experimental results of the AP Magnitude Response with: $C_1 = C_2 = C_3 = 470 \mu F$.

$R_1 = R_2 = R_3 = 5k\Omega$, $R_4 = 4k\Omega$. '—' Theoretical results, '*' Experimental results.

Fig. 2.6: (d) Experimental results of the AP Phase Response with: $C_1 = C_2 = C_3 = 470 \mu F$.

$R_1 = R_2 = R_3 = 5k\Omega$, $R_4 = 4k\Omega$. '—' Theoretical results, '*' Experimental results.
It can be seen from Fig. 2.6 that the experimental results are in good agreement with the theoretical analysis. The slight variation between the experimental and theoretical results are mainly due to the stray capacitance, the parasitic capacitance of the conveyors, and the nonideal characteristics of the AD844.

It can be seen from the previous discussion that there are some similarities and differences between the proposed filter based on CCII+ of Fig. 2.5 and the FTFN based filter of Fig. 2.3. Table 2.1 summarizes the results of a comprehensive comparison between these two filter realizations.

<table>
<thead>
<tr>
<th>Features</th>
<th>FTFN Universal Filter</th>
<th>CCII+ Universal Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_0$ and $\omega_0/Q_0$ Relation</td>
<td>Independent</td>
<td>Independent</td>
</tr>
<tr>
<td>Number of Active Elements</td>
<td>Four</td>
<td>Four</td>
</tr>
<tr>
<td>Number of Passive Elements</td>
<td>Seven</td>
<td>Eight</td>
</tr>
<tr>
<td>Capacitors and Resistors</td>
<td>Grounded</td>
<td>Grounded</td>
</tr>
<tr>
<td>Controlling $\omega_0$</td>
<td>Resistor</td>
<td>Resistor</td>
</tr>
<tr>
<td>Controlling $\omega_0/Q_0$</td>
<td>Capacitor</td>
<td>Resistor</td>
</tr>
<tr>
<td>Active and Passive Sensitivity</td>
<td>Low (less than unity)</td>
<td>Low (less than unity)</td>
</tr>
<tr>
<td>Realization Conditions</td>
<td>Only for AP</td>
<td>Only for AP</td>
</tr>
<tr>
<td>Accuracy</td>
<td>Good</td>
<td>Good</td>
</tr>
</tbody>
</table>
It can be seen from the Table that the FTFN and CCII+ based (MISO) filters are comparable exhibiting almost the same advantages. Although the CCII+ based filter bandwidth can be controlled by a resistor and that of the FTFN based filter by a capacitor, the FTFN filter incorporate less number of passive elements. Moreover, the experimental results of both FTFN and CCII+ based filters agree with theoretical analysis will even that the FTFN was implemented using discrete CCII+ (AD844). Thus, it is expected that if an integrated FTFN is used, the experimental results will improve.

2.6 Comparison with OTA-Op Amp Counterparts

In this section FTFN based (MISO) filter of Fig.2.3 will be compared with a new and attractive approach of designing current mode filters using OTA-Op Amp. The operational transconductance amplifier (OTA) based universal voltage mode filters exhibit several advantages such as programmability, integrability and simplicity [46-50]. These filters do not use resistors, however, they incorporate externally connected capacitors hence called OTA-C filters. The OTA-C voltage mode filters call for using voltage buffers to provide an appropriate low impedance the output voltages. On the other hand, voltage mode active-R filters based on the Op Amp are capacitorless and cascadable realizations. An active-R circuit is suitable for monolithic integration and exhibits higher frequency performance than the conventional design [51,52]. The current mode filters based on both OTA-C and Op Amp-R are superior compared to their voltage counterparts [34,53].
Moreover, OTA-C current mode filters are cascadable because the output currents are associated with high impedances in contrast to the Op Amp-R filters. This implies that totally active filters can be achieved using only OTAs and Op Amps. Also, it is expected that current mode capacitorless-resistorless filters based on OTA-Op Amp will combine the advantages of the active-R and OTA-C filters and can avoid their disadvantages. Actually, this new approach of designing filters using OTAs and Op Amps are attractive for monolithic integration, programmability and wide frequency range of operation [35].

We proposed a new current-mode totally active (MISO) universal filter incorporating only OTAs and Op Amps [54] (see the Appendix for the complete paper). The circuit has three inputs and one high impedance output as shown in Fig.2.7. The proposed circuit can realize all the five basic filtering functions without changing the circuit topology.

Fig. 2.7: Proposed (MISO) Current Mode Universal Filter Based on OTA-Op Amp
The OTA is characterized by \( I_o = g_m(V_i - V_r) \), where \( g_m = I_{ABC}/2V_T \) is the transconductance of the OTA. \( I_{ABC} \) is the auxiliary bias-current. \( V_T \) is the thermal voltage, and \( V_i \) and \( V_r \) are input voltages of the OTA. Also, the internally compensated Op Amp open-loop gain can be expressed as \( A = \frac{B}{s} \), where \( B \) is the gain-bandwidth product. Using these relations, routine analysis results in the following current transfer function:

\[
I_o = g_m \frac{s^2 g_{m3} g_{m7} I_{in1} - sB_2 g_{m1} g_{m4} I_{in2} + B_1 B_2 g_{m1} g_{m5} I_{in1}}{s^2 g_{m3} g_{m6} g_{m7} + sB_2 g_{m2} g_{m3} g_{m4} + B_1 B_2 g_{m1} g_{m4} g_{m5}} \tag{2.30}
\]

It can be seen from (2.30) that:

1. LP response is obtained when \( I_{m2} = I_{m3} = 0 \)
2. HP response is obtained when \( I_{m1} = I_{m2} = 0 \)
3. BP response is obtained when \( I_{m1} = I_{m3} = 0 \)
4. BR response is obtained when \( I_{m2} = 0 \) and \( I_{m1} = I_{m2} \)
5. AP response is obtained when \( I_{m1} = I_{m2} = -I_{m3} \) and \( g_{m2} = g_{m4} = g_{m6} = g_{m8} \).

Also, it can be seen from (2.30) that the parameters \( \omega_0, \omega_o/Q_o \) and \( Q_o \) can be expressed as:

\[
\omega_0 = \sqrt{\frac{B_1 B_2 g_{m1} g_{m4} g_{m5}}{g_{m3} g_{m6} g_{m7}}} \tag{2.31}
\]

\[
\frac{\omega_o}{Q_o} = \frac{B_2 g_{m3} g_{m4}}{g_{m6} g_{m7}} \tag{2.32}
\]

\[
Q_o = \frac{1}{g_{m2}} \sqrt{\frac{B_1 g_{m1} g_{m5} g_{m6} g_{m7}}{B_2 g_{m3} g_{m4}}} \tag{2.33}
\]
Also, it can be seen that the LP, HP, and BP gains are as follows:

\[ G_{LP} = \frac{g_{m8}}{g_{m1}} \quad (2.34) \]

\[ G_{HP} = \frac{g_{m8}}{g_{m9}} \quad (2.35) \]

\[ G_{BP} = \frac{g_{m8}}{g_{m2}} \quad (2.36) \]

It can be seen from (2.31) and (2.32) that the center frequency (\( \omega_c \)) can be tuned via \( g_{m1} \), and/or \( g_{m5} \) without disturbing the bandwidth and the gain. Therefore, \( \omega_c \) can be electronically adjusted using the auxiliary bias currents \( I_{ABC1} \), \( I_{ABC3} \) and/or \( I_{ABC5} \). Also, it is interesting to note that linear tuning of \( \omega_c \) can be achieved when the auxiliary bias currents \( I_{ABC1} \) and \( I_{ABC5} \) are adjusted to be equal.

Furthermore, it can be seen that the bandwidth (\( \omega_c/Q_o \)) can be adjusted via \( g_{m2} \) without disturbing \( \omega_c \) and the gain. Therefore, \( \omega_c/Q_o \) can be electronically tuned using \( I_{ABC2} \). Also, it can be seen from (2.34-2.36) that the LP, HP and BP gains can be controlled by \( g_{m8} \) or electronically by the auxiliary bias current \( I_{ABC4} \) without disturbing the filter's parameters namely \( \omega_c \), \( \omega_c/Q_o \) and \( Q_o \).

It is worth mentioning that the filter parameters \( \omega_c \), \( \omega_c/Q_o \) and \( Q_o \) are temperature-insensitive even though the transconductance \( g_m \) of an OTA is temperature dependent.
Also, it can be seen from that the sensitivities of the parameters \( \omega_o \) and \( Q_o \) can be expressed as:

\[
S_{\beta}^{\omega_o} = S_{\beta}^{\omega_o} = \frac{1}{2}
\]

\[
S_{\delta_{n_1}}^{\omega_o} = S_{\delta_{m_1}}^{\omega_o} = S_{\delta_{m_1}}^{\omega_o} = -S_{\delta_{m_1}}^{\omega_o} = -S_{\delta_{m_1}}^{\omega_o} = -S_{\delta_{m_1}}^{\omega_o} = \frac{1}{2}
\]

\[
S_{\delta_{n_2}}^{\omega_o} = S_{\delta_{m_2}}^{\omega_o} = 0
\]

\[
S_{\beta}^{Q_o} = -S_{\beta}^{Q_o} = \frac{1}{2}
\]

\[
S_{\delta_{n_1}}^{Q_o} = -S_{\delta_{m_1}}^{Q_o} = -S_{\delta_{m_1}}^{Q_o} = S_{\delta_{m_1}}^{Q_o} = S_{\delta_{m_2}}^{Q_o} = S_{\delta_{m_2}}^{Q_o} = \frac{1}{2}
\]

\[
S_{\delta_{n_2}}^{Q_o} = -1
\]

\[
S_{\delta_{m_2}}^{Q_o} = 0
\]

All of which are small (not more than unity).

The proposed circuit of Fig.2.7 was simulated using the PSPICE. The Op Amp (LF156) and OTA macromodel [34] were used in simulation. The multiple output OTA was simulated using parallel-connected single output OTAs. Sample of the simulation results are shown in Fig.2.8. In order to compare the simulation results with theoretical results, it is necessary to determine the gain-bandwidth (B) product of the LF156 Op Amp since it is involved in the transfer function. Simulation result of the open loop LF156 Op Amp shows that gain-bandwidth produce (B) is equal to 4.1345 MHz.
Fig. 2.8: (a) Simulation results of the LP with: $g_{m1} = g_{m4} = g_{m5} = g_{m6} = g_{m7} = 1923 \mu S$, $g_{m2} = 2720 \mu S$.

$g_{m3} = g_{m8} = 19230 \mu S$. '—' Theoretical results, 'x' Simulation results.

Fig. 2.8: (b) Simulation results of the LP with: $g_{m1} = g_{m4} = g_{m5} = g_{m6} = g_{m7} = 1923 \mu S$, $g_{m2} = 2720 \mu S$.

$g_{m3} = g_{m8} = 30000 \mu S$. '—' Theoretical results, 'x' Simulation results.
Fig. 2.8: (c) Simulation results of the BR with: $g_{m1} = g_{m4} = g_{m5} = g_{m6} = g_{m8} = 1923 \mu S$, $g_{m2} = 3500 \mu S$.

$g_{m3} = g_{m7} = 192300 \mu S$. "—" Theoretical results, 'x' Simulation results

In summary, a new current-mode biquad universal filter has been presented. The proposed filter is totally active using only OTAs and Op Amps. The proposed filter offers the following attractive features:

(a) Current control of the filter’s gain and the parameters $\omega_o$, $\omega_o/Q_o$ and $Q_o$.

(b) Independent control of any of filter’s parameters $\omega_o$, $\omega_o/Q_o$ and gains without disturbing the others.

(c) Low sensitivities of the parameters $\omega_o$ and $Q_o$ to the gain-bandwidth products of the Op Amps and the transconductances of the OTAs.
(d) realization of the five standard filter functions namely LP, HP, BP, BR and AP without changing the circuit topology

(e) insensitivity to temperature variations.

Table 2.2 summarizes the results of a comprehensive comparison between the FTFN based realization of Fig 2.3 and the OTA-Op Amp based realization of Fig 2.7.

Table 2.2: Comparison Between (MISO) Current Mode Filters Based on FTFN and OTA-Op Amp

<table>
<thead>
<tr>
<th>Features</th>
<th>FTFN Filter</th>
<th>OTA-Op Amp Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_o$ and $\omega_o/Q_o$ Relation</td>
<td>Independent</td>
<td>Independent</td>
</tr>
<tr>
<td>Number of Active Elements</td>
<td>Four</td>
<td>Ten</td>
</tr>
<tr>
<td>Number of Passive Elements</td>
<td>Seven</td>
<td>Zero</td>
</tr>
<tr>
<td>Controlling $\omega_o$</td>
<td>Grounded Resistor</td>
<td>Electronically</td>
</tr>
<tr>
<td>Controlling $\omega_o/Q_o$</td>
<td>Capacitor</td>
<td>Electronically</td>
</tr>
<tr>
<td>Active and Passive Sensitivities</td>
<td>Low (less than unity)</td>
<td>Low (less than unity)</td>
</tr>
<tr>
<td>Realization Conditions</td>
<td>Only for AP</td>
<td>Only for AP</td>
</tr>
</tbody>
</table>

According to Table 2.2 and the previous discussion, it can be seen there are some similarities and differences between FTFN and OTA-Op Amp based filters. The main parameters $\omega_o$ and $\omega_o/Q_o$ of both filters can be adjusted independently. Also, the OTA-Op Amp based filter uses ten active elements, and the FTFN based filter uses eleven elements.
four active and seven passive. Both filters do not require any realization conditions except for AP and they enjoy low active and passive sensitivities. Also, both filters are cascadable because their output currents are associated with high impedances. On the other hand, the OTA-Op Amp based filter enjoys the advantage of electronic tuning of the filter parameters. The bandwidth of the FTFN filter can be adjusted via a capacitor which is less attractive. Also, the center frequency of the FTFN based filter can be adjusted electronically only if the center frequency controlling resistor is replaced by a JEF performing as a variable resistor. Although, the OTA-Op Amp filter is more appropriate for integration since it is capacitorless, it suffers from the frequency limitation of the Op Amp.

2.7 Universal Single Input Multiple Output (SIMO) Current Mode Filters Based on FTFN

No universal filter with single input and multiple outputs (SIMO) mixed (current and voltage) or current mode has been reported using the FTFN. In this section two such filters are proposed. First, a canonical filter using minimum number of active and passive elements is presented. However, this filter suffers from some disadvantages. The second filter overcomes these disadvantages by using more active and passive elements.
2.7.1 Universal (SIMO) Current Mode Filters (I)

The first filter uses the minimum number of active and passive components required in such application. The basic circuit consists of only two FTFNs as shown in Fig.2.9. The main advantage of the proposed circuit is that all the five filter functions can be obtained using minimum number of active and passive components (only two FTFNs plus voltage to current converter, three resistors and two capacitors).

![Diagram of Universal (SIMO) Current Mode Filters](image)

Fig.2.9: Proposed (SIMO) Current Mode Universal Filter (I)

Routine analysis of the circuit, assuming ideal FTFN characterized by $I_x=I_y=0$, $V_x=V_y$ and $I_z=I_w$, shows that the proposed circuit offers the following transfer functions:

$$I_{o1} = \frac{G_2G_3}{s^2C_1C_4 + sC_1G_1 + G_1G_3} I_{in1}$$

(2.37)

$$I_{o2} = -\frac{sC_2G_3}{s^2C_1C_4 + sC_1G_3 + G_2G_3} I_{in1}$$

(2.38)
\[ I_{o1} = -\frac{sC_1G_2}{s^2C_1C_4 + sC_1G_3 + G_2G_3} I_{i\text{n}^2} \]  
(2.39)

\[ I_{o2} = \frac{s^2C_1C_4}{s^2C_1C_4 + sC_1G_3 + G_2G_3} I_{i\text{n}^2} \]  
(2.40)

\[ I_{o3} = \frac{G_3}{s^2C_1C_4 + sC_1G_3 + G_2G_3} I_{i\text{n}^2} \]  
(2.41)

It can be seen that with input current \( I_{i\text{n}1} \), then two currents, \( I_{o1} \) (LP) and \( I_{o2} \) (BP) associated with high impedance, are offered by the circuit. Also, if the input is \( I_{i\text{n}2} \), then BP and HP currents are obtained at \( I_{o1} \) and \( I_{o2} \) respectively. Moreover, applying a voltage to current converter at \( V_i \), LP current (\( I_{o3} \)) can be obtained. Thus, \( I_{o2} \) and \( I_{o3} \) can be combined to get the BR. Also, the AP is obtainable by adding the three currents with \( G_2 = G_3 = G_5 \), where \( G_5 \) is the transconductance of the voltage to current converter. Another way of obtaining AP is by applying both inputs simultaneously (\( I_{i\text{n}} = I_{i\text{n}1} = I_{i\text{n}2} \)) and adding the two output currents (\( I_{o1} \) and \( I_{o2} \)). The result of the some of the two currents (\( I_o \)) will be equal to:

\[ I_o = \frac{s^2C_1C_4 - sC_1G_2 - sC_4G_3 + G_2G_3}{s^2C_1C_4 + sC_1G_3 + G_2G_3} I_{i\text{n}} \]  
(2.42)

Thus, AP response is achieved by satisfying this condition: \( C_1G_3 = C_1G_2 + C_4G_3 \).

Furthermore, it can be seen from (2.37-2.41) that the filter center frequency, bandwidth and quality factor are expressed respectively as:
\[\omega_c = \sqrt{\frac{G_s G_3}{C_1 C_4}} \]  
\[\omega_n = \frac{G_s}{C_2} \]  
\[Q_n = \sqrt{\frac{C_1 G_2}{C_4 G_3}} \]  

Also, it can be seen with \(I_{in} = I_{in2}\) that the LP, HP and PB gains are respectively given by:

\[G_{LP} = \frac{G_s}{G_3} \]  
\[G_{HP} = 1 \]  
\[G_{BP} = \frac{G_s}{G_3} \]  

From (2.43-2.44), it can be seen that the center frequency can be tuned without disturbing the bandwidth via \(C_1\) and/or \(R_2 = 1/G_2\). However, the bandwidth and \(Q_n\) can not be changed without disturbing the center frequency. The second main disadvantage of this circuit is the use of a floating capacitor.

Simulation results of the proposed circuit of Fig. 2.9 are shown in Fig. 2.10. The FTFN was simulated using the power supply sensing technique of Fig. 1.4. It can be seen that the circuit is performing very well as expected, except for a very slight deviation related to the Op Amp nonideal behavior such as the finite gain.
Fig 2.10:(a): Simulation results of the LP (I_{a1}) with: I_{na}=I_{na1}, C_{1}=C_{4}=500pF, R_{2}=2k\Omega.

R_{3}=1k\Omega. '—' Theoretical results, 'x' Simulation results

Fig 2.10:(b): Simulation results of the BP (I_{a2}) with: I_{na}=I_{na1}, C_{1}=45pF, C_{4}=900pF,

R_{1}=1k\Omega, R_{3}=10k\Omega. '—' Theoretical results, 'x' Simulation results
Fig 2.10: (c): Simulation results of the HP ($I_{m2}$) with: $I_{in}=I_{m2}$, $C_1=1\text{nF}$, $C_4=1\text{nF}$, $R_2=6k\Omega$.

$R_4=3k\Omega$. "—" Theoretical results, 'x' Simulation results

Fig 2.10: (d): Simulation results of the BP ($I_{e4}$) with: $I_{in}=I_{m2}$, $C_1=45\text{pF}$, $C_4=900\text{pF}$.

$R_2=1k\Omega$, $R_3=10k\Omega$. "—" Theoretical results, 'x' Simulation results
2.7.2 Universal (SIMO) Current Mode Filter (II)

In order to overcome the disadvantages of the previous circuit, a new universal current-mode second order filter using FTFNs is presented. The new circuit, offering additional new advantages, is shown in Fig.2.11. However, it requires more active and passive elements. The circuit can realize all the standard filter function simultaneously without changing the circuit topology.

Fig.2.11. Proposed (SIMO) Current Mode Universal Filter (II)

Assuming ideal FTFNs characterized by $I_x=I_y=0$, $V_x=V_y$ and $I_z=I_m$, routine analysis shows that the transimpedance transfer functions of the circuit can be expressed as:
$$V_1 = -\frac{sC_2G_1}{s^2C_2C_4G_6 + sC_4G_1G_3 + G_1G_3G_6} I_m$$  \hspace{1cm} (2.49) \\
$$V_2 = \frac{G_1G_3}{s^2C_2C_4G_6 + sC_4G_1G_3 + G_1G_3G_6} I_m$$  \hspace{1cm} (2.50) \\
$$V_3 = \frac{s^2C_2C_4}{s^2C_2C_4G_6 + sC_4G_1G_3 + G_1G_3G_6} I_m$$  \hspace{1cm} (2.51)$$

Three different voltage mode outputs can be directly obtained using voltage buffers. The five basic current mode filter transfer functions are achievable after converting these voltages to currents using voltage to current converters as follows:

$$I_{o1} = -\frac{sC_2G_4G_{o1}}{s^2C_2C_4G_6 + sC_4G_1G_3 + G_1G_3G_6} I_m$$  \hspace{1cm} (2.52) \\
$$I_{o2} = \frac{G_1G_3G_{o2}}{s^2C_2C_4G_6 + sC_4G_1G_3 + G_1G_3G_6} I_m$$  \hspace{1cm} (2.53) \\
$$I_{o3} = \frac{s^2C_2C_4G_{o3}}{s^2C_2C_4G_6 + sC_4G_1G_3 + G_1G_3G_6} I_m$$  \hspace{1cm} (2.54)

Where \( G_{o1}, G_{o2} \) and \( G_{o3} \) are the admittances of the voltage-to-current converters. The BR function is obtained by adding \( I_{o2} \) and \( I_{o3} \) and the AP is achieved by combining the three currents with \( C_4=C_5, \ G_{o1}=G_5, \ G_{o2}=G_5 \) and \( G_{o3}=G_6 \). Also, it can be seen that the filter parameters are expressed as:

$$\omega_o = \sqrt{\frac{G_1G_3G_5}{C_2C_4G_6}}$$  \hspace{1cm} (2.55) \\
$$\frac{\omega_o}{Q_o} = \frac{C_4G_5G_3}{C_2C_4G_6}$$  \hspace{1cm} (2.56)$$
\[ Q_o = \frac{1}{C_s} \sqrt{\frac{C_s C_s G_s G_o}{G_s G_s}} \]  \hspace{1cm} (2.57)

Thus, the center frequency \( (\omega_o) \) can be adjusted using \( G_s \) without disturbing the bandwidth \( \omega_o Q_o \). Also, the bandwidth can be adjusted using \( C_s \) without disturbing \( \omega_o \). Moreover, the LP, HP, and BP gains are, respectively, given by

\[ G_{lo} = \frac{G_{o2}}{G_s} \]  \hspace{1cm} (2.58)

\[ G_{hp} = \frac{G_{o3}}{G_s} \]  \hspace{1cm} (2.59)

\[ G_{pb} = \frac{C_s G_{o1}}{C_s G_s} \]  \hspace{1cm} (2.60)

Nonideal analysis is required to find the active sensitivities of the filter. Assuming that the port relations of the FTFN, can be expressed as \( I_1 = I_2 = 0, V_2 = \beta V_1 \) and \( I_{o2} = \alpha I_{o1} \) where \( \beta = 1 - \epsilon \) \( (|\epsilon| << 1) \), denotes the voltage tracking error of the FTFN and \( \alpha = 1 - \delta \) \( (|\delta| << 1) \), represents the current-tracking error, routine analysis of the circuit shown in Fig. 2.11 yields the transimpedance transfer functions:

\[ I_1' = \frac{\beta \beta_1}{\alpha_1} \frac{s C_s G_s I_{in}}{s^2 C_s C_s G_s + s \frac{\beta \beta_1 \beta_2}{\alpha \alpha_2 \alpha_3} C_s G_s G_s + \frac{\beta \beta_1 \beta_2}{\alpha \alpha_2 \alpha_3} G_i G_s G_s} \]  \hspace{1cm} (2.61)

\[ I_2' = \frac{\beta \beta_2}{\alpha \alpha_2} \frac{G_s G_s I_{in}}{s^2 C_s C_s G_s + s \frac{\beta \beta_1 \beta_2}{\alpha \alpha_2 \alpha_3} G_s G_s G_s + \frac{\beta \beta_1 \beta_2}{\alpha \alpha_2 \alpha_3} G_s G_s G_s} \]  \hspace{1cm} (2.62)
\[ V_r = \frac{s^2 C_x C_m I_m}{s^2 C_x C_G + s \cdot \frac{\beta_1 \beta_2 \beta_3}{\alpha_1 \alpha_2 \alpha_3} C_G G_G G_G + \frac{\beta_1 \beta_2 \beta_3}{\alpha_1 \alpha_2 \alpha_3} G_G G_G} \]  

(2.63)

The main parameters of the filters are given by the following equations:

\[ \omega_0 = \frac{\beta_1 \beta_2 \beta_3 G_G G_G}{\sqrt{\alpha_1 \alpha_2 \alpha_3 C_x C_G}} \]  

(2.64)

\[ \frac{\omega}{Q} = \frac{\beta_1 \beta_2 \beta_3 C_G G_G}{\alpha_1 \alpha_2 \alpha_3 C_x C_G} \]  

(2.65)

\[ Q_0 = \frac{1}{C_G} \frac{\sqrt{\alpha_1 \alpha_2 \alpha_3 C_x C_G G_G}}{\beta_1 \beta_2 \beta_3 G_G} \]  

(2.66)

From (2.64) and (2.66) it is easy to show that the active and passive sensitivities of the parameters \( \omega_0 \) and \( Q_0 \) are:

\[ S_{\omega_0}^{a_1} = S_{\omega_0}^{a_2} = S_{\omega_0}^{a_3} = -S_{\omega_0}^{a_1} = -S_{\omega_0}^{a_2} = -S_{\omega_0}^{a_3} = \frac{1}{2} \]

\[ S_{\omega_0}^{c_1} = S_{\omega_0}^{c_2} = S_{\omega_0}^{c_3} = -S_{\omega_0}^{c_1} = -S_{\omega_0}^{c_2} = -S_{\omega_0}^{c_3} = \frac{1}{2}, \quad S_{\omega_0}^{c_4} = 0 \]

\[ S_{a_1}^{q_1} = S_{a_2}^{q_1} = S_{a_3}^{q_1} = -S_{a_1}^{q_1} = -S_{a_2}^{q_1} = -S_{a_3}^{q_1} = \frac{1}{2} \]

\[ S_{a_4}^{q_1} = S_{a_5}^{q_1} = S_{a_6}^{q_1} = -S_{a_4}^{q_1} = -S_{a_5}^{q_1} = -S_{a_6}^{q_1} = \frac{1}{2} \]

\[ S_{c_1}^{q_1} = S_{c_2}^{q_1} = S_{c_3}^{q_1} = -S_{c_1}^{q_1} = -S_{c_2}^{q_1} = -S_{c_3}^{q_1} = \frac{1}{2} \]

\[ S_{c_4}^{q_1} = -1 \]

All of which are small (less than or equal to unity)
In summary, a novel (SIMO) current mode filter based on FTFN and using grounded passive elements is proposed. At most six FTFNs and ten grounded passive elements are required to obtain any type of five filter functions. Also, it can be shown that to obtain all the five filter functions simultaneously, eleven FTFNs and fifteen passive elements are required. The proposed circuit exhibits the following advantages:

(a) Arbitrary biquadratic transfer functions are realized with single input.
(b) All the five basic filter functions can be obtained simultaneously.
(c) Independent control of filter parameters namely the center frequency and the bandwidth using grounded elements.
(d) All the passive elements are grounded.
(e) Can be easily converted into voltage mode biquad.
(f) Low active and passive sensitivities.

The proposed circuit was used to realize LP, HP, BP, BR and AP for verifying the theoretical analysis. The experimental results obtained agree very well with the presented theory. The FTFN was implemented using the two current conveyors (AD844) realization of Fig. 18(b). Sample of the experimental results of the LP, HP and BP are presented in Fig. 2. The slight deviation of the experimental results from the theoretical is mainly due to the stray capacitances and the nonideal practical performance of the current conveyors used to realize the FTFN.
Fig. 2.12: (a): Experimental results of the LP with: $R_1 = R_3 = R_5 = R_6 = 5.1k\Omega$, $C_2 = C_4 = C_5 = 1.2nF$

'—' Theoretical results, '••' Experimental results

Fig. 2.12: (b): Experimental results of the HP with: $R_1 = R_3 = R_5 = R_6 = 5.1k\Omega$, $C_2 = C_4 = C_5 = 1.2nF$

'—' Theoretical results, '••' Experimental results
Fig.2.12: (c): Experimental results of the BP with: $R_1=R_2=R_6=5.1k\Omega$, $R_5=2K$, $C_2=C_4=1.2nF$.

$C_v=200pF$. "--" Theoretical results. "*" Experimental results

2.8 Comparison with CCII+ Counterparts

No single FTFN based (SIMO) filter has been suggested in the literature. Thus, the proposed filters will be compared with CCs counterparts. As discussed before, all of the proposed filters require more than one type of CCs. However, a fair comparison should be between the FTFN based filters and those built around only one type of CCs. As mentioned previously the CCII+ based filters are selected for comparison because the
CCII+ can not be obtained from FTFN and it is proved to be very powerful and well established. A new versatile active biquad based only on CCII+, which can not be obtained from the FTFN, is introduced. The proposed configuration is shown in Fig.2.13.

Fig. 2.13: Proposed (SIMO) Current Mode Universal Filter Based on CCII+

Assuming non-ideal CCII+ characterized by \( i_x = \alpha v_x, v_x = \beta v_y \), where \( \alpha = 1 - \varepsilon_1 \) and \( \varepsilon_1 \) denotes the current-tracking error, \( \beta = 1 - \varepsilon_2 \) and \( \varepsilon_2 \) represents the voltage-tracing error, routine analysis of the circuit of Fig.2.13 results in the following transimpedance transfer functions:
\[ I'_1 = -\frac{\alpha_4 \beta_3 G_1 G_2}{\alpha_4 \alpha_2 \beta_3 \beta_3} s^2 C_2 C_1 G_1 + s \frac{C_2 G_4 G_5}{\alpha_4 \alpha_2 \beta_3 \beta_3} + \frac{\alpha_4 \beta_3 \beta_3 G_1 G_2}{\alpha_4 \alpha_2 \beta_3 \beta_3} I_{in} \quad (2.66) \]

\[ I'_2 = \frac{\alpha_4 \beta_3 s^2 C_2 C_1 G_1 + \alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3 G_2 G_1}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} I_{in} \quad (2.67) \]

\[ I'_3 = \frac{\alpha_4 \beta_3 s^2 C_2 C_1 G_1}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} s^2 C_2 C_1 G_1 + s \frac{C_2 G_4 G_5}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} + \frac{\alpha_4 \beta_3 \beta_3 G_1 G_2}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} I_{in} \quad (2.68) \]

From the above relations, it can be seen that three different voltage outputs can be obtained by adding proper voltage buffers. Also, it can be seen that these transimpedance transfer functions can be converted to current mode relations by adding voltage to current converters. This provides the following current mode LP, BR and BP biquad functions respectively:

\[ I_{e_1} = -\frac{\alpha_4 \beta_3 G_4}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} s^2 C_2 C_1 G_1 + s \frac{C_2 G_4 G_5}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} + \frac{\alpha_4 \beta_3 \beta_3 G_1 G_2}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} I_{in} \quad (2.69) \]

\[ I_{e_2} = \frac{\alpha_4 \beta_3 s^2 C_2 C_1 G_1 + \alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3 G_2 G_1}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} I_{in} \quad (2.70) \]

\[ I_{e_3} = -\frac{\alpha_4 \beta_3 G_4}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} s^2 C_2 C_1 G_1 + s \frac{C_2 G_4 G_5}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} + \frac{\alpha_4 \beta_3 \beta_3 G_1 G_2}{\alpha_4 \alpha_2 \alpha_3 \beta_3 \beta_3} I_{in} \quad (2.71) \]
Where \( G_{o1}, G_{o2} \) and \( G_{o3} \) are the admittances of the voltage-to-current converters. The AP is obtained by combining \( L_{o2} \) and \( L_{o3} \) with \( G_{o2} = G_{c} \) and \( G_{o3} = G_{c} \) and the HP is obtained by adding \( L_{o1} \) and \( L_{o2} \) with \( G_{d} = G_{c} \) and \( G_{o1} = G_{o2} \). Thus, at most seven CCII- and eight passive elements are required to obtain any type of the five filters. Moreover, it can be noticed that twelve CCII- and fourteen passive elements are required to offer all the five filter functions simultaneously.

Also, the filter parameters obtained from the above transfer functions are given by:

\[
\omega = \frac{\alpha_{2} \alpha_{3} \beta_{2} \beta_{3} G_{1} G_{3}}{\alpha_{1} \beta_{2} C_{2} C_{1}} \tag{2.72}
\]

\[
\frac{\omega}{Q} = \frac{G_{2} G_{o}}{\alpha_{2} \alpha_{3} \alpha_{4} \beta_{2} \beta_{3} C_{2} C_{1} G_{1}} \tag{2.73}
\]

\[
Q_{o} = \frac{\alpha_{2} \alpha_{3} \beta_{2} G_{3}}{G_{2} G_{o}} \times \frac{\sqrt{\alpha_{1} \alpha_{2} \beta_{1} \beta_{2} C_{2} G_{1} G_{3}}}{C_{2}} \tag{2.74}
\]

It can be seen that the center frequency can be adjusted using \( G_{1}, G_{3} \) and/or \( C_{2} \) without affecting the bandwidth \( \omega_{o}/Q_{o} \). Similarly, the bandwidth can be tuned without disturbing \( \omega_{o} \) via \( G_{d}, G_{o} \) and/or \( G_{5} \).

The active and passive sensitivities of filter parameters \( \omega_{o} \) and \( Q_{o} \) are given by:
\[ S_{\alpha_2} = S_{\beta_2} = -S_{\alpha_2} = S_{\alpha_2} = S_{\alpha_2} = -S_{\alpha_2} = \frac{1}{2} \]

\[ S_{\alpha_3} = S_{\beta_3} = S_{\alpha_3} = S_{\alpha_3} = 0 \]

\[ S_{\alpha_4} = S_{\beta_4} = -S_{\alpha_4} = -S_{\alpha_4} = \frac{1}{2} \]

\[ S_{\alpha_5} = S_{\beta_5} = S_{\alpha_5} = S_{\alpha_5} = 0 \]

\[ S_{\alpha_6} = S_{\beta_6} = S_{\alpha_6} = S_{\beta_6} = S_{\alpha_6} = \frac{1}{2} \]

\[ S_{\alpha_7} = S_{\beta_7} = \frac{1}{2} \]

\[ S_{\alpha_8} = S_{\beta_8} = S_{\alpha_8} = S_{\beta_8} = \frac{1}{2} \]

\[ S_{\alpha_9} = S_{\beta_9} = S_{\alpha_9} = S_{\beta_9} = 1 \]

\[ S_{\alpha_{10}} = S_{\beta_{10}} = S_{\alpha_{10}} = S_{\beta_{10}} = \frac{1}{2} \]

\[ S_{\alpha_{11}} = S_{\beta_{11}} = S_{\alpha_{11}} = S_{\beta_{11}} = 1 \]

\[ S_{\alpha_{12}} = S_{\beta_{12}} = S_{\alpha_{12}} = S_{\beta_{12}} = 0 \]

All of which are small (no more than or equal to unity)

In summary, the proposed circuit enjoys the following advantages:

(a) Arbitrary biquadratic transfer functions are realized with single input.

(b) With multiple outputs all the five basic filter functions can be obtained simultaneously.

(c) Grounded resistors control the filter parameters.

(d) All the passive elements are grounded.

(e) Easy conversion into voltage mode biquad.

(f) Suitable for IC integration.
The proposed filter was tested practically using CCII+ (AD844) and discrete passive components. Sample of the experimental results are shown in Fig. 2.14. It can be seen from Fig. 2.14 that the experimental results are in good agreement with the theoretical analysis except for small deviation due to the stray capacitances, the parasitic capacitances of CCII− (AD844) and the nonideal performance of the practical CCII−.

![Graph showing experimental results](image)

Fig. 2.14: (a) Experimental results of the LP Filter with: $C_2 = C_3 = 470\mu F$, $R_1 = 4k\Omega$, $R_3 = R_4 = 10k\Omega$, $R_5 = R_6 = 2k\Omega$. — Theoretical results, ∗ Experimental results.
Fig. 2.14: (b) Experimental results of the HP Filter with: $C_2 = C_1 = 470 \text{pF}$, $R_1 = 4 \text{k}\Omega$, $R_3 = R_4 = 10 \text{k}\Omega$.

$R_e = R_c = R_s = 2 \text{k}\Omega$. '—' Theoretical results, '*' Experimental results

Fig. 2.14: (c) Experimental results of the BR with: $C_2 = C_1 = 470 \text{pF}$, $R_1 = 4 \text{k}\Omega$, $R_3 = R_4 = 10 \text{k}\Omega$.

$R_e = R_c = R_s = 2 \text{k}\Omega$. '—' Theoretical results, '*' Experimental results.
The main features of both the FTFN universal (SIMO) current mode filter of Fig 2.11 and the CCII− counterparts of Fig 2.13 are summarized in Table 2.4.

Table 2.3: Comparison Between (SIMO) Current Mode Filters Based on FTFN and CCII+

<table>
<thead>
<tr>
<th>Features</th>
<th>FTFN Universal Filter</th>
<th>CCII− Universal Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_o$ and $\omega_o/Q_o$ Relation</td>
<td>Independent</td>
<td>Independent</td>
</tr>
<tr>
<td>Number of Elements to Obtain any Filter Function</td>
<td>Six FTFNs, Ten Passive</td>
<td>Seven CCII−, Eight Passive</td>
</tr>
<tr>
<td>Number of Elements to Obtain All the Filter Types</td>
<td>Eleven FTFNs, Fifteen Passive</td>
<td>Twelve CCII−, Fourteen Passive</td>
</tr>
<tr>
<td>Capacitors and Resistors</td>
<td>Grounded</td>
<td>Grounded</td>
</tr>
<tr>
<td>Controlling $\omega_o$</td>
<td>Resistor</td>
<td>Resistor</td>
</tr>
<tr>
<td>Controlling $\omega_o/Q_o$</td>
<td>Capacitor</td>
<td>Resistor</td>
</tr>
<tr>
<td>Active and Passive Sensitivity</td>
<td>Low (less than unity)</td>
<td>Low (less than unity)</td>
</tr>
<tr>
<td>Realization Conditions</td>
<td>Only for All-pass</td>
<td>For All-pass and HP</td>
</tr>
<tr>
<td>Accuracy</td>
<td>Good</td>
<td>Good</td>
</tr>
</tbody>
</table>

It is clear from the above Table that the FTFN and CCII+ based (SIMO) filters exhibit almost the same advantages. Although the CCII+ based filter bandwidth can be controlled by a resister and that of the FTFN based filter by a capacitor, the FTFN filter incorporates less number of active elements. Moreover, the experimental results of both FTFN and CCII− based filters agree with the theoretical analysis very well even that the FTFN was
implemented using discrete CCII+ (AD844). So, it is expected that if a real FTFN is used, the practical results will improve. Also, it can be noticed that extensive number of active and passive elements are needed for this type of filter. It is worth mentioning that incorporating an additional output current to CCII or the FTFN would reduce the number of active and passive elements required in such applications tremendously.

2.9 Universal (MISO) Voltage Mode Filters Using FTFN

A new FTFN based universal voltage mode active biquadratic filter configuration is proposed. The circuit, with three inputs and one output, is shown in Fig. 2.15.

![Fig 2.15: Proposed (MISO) Voltage Mode Universal Filter](image)

Using the ideal FTFN port relations characterized by \(I_x=I_y=0\), \(V_x=V_y\) and \(I_z=I_w\), routine analysis of the filter circuit of Fig. 2.15 produces the following voltage transfer function:
\[
I_o = \frac{s^2C_1^2I_+^2 + sC_2G_1I_+ I_- + G_1G_2I_-^2}{s^2C_1^2 + sG_1(C_1 + C_2) + G_1G_2},
\]

(2.75)

Also, an additional output current \( I_o \) is provided. This output current can be expressed as

\[
I_o = sC_1(I_+ - I_-)
\]

(2.76)

It can be seen from (2.75) and (2.76) that:

1) \( V_o \) is LP and \( I_o \) is BP when \( V_1 = V_3 = 0 \)

2) \( V_o \) is BP and \( I_o \) is HP when \( V_1 = V_3 \)

3) \( V_o \) is HP when \( V_2 = V_3 = 0 \)

4) \( V_o \) is notch and \( I_o \) is HP when \( V_1 = V_3 \) and \( V_2 = 0 \)

5) \( V_o \) is AP and \( I_o \) is HP when \( V_1 = -V_2 = -V_3 \) and \( C_1 = 0 \)

The circuit incorporates the minimum number of active and passive components with no requirement for matching components. This circuit is attractive for realizing LP since the circuit, in this case, comprises grounded capacitors. Also, it can be found from (2.75) that the center frequency and the bandwidth are given by the following relations respectively

\[
\omega_c = \frac{G_1G_2}{\sqrt{C_2C_4}}
\]

(2.77)

\[
\frac{\omega_n}{Q_o} = \frac{(C_1 + C_2)G_3}{C_2C_4}
\]

(2.78)

\[
Q_o = \frac{\sqrt{C_2C_4G_1}}{\sqrt{G_3(C_1 + C_4)}}
\]

(2.79)
It can be seen from (2.77) and (2.78) that the parameter \( \omega_n \) can be adjusted by controlling \( R_1 = 1/G_1 \) without disturbing the parameter \( \omega_c/Q_0 \). Similarly, the parameter \( \omega_n, Q_0 \) can be tuned without varying the center frequency \( \omega_c \) via \( C_1 \). Thus, the HP and BP realizations enjoy grounded control of the filter parameters \( \omega_n \) and \( \omega_c/Q_0 \). Moreover, it can be seen from (2.75) that the LP, HP and BP gains can be expressed as

\[
G_{LP} = 1 \quad (2.80)
\]

\[
G_{HP} = 1 \quad (2.81)
\]

\[
G_{BP} = \frac{C_2 G_3}{(C_1 + C_4)G_3} \quad (2.82)
\]

Taking into account the nonideal behavior of the FTFN, the port relations of the FTFN considering the voltage and the current tracking errors, are expressed as: \( I_1 = I_2 = 0, V_2 = \beta V_1 \) and \( I_o = \alpha I_o \) where \( \beta = 1 - \varepsilon \) (\( |\varepsilon| < < 1 \)), denotes the voltage tracking error of the FTFN and \( \alpha = 1 - \delta \) (\( |\delta| < < 1 \)), represents the current-tracking error. Reanalysis of the circuit yields the following transfer functions:

\[
I_o^* = \frac{s^2 \beta C_2 C_1 V_1^* + sC_2 G_1 V_1^* + \alpha G_1 G_2 V_1^*}{s^2 \beta C_2 C_2 + s(\alpha C_1 G_1 + \beta C_2 G_3) + \alpha G_1 G_3} \quad (2.83)
\]

\[
I_o = \alpha_s C_2 (V_1^* - \beta I_o^*) \quad (2.84)
\]

From (2.83), the center frequency, bandwidth, and the equality factor can be, respectively, expressed as:
\[ \omega_\alpha = \frac{\alpha \beta C_1 G_1}{\sqrt{\beta_1 \beta_2 C_1 C_2}} \]  \quad (2.85)

\[ \frac{\omega_\alpha}{Q_\alpha} = \frac{(\alpha \beta C_1 + \beta_1 \beta_2 C_2)G_1}{\beta_1 \beta_2 C_1 C_2} \]  \quad (2.86)

\[ Q_\alpha = \frac{\sqrt{\alpha \beta_1 \beta_2 C_1 C_2 G_1}}{\sqrt{G_1 (C_1 + \alpha \beta_1 \beta_2 C_2)}} \]  \quad (2.87)

Thus, it is easy to show that the passive and active sensitivities of the parameters \( \omega_\alpha \) and \( Q_\alpha \) can be found to be:

\[ S_{\alpha_\alpha}^{\omega_\alpha} = S_{\alpha_\alpha}^{Q_\alpha} = -S_{\alpha_\alpha}^\omega = -S_{\alpha_\alpha}^{Q_\alpha} = \frac{1}{2} \]

\[ S_{\alpha_1}^{\omega_\alpha} = S_{\alpha_1}^{Q_\alpha} = S_{\alpha_1}^\omega = -S_{\alpha_1}^{Q_\alpha} = 0 \]

\[ S_{C_1}^{\alpha_\alpha} = S_{C_1}^{Q_\alpha} = -S_{C_1}^{\alpha} = \frac{1}{2} \]

\[ S_{\alpha_1}^{Q_\alpha} = S_{\alpha_1}^{Q_\alpha} = S_{\alpha_1}^{Q_\alpha} = \frac{1}{2} \frac{\alpha \beta_1 \beta_2 C_2}{C_1 + \alpha \beta_1 \beta_2 C_2} \]

\[ S_{C_1}^{Q_\alpha} = -\frac{C_1}{C_1 + \alpha \beta_1 \beta_2 C_2} \]

Therefore, it is clear that the active and passive sensitivities are small (no more than unity).

In summary, the proposed circuit exhibits independent control of the center frequency and the bandwidth in contrast to the previously known FTFN based filters [20]. Also, the proposed circuit offers the following advantageous features:
(a) The realization of all the five generic filter functions from the same configuration without requiring matching conditions

(b) Use of two FTFNs with minimum number of passive elements (five) required for independent control of the filter parameters

(c) Bandwidth control through a grounded capacitor

(d) Additional current mode filter functions are offered

Simulation results confirming the theoretical analysis are shown in Fig 2.16. The FTFNs are simulated using the current sensing technique of Fig 1.4. The simulation results are in excellent agreement with the theoretical analysis.

Fig 2.16: (a): Simulation results of the LP with: \( R_1=R_2=1\, \text{k}\Omega, C_1=500\, \text{pF}, C_2=1\, \text{nF}, C_3=500\, \text{pF} \)

'-' Theoretical results, 'x' Simulation results
Fig 2.16: (b): Simulation results of the HP with: \( R_1 = R_3 = 1 \text{k}\Omega \), \( C_1 = 500 \text{pF} \), \( C_2 = 1 \text{nF} \), \( C_3 = 500 \text{pF} \)

'—' Theoretical results, 'x' Simulation results

Fig 2.16: (c): Simulation results of the BP with: \( R_1 = 1 \text{k}\Omega \), \( R_2 = 30 \text{k}\Omega \), \( C_1 = 100 \text{pF} \), \( C_2 = 4 \text{nF} \), \( C_3 = 1 \text{nF} \).

'—' Theoretical results, 'x' Simulation results
Fig 2.16: (d): Simulation results of the BR with: $R_1=1k\Omega$, $R_2=30k\Omega$, $C_1=100\text{pF}$, $C_2=C_3=1\text{nF}$

'-' Theoretical results, 'x' Simulation results

Fig 2.16: (e): Simulation results of the AP Magnitude Response with: $R_1=R_3=1k\Omega$, $C_1=0$, $C_2=1\text{nF}$, $C_4=500\text{pF}$. '-' Theoretical results, 'x' Simulation results
Fig. 2.16: (f): Simulation results of the AP Phase response with: R_1=R_2=1kΩ, C_1=0, C_2=1nF.

C_4=500pF. ‘—’ Theoretical ‘x’ Simulation results

In addition to the voltage mode filters, a cascadable current mode BP filter can be obtained from the same proposed circuit as shown in Fig. 2.17. Using ideal port relations for the FTFN, routine analysis yields the following current mode band-pass transfer function:

$$I_o = -\frac{sC_2G_3}{s^2C_2C_4 + s(C_1G_2 + C_4G_3) + G_1G_3} I_{in}$$

(2.88)

Simulation result of the BP current filter is shown in Fig. 2.18. A slight deviation between the theoretical analysis and the simulation is mainly due to the nonideal characteristics of the Op Amp used to realize FTFN.
Fig. 2.17: Additional Current Filter Obtained From the Circuit of Fig. 2.15

Fig. 2.18: Current-Mode BP Filter. 

R1=1kΩ, R2=20kΩ, C1= C4= 100pF, C2=2nF,

'--' Theoretical 'x' Simulation results
2.10 Comparison with CCII+ Counterparts

Recently, a universal filter with three inputs and single output using CCII+ was proposed [55]. The filter circuit incorporates three CCII+, three floating resistors and two capacitors one of them is grounded. This filter exhibits low output impedance and low passive sensitivities. However, the filter requires matching resistors to realize HP, BR and the AP. The main features of this filter and those of the FTFN based filter of Fig. 2.15 are summarized in Table 2.4.

<table>
<thead>
<tr>
<th>Features</th>
<th>FTFN Universal Filter</th>
<th>CCII+ Universal Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \omega_o ) and ( \omega_o/Q_o ) Relation</td>
<td>Independent</td>
<td>Independent</td>
</tr>
<tr>
<td>Number of Active Elements</td>
<td>Two</td>
<td>Three</td>
</tr>
<tr>
<td>Number of Passive Elements</td>
<td>Five</td>
<td>Five</td>
</tr>
<tr>
<td>Controlling ( \omega_o )</td>
<td>Floating resistor</td>
<td>Grounded capacitor</td>
</tr>
<tr>
<td>Controlling ( \omega_o/Q_o )</td>
<td>Grounded capacitor</td>
<td>Floating resistor</td>
</tr>
<tr>
<td>Active Sensitivity</td>
<td>Low (less than unity)</td>
<td>Not clear</td>
</tr>
<tr>
<td>Passive Sensitivity</td>
<td>Low (less than unity)</td>
<td>Low (less than unity)</td>
</tr>
<tr>
<td>Realization Conditions</td>
<td>Cancellation/ for AP</td>
<td>Matching/ for HP, BR and AP</td>
</tr>
<tr>
<td>Additional outputs</td>
<td>Four</td>
<td>Non</td>
</tr>
</tbody>
</table>
It is clear from Table 2.4 that the FTFN based filter offers some advantages over the CCII+ based filter. First, it uses less number of active elements, only two, while CCII+ based filter uses three. Second, the FTFN based filter requires matching conditions for AP only, however, the CCII+ based filter relies on matching conditions to realize HP, BR and AP. Also, FTFN based filter can provide additional outputs in addition to the basic output whereas CCII+ circuit can not. Moreover, the active and passive sensitivities of the FTFN filter are low but the CCII+ based filter active sensitivities are not clear. The CCII+ filter is associated with low output impedance allowing it to be cascaded successfully. Also, the FTFN filter can be made cascadable by using a voltage buffer.

2.11 Comparison with CFA Counterparts

As mentioned previously, the proposed voltage mode filter of Fig.2.15 enjoys independent control of the center frequency and the bandwidth in contrast to the filter of [20]. Also, the proposed filter appears to be more efficient than CCII+ counterparts. However, one may ask what is the advantages of using the FTFN over other devices-based filters in such application. In other words, how is the FTFN-based filters compared with other devices counterparts. In fact, the CFA proved to be very powerful in designing universal voltage mode filters compared with Op Amp, OTA and CCII [56-62]. The current feedback amplifier CFA is no more than a current conveyor associated with an output buffer. This output buffer provices a low impedance output voltage. Also, it gives the designers more
freedom in their work for designing new universal filters. Thus, the proposed voltage mode universal filter based on FTFN will be compared with CFA counterparts. Recently, several voltage mode filters based on the CFA have been proposed. The filters suggested in [56-60] are not universal filters. A (MISO) universal filter using two CFAs and six passive elements was proposed by Liu.[61]. This filter is capable of realizing any of the five filter functions without changing the circuit topology. Another (MISO) universal filter, which reduces the number of passive components to five, was reported in [62].

A new minimum component universal voltage mode filter using two CFAs, two capacitors and three resistors is presented here. The proposed circuit exhibits all the advantages of the circuit of [62] and can offer additional outputs simultaneously with the main outputs. The proposed circuit is shown in Fig.2.19.

![Diagram](image-url)

**Fig.2.19: Proposed (MISO) Voltage Mode Universal Filter Based on CFA**
Considering the nonideal characteristics of the CFA, namely \( i_z = \alpha i_z, v_z = \beta v_z, v_n = \gamma v_z \), where \( \alpha = 1 - \varepsilon_i \) and \( |\varepsilon_i| << 1 \) denotes the current-tracking error, \( \beta = 1 - \varepsilon_v \) and \( |\varepsilon_v| << 1 \) represents the input voltage-tracking error, \( \gamma = 1 - \varepsilon_o \) and \( |\varepsilon_o| << 1 \) denotes the output voltage tracking error, routine analysis shows that the transfer function of the circuit of Fig 2.19 can be expressed as:

\[
V_o = \frac{s^2C_2C_3V_i - \alpha_2\gamma_1sC_2G_5V_4 + \alpha_1\alpha_2\gamma_1G_1G_3V_i + \alpha_2\beta_2sC_2(G_1 + G_2)V_i}{s^2C_2C_3 + \alpha_2\gamma_2sC_2G_4 + \alpha_1\alpha_2\beta_2\gamma_2G_2G_4}
\]  
(2.89)

\[
V_{o1} = \gamma_1V_4 - \alpha_1\gamma_1 \frac{G_1}{sC_2} (V_1 - \beta_2\gamma_2V_o)
\]  
(2.90)

From (2.89) and (2.90) it can be seen that

1) \( V_o \) is LP when \( V_2 = V_3 = V_4 = 0 \)
2) \( V_o \) is BP and \( V_{o1} \) is LP when \( V_1 = V_3 = V_4 = 0 \)
3) \( V_o \) is HP and \( V_{o1} \) is BP when \( V_1 = V_2 = V_4 = 0 \)
4) \( V_o \) is notch when \( V_1 = V_3 \) and \( V_2 = V_4 = 0 \)
5) \( V_o \) is AP when \( V_1 = V_3 = V_4 \) and \( V_2 = 0 \) and \( G_4 = G_5 \)

Also, from the (2.89) the parameters \( \omega_o, \omega_o/Q_o \) and \( Q_o \) can be expressed as:

\[
\omega_o = \sqrt{\frac{\alpha_2\beta_2\gamma_2G_1G_4}{C_2C_3}}
\]  
(2.91)
\[ \frac{\omega_o}{Q_o} = \frac{\alpha \gamma G_z}{C_z}, \]  
(2.92)

\[ Q_o = \frac{1}{G_z} \sqrt{\frac{\alpha \beta \gamma C_z G_z G_z}{\alpha \beta \gamma C_z}}, \]  
(2.93)

Moreover, it can be seen from (2.89) that the LP, HP and BP gains can be approximately expressed as:

\[ G_{LP} = 1 \]  
(2.94)

\[ G_{HP} = 1 \]  
(2.95)

\[ G_{BP} \approx \frac{G_z + G_z}{G_z} \]  
(2.96)

It is easy to show that the passive and active sensitivities of the parameters \( \omega_o \) and \( \omega_o/Q_o \) can be expressed as:

\[ S_{\alpha}^{\omega_o} = S_{\beta}^{\omega_o} = -S_{\gamma}^{\omega_o} = -S_{C_z}^{\omega_o} = \frac{1}{2} \]

\[ S_{\alpha_z}^{\omega_o} = S_{\alpha_z}^{\omega_o} = S_{\beta_z}^{\omega_o} = S_{\gamma_z}^{\omega_o} = S_{C_z}^{\omega_o} = \frac{1}{2} \]

\[ S_{\alpha}^{\omega_o} = S_{\beta}^{\omega_o} = S_{\gamma}^{\omega_o} = -S_{C_z}^{\omega_o} = \frac{1}{2} \]

\[ S_{\alpha_z}^{\omega_o} = -1 \]

\[ S_{\alpha_z}^{\omega_o} = S_{\beta_z}^{\omega_o} = S_{\gamma_z}^{\omega_o} = -S_{\alpha_z}^{\omega_o} = -S_{\beta_z}^{\omega_o} = \frac{1}{2} \]

Thus all the active and passive sensitivities are no more than unity.
In summary, the proposed circuit enjoys the following advantages:

(a) Realizing all the five filter functions without changing the circuit topology

(b) Offering an additional function simultaneously with some of the main output filter

(c) Independent control of the center frequency and the bandwidth.

(d) Low active and passive sensitivities

(e) Easily cascadable since the circuit exhibits low output impedance.

The proposed circuit was tested experimentally results using AD844 CFA (current conveyor and a voltage buffer). The experimental results of LP, HP and BR are shown in Fig.2.20. The experimental results are in good agreement with the theory presented.

Fig 2.20: (a): Experimental results of the LP with: \( R_l=1k\Omega \), \( R_c=R_e=2k\Omega \), \( C_2=C_3=100\mu F \).

'- ' Theoretical results, '* ' Experimental results
Fig. 2.20: (b): Experimental results of the HP with: $R_t=1\,k\Omega$, $R_e=R_c=2\,k\Omega$, $C_2=C_3=100\,pF$.

'--' Theoretical, '*-' Experimental results

Fig. 2.20: (c): Experimental results of the BR with: $R_t=1\,k\Omega$, $R_e=10\,k\Omega$, $R_c=2\,k\Omega$, $C_2=C_3=100\,pF$.

'--' Theoretical, '*-' Experimental results
A comprehensive comparison between the FTFN universal (MISO) voltage mode filter of Fig. 2.17 with the counterparts of the CFA of Fig. 2.19 is summarized in Table 2.5.

<table>
<thead>
<tr>
<th>Features</th>
<th>FTFN Universal Filter</th>
<th>CCII- Universal Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\omega_o$ and $\omega_o/Q_o$ Relation</td>
<td>Independent</td>
<td>Independent</td>
</tr>
<tr>
<td>Number of Active Elements</td>
<td>Two</td>
<td>Two</td>
</tr>
<tr>
<td>Number of Passive Elements</td>
<td>Five</td>
<td>Five</td>
</tr>
<tr>
<td>Controlling $\omega_o$</td>
<td>Resistor</td>
<td>Resistor</td>
</tr>
<tr>
<td>Controlling $\omega_o/Q_o$</td>
<td>Capacitor</td>
<td>Resistor</td>
</tr>
<tr>
<td>Active and Passive Sensitivity</td>
<td>Low (less than unity)</td>
<td>Low (less than unity)</td>
</tr>
<tr>
<td>Realization Conditions</td>
<td>Cancellation restriction</td>
<td>Matching condition</td>
</tr>
<tr>
<td></td>
<td>Only for All-pass</td>
<td>Only for All-pass</td>
</tr>
<tr>
<td>Additional outputs</td>
<td>Four</td>
<td>Two</td>
</tr>
<tr>
<td>Accuracy</td>
<td>Good</td>
<td>Good</td>
</tr>
</tbody>
</table>

From the above Table, it is obvious that both devices appear to be very effective in designing voltage mode (MISO) universal filters. The CFA voltage mode filter exhibits two advantages over that of the FTFN. The first is that controlling the bandwidth of CFA based filter is via a resistor and that of the FTFN is via a capacitor. The second advantage is that the output of the CFA filter output is cascadable because it is associated with low impedance in contrast to the FTFN filter. However, the FTFN based filter can be easily
made cascadable by using a voltage buffer. On the other hand, the FTFN based filter requires only a cancellation restriction for realizing the AP while the CFA based filter relies on a matching condition. Also, the FTFN based filter offers more additional outputs than the CFA based filter.

2.12 Summary

Several biquad universal filters based on the FTFN have been proposed. First, a (MISO) current mode filter using only three FTFNs has been discussed. That filter suffers from some disadvantages: requiring matching/cancellation conditions for the LP, BR and AP. Also, these conditions deprive the independent control of the center frequency and the bandwidth of those filter functions. Another (MISO) filter has been suggested using four FTFNs and grounded passive elements. This filter configuration enjoys independent control of the center frequency and the bandwidth and requires no passive components matching except for the AP. Moreover, this filter has been compared with a new CCII+ based counterpart. It was found that the two filters almost enjoy the same features although the FTFN filter uses less passive elements.

Furthermore, the FTFN realization was compared with a totally active filter based on OTAs and Op Amps. The OTA-Op Amp filter is more suitable for integration, however,
its frequency performance suffers from the Op Amp limited bandwidth and slew rate. Also, the OTA-Op Amp based filter parameters can be electronically tuned. This advantage can be also offered by the FTFN by designing filter with grounded resistor control of the filter parameters. Then, these resistors can be replaced by JFETs to adjust the filter parameters electronically. The center frequency of the proposed filter based on FTFN can be made electronically tunable but it is difficult to make the bandwidth too. Thus it can be concluded that the FTFN is very strong candidate in designing MISO current mode universal filter.

Secondly, two new (SIMO) filters have been proposed. The first one incorporates the minimum number of active elements (three) in such application. However, it suffers from some disadvantages: the bandwidth can not be adjusted without disturbing the center frequency and a floating capacitor is required. Another new FTFN based filter that comprises six FTFNs and grounded passive elements was proposed. This configuration exhibits independent control of the center frequency and the bandwidth. Furthermore, the later filter has been compared with a new CCI+ based filter. It was found that the number of FTFNs required in such application is less than the number of CCI+ and still enjoying almost all the attractive features of the optimum filter.

Finally, a new voltage mode (MISO) universal filter using the FTFN has been proposed. The proposed filter incorporates only two FTFNs and five passive elements and enjoys independent control of the center frequency and the bandwidth. This filter was compared
with CCII+ counterparts. It was found that the FTFN based filter using less active element achieves almost the same advantages of the CCII+ based filter and offers additional features. For example, it provides accessory outputs in additional to the main output. Also, this filter was compared with a new CFA-based filter and it was found that they are very similar. This implies that the FTFN is very powerful in such application.

In addition, it can be expected, according to the experimental results that FTFN-based filters would be superior in practice if a real FTFN is available. In conclusion, the work done in this chapter shows that the FTFN appears to be a very strong candidate for current mode and voltage mode universal filters.
CHAPTER 3

APPLICATIONS OF THE FTFN:

II. SINUSOIDAL OSCILATORS

3.1 Introduction

Sinusoidal oscillators are essential for analog signal processing circuits. They have a wide range of applications in communication, control, testing and measurement systems. Actually, sinusoidal oscillators are nonlinear circuits. However, their design procedure
consists of two parts. The first is a linear analysis in the frequency domain. The second step is to provide a nonlinear mechanism for amplitude control.

In most cases, the amplitude of the generated sine wave is limited automatically by the nonlinear characteristics of the active devices involved. However, sometimes a separate circuit is added to perform amplitude limitation. A simple limiting circuit can be implemented using two back-to-back diodes.

Over the years sinusoidal oscillators were designed using active integrated circuits such as the Op Amp, the current conveyor and the current-feedback amplifier. Recently, the design of sinusoidal oscillators based on FTFN is attracting the attention of several researchers in the field [21-23]. The main advantage of using FTFN to design sinusoidal oscillators is its capability of providing output current. In fact, a single FTFN sinusoidal oscillator associated with high impedance output current is easy to design. This feature was not reported for oscillators based on other active devices.

New current mode sinusoidal oscillator circuits are proposed in this chapter. A new generic oscillator circuit built around single FTFN is proposed in section 3.3. Different categories of sinusoidal oscillators are obtained from this generic circuit. In section 3.4 new sinusoidal oscillators using two FTFN are proposed. Each of these oscillators are associated with interesting features.
3.2 Assessing a Sinusoidal Oscillator

The main factors considered in evaluating a sinusoidal oscillator are as follows

1) The relation between the condition of oscillation and the frequency of oscillation can be either independent, orthogonal (only one can be changed without disturbing the other), or interdependent. The optimal oscillator characterized by independent control of frequency of oscillation and the condition of oscillation is described as variable frequency and amplitude oscillator where the frequency and condition of oscillation are variable. Thus allowing control of the frequency and the amplitude of the generated sine wave. There are two types of orthogonal control oscillators: variable frequency constant amplitude, which results if only the frequency of oscillation is adjustable, and variable amplitude single frequency oscillator, which results if only the condition of oscillation is controllable. If the frequency of oscillation and condition of oscillation are interdependent then the circuit reduces to a single frequency oscillator with constant amplitude.

2) The number of floating capacitors. It is highly preferred if the oscillator employs only grounded capacitors to reduce the effect of the stray capacitances and to be suitable for integration.

3) It is advantageous to obtain grounded resistor control of the frequency of oscillation and the condition of oscillation. If the frequency of oscillation and condition of oscillation
are adjusted via grounded resistors, then programmable control can be easily obtained by replacing these resistors with active resistor based on JETF or MOSFT transistors. Also, if the control is via a grounded capacitor, electronic tuning is achievable using a varactor. However, resistance control is easier and more attractive.

4) The optimum oscillator is canonical: using minimum number of passive and active components.

These characteristics will be used in the following sections for evaluating the proposed oscillator circuits.

3.3 Current-Mode Sinusoidal Oscillators Using Single FTFN

New current mode sinusoidal oscillator circuits using single FTFN are designed. The proposed circuits are systematically derived from a generic circuit. The obtainable circuits are classified in three categories. In the first group, comprising six passive elements with floating capacitors, the condition of oscillation can be adjusted without disturbing the frequency of oscillation while the frequency of oscillation can not be adjusted without affecting the condition of oscillation. In the second group, using seven passive elements with floating capacitors, the frequency and the condition of oscillation can be
independently controlled. In the third group, using eight passive elements with grounded capacitors, the frequency and the condition of oscillation are independently adjustable.

Fig.3.1 shows the generic circuit from which the above mentioned three groups can be obtained.

![Proposed Generic Oscillator Circuit](image)

**Fig.3.1: Proposed Generic Oscillator Circuit**

Assuming ideal FTFN characterized by $I_x=I_y=0$, $V_x=V_y$ and $I_z=I_w$, routine analysis of the generic oscillator structure shown in Fig.3.1 yields the following characteristic equation:
\[ Y_1Y_2Y_4^*Y_1Y_3Y_4^*Y_1Y_4Y_6^*Y_1Y_3Y_6^*Y_2Y_3Y_5 = 0 \] (3.1)

Using the above equation it is possible to obtain a large number of oscillator structures.

The analysis in the following sections will proceed by considering some selected special cases.

Case I: Using eight passive elements

Based on (3.1), using eight passive elements only, appropriate RC element combinations utilizing grounded capacitors, produce the oscillator circuits summarized in Table 3.1.

**Table 3.1: Frequency and condition of oscillation of the oscillator circuits of Case I**

<table>
<thead>
<tr>
<th>Circuit</th>
<th>( Y_1 )</th>
<th>( Y_2 )</th>
<th>( Y_3 )</th>
<th>( Y_4 )</th>
<th>( Y_5 )</th>
<th>( Y_6 )</th>
<th>Frequency of oscillation ( (\omega_0^2) )</th>
<th>Condition of Oscillation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>( G_1 )</td>
<td>( G_2 )</td>
<td>( G_3 )</td>
<td>( sC_4 )</td>
<td>( G_1^* )</td>
<td>( sC_6 )</td>
<td>[ \frac{G_5(G_1G_6 - G_5G_1)}{G_1C_4C_6} ]</td>
<td>( C_5G_2G_3 = G_1[G_2+G_3+G_6+C_4+(G_5+G_4)C_6] )</td>
</tr>
<tr>
<td>2</td>
<td>( G_1 )</td>
<td>( G_2 )</td>
<td>( G_3 )</td>
<td>( G_1^* )</td>
<td>( sC_4 )</td>
<td>( G_1^* )</td>
<td>( sC_5 )</td>
<td>( sC_6 )</td>
</tr>
</tbody>
</table>

Case II: using seven passive components

Using seven passive components oscillators are generated from four different subclasses as follows:

1) \( Y_5 = \infty \) (short circuit) and \( Y_4 = 0 \)
If $Y_3 = \infty$ and $Y_6 = 0$, (3.1) becomes

$$Y_1 Y_6 - Y_2 Y_5 = 0$$  \hspace{1cm} (3.2)

Based on (3.2), and using seven passive appropriate RC elements only yields the oscillator circuits summarized in Table 3.2.

### Table 3.2: Frequency and condition of oscillation of the oscillator circuits of Case II-1

<table>
<thead>
<tr>
<th>Circuit</th>
<th>$Y_1$</th>
<th>$Y_2$</th>
<th>$Y_3$</th>
<th>$Y_6$</th>
<th>Frequency of oscillation ( (\omega_0^2) )</th>
<th>Condition of Oscillation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$G_1 + \frac{sC_1}{s}$</td>
<td>$G_2$</td>
<td>$G_5 + \frac{G_6}{sC_6}$</td>
<td>$G_6$</td>
<td>$G_1 G_6 - G_2 G_5$ \hspace{1cm} $C_3 C_6$</td>
<td>$C_2 G_2 = C_1 G_6 + C_6 G_1$</td>
</tr>
<tr>
<td>2</td>
<td>$G_1 + \frac{sC_1}{s}$</td>
<td>$G_2 + \frac{sC_2}{s}$</td>
<td>$G_5$</td>
<td>$G_6 + \frac{G_6}{sC_6}$</td>
<td>$G_1 G_5 - G_2 G_4$ \hspace{1cm} $C_1 C_6$</td>
<td>$C_2 G_3 = C_1 G_6 + C_6 G_1$</td>
</tr>
<tr>
<td>3</td>
<td>$G_1 + \frac{sC_1}{s}$</td>
<td>$G_2 + \frac{sC_2}{s}$</td>
<td>$G_5 + \frac{G_6}{sC_6}$</td>
<td>$G_5$</td>
<td>$G_1 G_5 - G_2 G_4$ \hspace{1cm} $C_1 C_6$</td>
<td>$C_2 G_3 = C_1 G_6 + C_6 G_1$</td>
</tr>
<tr>
<td>4</td>
<td>$G_1$</td>
<td>$G_2 + \frac{sC_2}{s}$</td>
<td>$G_5 + \frac{G_6}{sC_6}$</td>
<td>$G_6$</td>
<td>$G_2 G_5 - G_1 G_6$ \hspace{1cm} $C_2 C_1$</td>
<td>$C_2 G_4 = C_1 G_2 + C_2 G_5$</td>
</tr>
</tbody>
</table>

2) $Y_3 = \infty$ and $Y_6 = 0$

Equation (3.1) reduces to the following relation if $Y_3 = \infty$ and $Y_6 = 0$.

$$Y_1 Y_4 - Y_2 Y_5 = 0$$  \hspace{1cm} (3.3)

By selecting seven appropriate RC combinations, the oscillator circuits summarized in Table 3.3 are obtained from (3.3).
Table 3.3: Frequency and condition of oscillation of the oscillator circuits of Case II-2

<table>
<thead>
<tr>
<th>Circuit</th>
<th>$Y_1$</th>
<th>$Y_2$</th>
<th>$Y_4$</th>
<th>$Y_5$</th>
<th>Frequency of oscillation ($\omega_o^2$)</th>
<th>Condition of Oscillation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$G_1$ + s$C_1$</td>
<td>$G_2$ + s$C_2$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$ + s$C_5$</td>
<td>$\frac{G_1G_4 - G_2G_5}{C_1C_4}$</td>
<td>$C_3G_2 = C_1G_4 + C_4G_1$</td>
</tr>
<tr>
<td>2</td>
<td>$G_1$ + s$C_1$</td>
<td>$G_2$ + s$C_2$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$</td>
<td>$\frac{G_1G_4 - G_2G_5}{C_1C_4}$</td>
<td>$C_2G_5 = C_1G_4 + C_4G_1$</td>
</tr>
<tr>
<td>3</td>
<td>$G_1$ + s$C_1$</td>
<td>$G_2$ + s$C_2$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$</td>
<td>$\frac{G_2G_4 - G_1G_5}{C_2C_4}$</td>
<td>$C_1G_4 = C_2G_3 + C_3G_2$</td>
</tr>
<tr>
<td>4</td>
<td>$G_1$</td>
<td>$G_2$ + s$C_2$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$</td>
<td>$\frac{G_2G_4 - G_1G_5}{C_2C_4}$</td>
<td>$C_4G_1 = C_2G_3 + C_3G_2$</td>
</tr>
</tbody>
</table>

3) $Y_2 = \infty$ and $Y_6 = 0$

For $Y_2 = \infty$ and $Y_6 = 0$, the following equation is obtained from (3.1).

$$Y_1Y_2 - Y_1Y_5 = 0 \quad (3.4)$$

Based on (3.4), the oscillator circuits listed in Table 3.4 are obtained by selecting seven appropriate RC elements combinations.

Table 3.4: Frequency and condition of oscillation of the oscillator circuits of Case II-3

<table>
<thead>
<tr>
<th>Circuit</th>
<th>$Y_1$</th>
<th>$Y_3$</th>
<th>$Y_4$</th>
<th>$Y_5$</th>
<th>Frequency of oscillation ($\omega_o^2$)</th>
<th>Condition of Oscillation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$G_1$ + s$C_1$</td>
<td>$G_3$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$ + s$C_5$</td>
<td>$\frac{G_1G_4 - G_3G_5}{C_1C_4}$</td>
<td>$C_3G_2 = C_1G_4 + C_4G_1$</td>
</tr>
<tr>
<td>2</td>
<td>$G_1$ + s$C_1$</td>
<td>$G_3$ + s$C_3$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$</td>
<td>$\frac{G_1G_4 - G_3G_5}{C_1C_4}$</td>
<td>$C_2G_5 = C_1G_4 + C_4G_1$</td>
</tr>
<tr>
<td>3</td>
<td>$G_1$ + s$C_1$</td>
<td>$G_3$ + s$C_3$</td>
<td>$G_4$</td>
<td>$G_5$ + s$C_5$</td>
<td>$\frac{G_2G_4 - G_3G_5}{C_2C_4}$</td>
<td>$C_1G_4 = C_2G_3 + C_3G_2$</td>
</tr>
<tr>
<td>4</td>
<td>$G_1$</td>
<td>$G_3$ + s$C_3$</td>
<td>$G_4$ + s$C_4$</td>
<td>$G_5$</td>
<td>$\frac{G_2G_4 - G_3G_5}{C_2C_4}$</td>
<td>$C_4G_1 = C_2G_3 + C_3G_2$</td>
</tr>
</tbody>
</table>
4) $Y_3 = \infty$

For $Y_3 = \infty$, equation (3.1) reduces to following relation:

$$Y_1Y_4 + Y_1Y_6 - Y_2Y_5 = 0$$

(3.5)

Based on (3.5) and using seven appropriate RC elements combinations yields the oscillator circuits listed in Table 3.5.

<table>
<thead>
<tr>
<th>Circuit</th>
<th>$Y_1$</th>
<th>$Y_2$</th>
<th>$Y_4$</th>
<th>$Y_5$</th>
<th>$Y_6$</th>
<th>Frequency of oscillation $(\omega_n^2)$</th>
<th>Condition of Oscillation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$G_1$</td>
<td>$G_2 + \frac{sC_2}{sC_3}$</td>
<td>$G_4$</td>
<td>$G_5 + \frac{sC_5}{C_3}$</td>
<td>$\frac{G_2G_6 - G_4G_5}{C_2C_3}$</td>
<td>$C_6G_1 = C_5G_2 + C_2G_5$</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>$G_1$</td>
<td>$G_2 + \frac{sC_4}{sC_5}$</td>
<td>$G_4$</td>
<td>$G_5 + \frac{sC_5}{C_4}$</td>
<td>$\frac{G_2G_6 - G_4G_5}{C_2C_3}$</td>
<td>$C_4G_1 = C_2G_5 + C_5G_2$</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>$G_1 + \frac{1}{sC_1}$</td>
<td>$G_2$</td>
<td>$G_4$</td>
<td>$G_5 + \frac{sC_5}{C_5}$</td>
<td>$\frac{G_2G_6 - G_4G_5}{C_2C_3}$</td>
<td>$C_5G_2 = C_1G_4 + C_6G_1$</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>$G_1 + \frac{1}{sC_1}$</td>
<td>$G_2 + \frac{sC_4}{sC_2}$</td>
<td>$G_4$</td>
<td>$G_5$</td>
<td>$\frac{G_2G_6 - G_4G_5}{C_2C_3}$</td>
<td>$C_2G_5 = C_4G_1 + C_1G_6$</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.5: Frequency and condition of oscillation of the oscillator circuits of Case II-4

Case III: Incorporating six passive elements

If $Y_6 = 0$ (open circuit), equation (3.1) reduces to

$$Y_1Y_4(Y_2 + Y_3) - Y_2Y_3Y_5 = 0$$

(3.6)

Based on (3.6), and incorporating six passive elements only, proper RC element combinations yields the oscillator circuits summarized in Table 3.6.
Table 3.6: Frequency and condition of oscillation of the oscillator circuits of Case III

<table>
<thead>
<tr>
<th>Circuit</th>
<th>$y_1$</th>
<th>$y_2$</th>
<th>$y_3$</th>
<th>$y_4$</th>
<th>$y_5$</th>
<th>Frequency of oscillation ($\omega_s^2$)</th>
<th>Condition of Oscillation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$G_1$</td>
<td>$sC_2$</td>
<td>$G_3$</td>
<td>$G_4$+</td>
<td>$G_5$</td>
<td>$\frac{G_2G_4}{C_2C_4}$</td>
<td>$C_2G_3G_4 = G_1(G_3C_2+G_5C_4)$</td>
</tr>
<tr>
<td>2</td>
<td>$G_1+$</td>
<td>$sC_1$</td>
<td>$G_3$</td>
<td>$G_4$</td>
<td>$G_5$</td>
<td>$\frac{G_1G_3}{C_1C_2}$</td>
<td>$C_2G_3G_4 = G_4(G_3C_1+G_1C_2)$</td>
</tr>
<tr>
<td>3</td>
<td>$G_1+$</td>
<td>$G_2$</td>
<td>$sC_3$</td>
<td>$G_4$</td>
<td>$G_5$</td>
<td>$\frac{G_2G_5}{G_4C_3}$</td>
<td>$C_1G_3G_4 = G_4(G_2C_1+G_1C_3)$</td>
</tr>
<tr>
<td>4</td>
<td>$G_1$</td>
<td>$G_2$</td>
<td>$sC_3$</td>
<td>$G_4$+</td>
<td>$G_5$</td>
<td>$\frac{G_2G_4}{C_3C_4}$</td>
<td>$C_1G_2G_4 = G_1(G_4C_2+G_2C_4)$</td>
</tr>
</tbody>
</table>

From Table 3.1, it can be seen that circuits (1) and (2) enjoy independent grounded-control of the frequency of oscillation and the condition of oscillation. The frequency of oscillation and the condition of oscillation can be controlled independently by adjusting the grounded resistance $G_5$ and the grounded capacitor $C_4$ respectively.

From Table 3.2, it can be seen that circuit (1) enjoys independent control of the frequency of oscillation via the grounded resistor ($R_s = 1/G_5$) and the condition of oscillation via the grounded capacitor $C_5$. Similarly, the frequency of oscillation of circuit (4) can be controlled by adjusting the grounded resistor ($R_6 = 1/G_6$) without varying the condition of oscillation. Also, the condition of oscillation is independently controllable via the grounded capacitor $C_6$. 
From Table 3.3, it can be seen that the frequency and the condition of oscillation of circuit (1) are independently controllable using the grounded resistor \((R_5=1/G_5)\) and the grounded capacitor \(C_5\) respectively. Moreover, it can be seen that circuit (4) enjoys independent control of the frequency of oscillation via the grounded resistor \((R_4=1/G_4)\). Also, the condition of oscillation can be adjusted without disturbing the frequency of oscillation via the grounded capacitor \(C_4\).

From Table 3.4, it can be seen that circuit (1) enjoys independent control of the frequency of oscillation via the grounded resistance \((R_5=1/G_5)\) and the condition of oscillation via the grounded capacitor \(C_5\). Similarly, the frequency of oscillation of circuit (4) can be controlled by adjusting the grounded resistor \((R_4=1/G_4)\) without varying the condition of oscillation. Also, the condition of oscillation is independently controllable via the grounded capacitor \(C_4\).

From Table 3.5, it can be seen that for the circuit (1), the frequency of oscillation can be controlled by adjusting the grounded resistor \((R_5=1/G_5)\) without varying the condition of oscillation and the condition of oscillation can be controlled by adjusting the grounded capacitor \(C_6\) without disturbing the frequency of oscillation. Similarly, the frequency and the condition of oscillation of circuit (2) can independently be controlled using the grounded resistor \((R_6=1/G_6)\) and the grounded capacitor \(C_4\) respectively.
From Table 3.6, it can be seen that for all the four circuits the condition of oscillation can be controlled using a grounded resistor \((R_S = 1/G_S)\). However, the frequency of oscillation can not be varied without disturbing the condition of oscillation. Thus, all of these circuits are single frequency oscillators. Such oscillators are suitable for systems incorporating automatic amplitude control systems.

In all the proposed circuits, the output current can be obtained from a high impedance outlet, without adding any additional active elements. Moreover, it is worth mentioning that all the circuits of Table 3.2, Table 3.3 and Table 3.4 incorporate a resistor in parallel with each capacitor. This allows the use of lossy capacitors in realizing these oscillator circuits. Also, it can be seen from Tables 3.1-3.5 that all the proposed oscillator circuits are capable for generating low and high frequencies of oscillation. This attribute is due to the difference terms in the frequency of oscillation expressions.

### 3.3.1 Experimental Results

The theoretical analysis of the proposed circuits were verified experimentally. The experimental results reported were obtained using the two-current conveyor realization of the FTFN shown in Fig.1.8(b). According to the previous discussion, circuit (1) of Table 3.1, shown in Fig. 3.2, and Circuit (4) of Table 3.4, shown in Fig.3.3 [63] (see the Appendix for the complete paper), are among the most attractive circuits. Sample of the experimental results of circuit of Fig.3.2 are shown in Fig.3.4.(a) and a typical generated wave form is shown in Fig.3.4(b).
Fig. 3.2: Proposed Oscillator of Circuit (1) of Table 3.1

Fig. 3.3: Proposed Oscillator of Circuit (4) of Table 3.4
Fig. 3.5 (a): Experimental results of the circuit of Fig. 3.2: Experimental results (-). Simulation results of Sloboda model (---). Simulation results using CCH+ (+). Theoretical analysis (--). $(R_1 = R_2 = R_3 = R_4 = 10k\Omega, C_1 = C_2 = 100pF, C_3 = 470pF)$. 

Fig. 3.5(a): Typical Output Wave-form Measured Across 500Ω Resistor Connected to Node W of the FTFN Oscillator of Fig. 3.2 with: $R_1 = R_2 = R_3 = R_4 = 10k\Omega$, $R_5 = 11k\Omega$, $C_1 = C_2 = 100pF$, $C_3 = 470pF$. ($f_c = 43.2kHz$, $V_{p-p} = 380mV$)
It can be noticed from Fig. 3.4(a) that there is slightly large deviation about 18% between experimental results and the theoretical analysis. One of the main sources of error is the nonideal characteristic of CCs used in implementing the FTFN. To confirm this conjecture, simulation results obtained using nonideal models of CC's are also included in Fig. 3.4(a).

First, Svoboda model of CCII+ [64], and shown in Fig. 3.5, was used in the simulation. The Svoboda parameters of the practical CCII+ type AD844 are \((R_x=50\Omega, R_y=10M\Omega, R_z=3M\Omega, C_z=4.5pF)\). The results, shown in Fig. 3.5(a), indicate that the error between measurement and simulation is small; not more than 6%.

Also, the circuit was simulated using transistor-level realization of the CCII+ (AD844). The transistors parameters used were obtained from AT & T with these model numbers (NR100N and PR100N). The simulation results obtained are also shown in Fig. 3.5(a) (Appendix E includes the input file of the program). It appears that the error does not exceed 4.5%. The small error remaining between the simulation and experimental results is mainly due to the stray capacitances resulting from the breadboard realization.
Fig.3.5: Svooboda Model of Non-Ideal Current Conveyor CCII+

Also, experimental and theoretical results obtained from the circuit of Fig.3.3 are shown in Fig.3.6(a). A typical wave form signal obtained from the experimental results is shown in Fig.3.6(b). It was found that the error between the experimental and theoretical results is a bit large about 11%. Thus, the circuit was reanalyzed with a more complicated model (Svooboda model) [64] of the CCII+ used to implement the FTFN. The resulting characteristic equation is given by:

\[ Y_{27}Y_1(Y_{27} + Y_1) + Y_{57}Y_1(Y_{27} + Y_3) + Y_1(Y_{27} + Y_3)(Y_{57} + Y_1) + \frac{1}{2} G_x (Y_5 Y_{27} - Y_{27} Y_1) = 0 \quad (3.7) \]

where \( Y_{27} = G_4 + G_7 + sC_4 \), \( Y_3 = G_1 + sC_3 \), \( Y_{57} = G_5 + G_7 + sC_5 \), \( Y_1 = G_1 \), \( Y_2 = G_x + sC_2 \) and \( G_x = 1/R_x \).
From the frequency of oscillation can be expressed as:

\[
\omega_c^2 = \frac{(G_1 - G_4)(G_2 + G_4) + G_1(G_2 + G_4 + G_5) + (G_2 + G_4)(G_2(G_4 + G_5 + G_6) + 0.5G_3G_4) - 0.5G_3G_4(G_4 - C_4)}{C_j[G_1(G_4 - G_2 - G_4) + G_2G_3] + (G_3 + G_4)(G_2 + G_4 + G_5) + 0.5G_3G_4} + (C_2 + C_3)(C_2 + G_2) + (C_4 - G_4 + G_4)
\]

(3.8)

The Svoboda model analysis reduces the gap between the experimental results and the theory to a maximum error of 8% as shown in Fig 3.6(a). Moreover, it is expected that the measured frequency of oscillation is less than the theoretical calculation due to the effect of the stray capacitances especially at high frequencies as shown in the Fig 3.6(a).

\[\text{Fig.3.6(a): Experimental results obtained from the circuit of Fig.3.3 Experimental results (•),} \]

\[\text{Svoboda model (- -), Theoretical (—)(R_1=R_2=R_3=10k\Omega, C_1=200pF, C_2=C_3=100pF).}\]
Fig. 3.6(b): Typical Output Wave form Measured Across 100Ω Resistor Connected to Node W of the FTFN Oscillator of Fig. 3.3 with: \( R_1 = R_3 = R_5 = 10kΩ, \ R_4 = 1MΩ, \ C_3 = C_5 = 100pF, \ C_4 = 200pF. \)

\( (f_s = 139.2kHz, \ V_{p-p} = 100mv) \)

3.3.2 Summary

The main advantage of using single FTFN in designing sinusoidal oscillators is that it is possible to obtain output current outlet at high impedance. This feature is not reported for
any other active device. Among the previously proposed oscillators [18-20], only one of
Senani's circuit [18] enjoys independent control of the parameters and can provide output
current without adding additional active elements. However, this circuit comprises two
floating capacitors. Circuit of Fig. 3.3, using same number of passive components (seven),
enjoys independent control of parameters and can provide output current while
incorporating only one floating capacitor [63]. In addition, it is capable of generating low
frequency of oscillation without calling for large capacitor or resistor values. Furthermore,
lossy capacitors can be used in implementing this oscillator because each of its capacitors
is connected in parallel with a resistor.

Moreover, some oscillator circuits using single FTFN but can not provide an output
current were previously reported [19,20]. Among those the oscillator circuit of [19] seems
to be the best. It exhibits interesting features: enjoying independent control of the
frequency of oscillation, incorporating grounded capacitors and can provide two output
currents in quadrature.

However, this circuit suffers from some disadvantages. A floating resistor is used to adjust
the condition of oscillation. Also, two additional current conveyors must be used to sense
the two quadrature currents through the resistor ($R_4$) and capacitor $C_3$ and convey them to
high impedances. Consequently, $R_4$ which is responsible of adjusting the frequency of
oscillation becomes a floating resistor.
The circuit Fig 3.2 using eight passive elements can provide an output current at high impedance. Also, two quadrature currents flowing through $R_e$ and $C_e$ can be regenerated at high output impedances by using two current followers. A typical quadrature waveforms obtained from experimenting the circuit of Fig 3.2 is shown in Fig 3.7. Moreover, this circuit incorporates grounded capacitors and enjoys independent grounded element control of its parameters. Also, the circuit is capable of generating low and high frequencies of oscillation.

Fig.3.7: Typical Current Mode Quadrature Signals Measured Across 2.5kΩ for the Upper Signal (Current through $R_e$) and 500Ω for the lower signal (Current through $C_e$) from the Circuit of Fig.3.2 with: $R_1=R_2=R_3=10k\Omega$, $R_e=1\Omega$, $C_4=C_6=100pF$, $C_5=470pF$

$\{V_{p-p\text{across the }2.5k\Omega\text{ resistors}}=480mV$, $V_{p-p\text{across the }500\Omega\text{ resistors}}=144mV$, $f_o=43.2kHz\}$
3.4 SINUSODAL OSCILLATORS USING TWO FTFNs

Sinusoidal oscillators built around more than one FTFN has not been investigated yet. Although it is expected that using two FTFN to design sinusoidal oscillators can provide new interesting features over single FTFN oscillator, no single paper has been reported discussing these advantages. In this section, the inherent advantages of designing sinusoidal oscillators based on two FTFNs are investigated. New sinusoidal oscillators incorporating two FTFNs are proposed. These new oscillators offer the important characteristics required for some engineering applications. The first proposed circuit using two FTFNs is shown in Fig. 3.8

Fig. 3.8: Variable Phase Proposed Oscillator Using Two FTFNs
Routine analysis, assuming ideal FTFNs characterized by $l_1 = l_2 = 0$, $V_1 = V_2$, and $l_3 = l_4$, yields the following characteristic equation:

$$s^2 C_1 C_2 G_1 G_2 + s C_1 G_1 G_3 G_4 + s C_2 G_1 G_5 G_6 - s C_1 G_2 G_3 G_4 + G_2 G_4 G_6 = 0$$  \hspace{1cm} (3.9)

It can be seen that the frequency of oscillation and the condition of oscillation are respectively expressed as:

$$\omega = \sqrt{\frac{G_2 G_4 G_6}{C_1 C_2 G_1 G_2}} = \sqrt{\frac{R_1}{C_1 C_2 R_4 R_5 R_6}}$$  \hspace{1cm} (3.10)

$$\frac{C_2}{C_1} = \frac{G_2 G_4}{G_1 G_6} - 1 = \frac{R_1 R_4}{R_2 R_5} - 1$$  \hspace{1cm} (3.11)

Thus, from (3.10) and (3.11), the frequency of oscillation can be controlled by adjusting the grounded resistor $R_6$ without disturbing the condition of oscillation. Also, the floating resistor $R_4$, which is not involved in the condition of oscillation can be selected to generate the desired range of frequency of oscillation. For example, $R_4$ can be selected to be small to generate low frequencies of oscillation without requiring large capacitors. Moreover, it can be seen that the condition of oscillation can not be changed without disturbing the frequency of oscillation. However, adding two resistors $R_{C1} = G_{C2}$ and $R_{C2} = 1/G_{C2}$ as shown in Fig.3.9, do not change the characteristic equation of the circuit and can be used to control the output current amplitudes and phases.
Fig. 3.9: Adding Resistors to the Circuit of Fig. 3.8 for Controlling the Phase Shift

It can be shown that $R_{c1}$ can be used to adjust the amplitude and the phase of the output current $I_{o1}$. Similarly, $R_{c2}$ can be used to vary the amplitude and the phase of the output current $I_{o2}$. Thus, by adding an optional resistor the phase shift between the two output currents can be obtained from the following relation:

\[
\frac{I_{o2}}{I_{o1}} = \frac{C_s \left[ G_4 G_7 G_4 - \omega_0^2 C_1 C_2 (G_1 + G_4 + G_{c4}) \right] + j \omega_0 \left[ (C_4 G_4 + C_2 G_3) (G_1 + G_4 + G_{c4}) + C_1 G_s G_{c4} \right]}{G_3 \left[ G_2 G_5 G_6 + G_7 G_3 G_6 + G_7 G_3 (G_3 + G_{c4}) \right] - j \omega_0 C_1 G_s G_3} \tag{3.12}
\]

Therefore, the phase shift between the two currents is given by:
\[ \theta = \tan^{-1} \left( \frac{\omega \left[ (C_1 \cdot G_2 + C_2 \cdot G_3)(G_1 + G_2 + G_4) + C_1 \cdot G_2 \cdot G_3 \right]}{G_2 \cdot G_2 - \omega^2 C_2 \cdot C_2 (G_1 + G_2 + G_4)} + \tan^{-1} \left( \frac{\omega C_1 \cdot G_2 \cdot G_3}{G_2 \cdot G_2 + G_1 \cdot G_2 + G_2 \cdot G_3 (G_1 + G_4)} \right) \right) \]  

(3.13)

The proposed circuit was simulated with FTFN realized using the circuit of Fig. 14 using the Op Amp (UA709) macromodel. However, problems occurred when running the program due to performing transient analysis of very complicated circuit. Then, a current control current source is added at the output node of the Op Amp, instead of using the current mirrors, to sense the current and convey it to the fourth terminal of the FTFN. Simulation results of the variation of the frequency of oscillation of the circuit of Fig. 3.8 is shown in Fig. 3.10. The results agree very well with the theoretical analysis with maximum error of 5%.

Also, simulation results demonstrating the relation between the phase shift of the two currents and the added controlling resistors are shown in Fig. 3.11 and Fig. 12. It can be seen from the two figures that, the phase shift is varied throughout the range 87-156 degree at frequency of oscillation 72kHz. Typical waveforms of the generated signals with phase difference of 90 degree (quadrature oscillator) is shown in Fig. 3.13. The simulation results are in good agreement with the theory proposed except for small deviation which does not exceed 2.5% due to the nonideal characteristics of the Op Amp used to realize the FTFN.
Fig. 3.10: Simulation results of the Oscillator of Fig 3.8: \( R_1 = 2.2\, \text{k}\Omega, \) \( R_2 = R_3 = R_4 = R_5 = 1\, \text{k}\Omega, \)

\( C_1 = C_2 = 1\, \text{nF}. \quad \cdot \cdot \cdot \) Simulation results, \( \cdot \cdot \cdot \) Theoretical results.

Fig. 3.11: Simulation results of Phase Shift of the Oscillator of Fig. 3.9: \( R_1 = 2.2\, \text{k}\Omega, \) \( R_2 = 0.8\, \text{k}\Omega \)

\( R_3 = R_4 = R_5 = 1\, \text{k}\Omega, \) \( R_6 = 10\, \text{k}\Omega, \) \( C_1 = C_2 = 1\, \text{nF}. \quad \cdot \cdot \cdot \) Simulation results, \( \cdot \cdot \cdot \) Theoretical results.
Fig. 3.12: Simulation results of the Phase Shift of the Oscillator of Fig. 3.9: $R_1=2.2\,\text{k}\Omega$, $R_{c1}=\infty$, $R_3=R_4=R_5=1\,\text{k}\Omega$, $R_6=10\,\text{k}\Omega$, $C_1=C_2=1\,\text{nF}$. '-' Simulation results, '—' Theoretical results.

Fig. 3.13: Quadrature Current Signals Obtained from the Oscillator of Fig. 3.9
In summary, a new oscillator circuit based on two FTFNs is presented. The proposed circuit enjoys the following attractive features:

(a) Two output currents with variable phase shifts controlled via grounded resistors.
(b) Low frequencies of oscillation is generated without using large capacitors or resistors.
(c) All the capacitors are grounded.
(d) Grounded resistors control the frequency of oscillation and the amplitude of the output currents.
(e) Quadrature output currents can be provided.

3.4.1 Sinusoidal Oscillators Derived From the Proposed Oscillator Based on Two FTFNs

A number of oscillators which may exhibit new attractive features can be developed from the basic circuit of Fig.3.8. In this section the idea of [21], which discusses a single FTFN based oscillator, will be extended to be applied to the two FTFNs based oscillator circuit of Fig.3.8. Thirteen new oscillators can be developed from the oscillator circuit of Fig.3.8, since it has seven nodes as shown in Fig.3.14.
Fig. 3.14: Proposed Oscillators Derived From the Circuit of Fig. 3.8
The oscillator circuits of Fig.3.14 have the same characteristic equation of the original circuit of Fig.3.8 and given by (3.9). Thus, they exhibit the same frequency and condition of oscillation of the circuit of Fig.3.8 and given by (3.10) and (3.11) respectively. Among these oscillators, the circuit of Fig.3.14(g) appears to be attractive although it suffers from using two floating capacitors. The circuit of Fig.3.14(g) provides two trends of varying the frequency of oscillation using two grounded resistors $R_3$ and $R_6$. Frequencies of oscillation proportional to $\sqrt{R}$ are achieved by controlling $R_3$ and frequencies of oscillation proportional to $1/\sqrt{R}$ are obtained by controlling $R_6$. Thus, low and high frequencies of oscillation are achieved easily without requiring large or small capacitors respectively. Also, an output current associated with high impedance is available. Therefore, an additional resistor can be added to the circuit for adjusting the amplitude of the output current as shown in Fig.3.15.

Fig.3.15: Adding a Resistor for Controlling the Amplitude of the Circuit of Fig.3.14(g)
CHAPTER 4

NONLINEAR ANALYSIS OF THE FTFN FRONT-END

4.1 Introduction

One of the most appreciated sources of error between theoretical analysis and experimental results of an active circuit is the nonlinear behavior of the active device. For example, frequency domain (linear) analysis is incorporated in designing active filters
assuming linear operation of the FTFN. This produces some variations between the experimental result and the theoretical analysis.

Also, as indicated previously that a sinusoidal oscillator is a non-linear circuit. A linear analysis is used to approximate the design of the oscillator. However, the nonlinear behavior of the active device both controls the amplitude of the oscillation and affects the frequency of oscillation. This effect can be seen clearly from the deviation between the ideal analysis and the practical results of the proposed oscillators based on the FTFN. In experimenting the proposed oscillators practically, the FTFN was implemented by the two CCII+ realization. Thus, it is important, for the sake of completeness to consider the possible sources of nonlinear performance of practical FTFN.

4.2 Nonlinear Analysis of the FTFN Based on the (CCII+)

The two commonly used mixed translinear loops, shown in Fig. 4.1, are the heart of several current-mode integrated circuits [65-67]. Moreover, the mixed translinear loop is the main part of the circuit of available CCII+ (AD844) and can be considered to be a strong candidate to implement the FTFN. Thus the nonlinear behavior of the mixed translinear loop is analyzed in this section.
Fig. 4.1: Mixed Translinear Loops

Assuming that the current gains $\beta >> 1$, the collector currents of transistors $Q_1$ and $Q_3$ are equal to the bias current $I_0$. With no load connected at port $X$, that is $i_x = 0$, the circuit behaves as an ideal voltage follower with $V_{xy} = 0$. However, when a load is connected to port $X$, then $i_x \neq 0$, a voltage difference is established between ports $X$ and $Y$. Under small
signal conditions, with $i_1 << 2I_o$, it was shown that the equivalent resistance between ports X and Y can be expressed as [67]

$$R_e = \frac{I_e}{2I}$$  \hspace{1cm} (4.1)

where $V_T$ is the thermal voltage. Equation (4.1) indicates that the relationship between the input current $i$, and the voltage established between ports Y and X is linear. However, SPICE simulation shows that this is not the case [67]. The simulation results reported in [67] shows that a Total Harmonic Distortion (THD) of 3% was obtained with input current amplitudes of the order of $2.5 I_o$. Actually, by virtue of its derivation, (4.1) is valid only for sufficiently small values of $i$, compared to $I_o$. Therefore, it is necessary to derive an expression for the voltage established between ports Y and X when the mixed translinear loop of Fig 4.1 is driven by a relatively large input current $i$.

Considering the mixed translinear loop shown in Fig.4.1. Applying the translinear principle [68] (ignoring the Early effect and assuming large $\beta$) results on

$$I_2 I_2 = I_o$$  \hspace{1cm} (4.2)

Using the well known exponential characteristic of a transistor ($I_e = I_e \left(\frac{I_{CE}}{I_e}\right)$), the currents $I_2$ and $I_o$ can be expressed as:

$$I_2 = I, \exp\left(\frac{I_e - I_{CE}}{I_e}\right)$$  \hspace{1cm} (4.3)
and

\[ I_1 = I \exp \left( \frac{I^* - I^*}{I^*} \right) \]  \hspace{1cm} (4.4)

where \( I_1 \) is the reverse-bias saturation current of the p-n junction, \( V_T \) is the thermal voltage, and \( V \) is the voltage at the bases of \( Q_1 \) and \( Q_2 \).

Combining (4.2-4.4), the current \( i_x = I_2 - I_4 \) can be expressed as:

\[ i_x = 2I \sinh \left( \frac{I^* - I^*}{I^*} \right) \]  \hspace{1cm} (4.5)

or

\[ \frac{I^* - I^*}{I^*} = \frac{V_{sx}}{I^*} = \sinh \left( \frac{I_x}{2I_c} \right) \]  \hspace{1cm} (4.6)

Equation (4.6) indicates that the relationship between the voltage established between ports Y and X and the input current is nonlinear. For sufficiently small values of \( \frac{I_x}{2I_c} \), the \( \sinh^{-1} \) function can be approximated by the first-term of its Taylor series expansion, thus

\[ V_{sx} = V_T \frac{I_x}{2I_c} \]  \hspace{1cm} (4.7)

Using (4.7), equation (4.1) can be easily obtained.
Equations (4.5) and (4.6) can be used in the nonlinear analysis of the oscillators. Performing nonlinear design of oscillators is beyond the scope of the thesis, however. This detected nonlinearity explains some of the error between the theory and practice of the proposed FTFN-based oscillators and filters. Moreover, one essential application of the (4.6) is in the predication of the amplitudes of the harmonic and intermodulation products in the normalized voltage $V_N$ when the normalized input current $\frac{I}{2I}$ is formed of a multisinusoidal signal.

4.3 Harmonic and Intermodulation Products

The prediction of the harmonic and intermodulation product is very important in filter design to ensure that no undesired signals are generated unintentionally by FTFN nonlinearity. Also, it is highly critical for designing sinusoidal oscillators where the harmonic product is one of the main features of the oscillator. Similarly, harmonic content of an output signal is very important for nonlinear applications of the FTFN such as multipliers.

As indicated previously, equation (4.6) is nonlinear so that it can be used to predict the harmonic and intermodulation performance resulting from a multisinusoidal input current. However, in its present form it can not yield closed form expressions for the amplitudes of
the harmonic and intermodulation components. Therefore, equation (4.6) is approximated by the truncated Fourier-series model of (4.8)

\[
\frac{I\gamma}{I_T} = \sum_{n=0}^{\infty} \alpha_{2n+1} \sin \left( \frac{(2n+1)\pi}{T} \frac{I_x}{2I_o} \right)
\]  

(4.8)

Normally, the coefficients \(\alpha_{2n+1}\) are calculated using standard least square approximations or discrete Fourier transform (DFT) techniques by finding the best fit between (4.6) and (4-8). However, these approaches demand extensive computing facilities and an advance software. Alternatively, these coefficients can be obtained using the procedure suggested in [69.70]. For \(N=9\), values of \(\alpha_1=1.307, \alpha_3=-0.0458, \alpha_5=0.0316, \alpha_7=-0.0143, \alpha_9=-0.0089, \alpha_{11}=-0.0059, \alpha_{13}=0.0043, \alpha_{15}=-0.0032, \alpha_{17}=0.0025, \alpha_{19}=-0.0020\) and \(T=4.0\), yields a relative root-mean-square (RRMS) error of 0.0027 for values of \(\|I_x\| \leq 4I_o\).

Now, the amplitudes of the harmonic and intermodulation products of the normalized voltage \(V_{yx}\) can be predicted using (4.8) when the normalized input current \(\frac{I_x}{2I_o}\) consists of a multisinusoidal signal of the form:

\[
\frac{I_x}{2I_o} = \sum_{m=1}^{M} X_m \sin(\omega_m t), \sum_{m=1}^{M} |X_m| \leq 2.0
\]  

(4.9)

Combining (4-8) and (4-9) and using the identities:
\[
\sin(\beta \sin(\omega t)) = 2 \sum_{l, \infty}^r J_{2l+1}(\beta) \sin(2l + 1) \omega t
\]

\[
\cos(\beta \sin(\omega t)) = J_0(\beta) - 2 \sum_{l=1}^r J_{2l}(\beta) \cos(2l) \omega t
\]

where \( J_0(\beta) \) is the Bessel function of order 0. It can be shown, after simple mathematical manipulations, that the amplitude of the normalized voltage component of frequency \( \sum \gamma_m \omega_m \) and order \( \sum \gamma_m \), where \( \gamma_m \) is an integer, is given by

\[
\frac{I_{\text{vn+1,2}}}{I_2} = 2 \sum_{n=1}^r \prod_{m=1}^M J_0 \left( \frac{(2n+1)\pi}{T} X_m \right) \quad (4.10)
\]

Generally, the Bessel functions are evaluated using built-in subroutines in most mainframe computers. In some cases with very large or very small arguments, however, simple approximations of Bessel functions can be used [71].

To investigate the total harmonic distortion (THD) of the normalized voltage \( \frac{I_{\text{vn}}}{I_2} \), the following case of a normalized current is considered

\[
\frac{I_x}{2I_n} = X \sin(\omega t) \quad (4.11)
\]

Combining (4.10) and (4.11), it can be shown that the relative kth-harmonic current components, \( k = \text{odd integer} \), can be expressed as
\[ H_s \equiv \frac{1}{1 - \frac{X^2}{(2n-1)\pi}} \sum_{n=1}^{\infty} \frac{\alpha_{2n-1}}{\frac{X^2}{2n-1}} \cdot k = 3, 5, 7. \quad (4.12) \]

For sufficiently small values of \( X \) so that \( 19 \pi X / T \ll 1 \), the Bessel functions can be approximated by

\[ J_n(\beta) \approx 1 - \left( \frac{\beta}{\sqrt{2}} \right)^2 \]

and

\[ J_n(\beta) \approx \left( \frac{\beta}{\sqrt{2}} \right)^k \text{ for } k \neq 0, k = 0 \]

and (4.12) reduces to:

\[ H_s \equiv \frac{1}{k! \sqrt{2T}} \sum_{n=0}^{\infty} \frac{(2n+1)^k \alpha_{2n-1}}{(2n+1)\alpha_{2n+1}} \cdot X^{k+1}, k = 3, 5, 7. \quad (4.13) \]

Calculations using (4.13), yields

\[ H_3 \approx 0.079 X^2 \quad (4.14) \]

\[ H_5 \approx 0.59 X^2 \quad (4.15) \]

and

\[ H_7 \approx 1.03 X^2 \quad (4.16) \]
4.4 Simulation Results

The mixed translinear loops of Fig 4.1 were simulated using SPICE with identical high β npn and pnp default transistors. The circuits were supplied from dc sources of ±2.5V and two DC bias currents $I_n=100\mu A$. Variation of the total-harmonic distortion (THD) of the voltage $V_{in}$ with the peak to peak magnitude of the input current $i_n$ are shown in Fig. 4.2. Also, the calculated results using (4.12) are shown in the same figure. It can be seen that the agreement between the simulation results and calculated results using (4.12) is very good. This confirms the validity of the analysis presented.

![Graph showing THD vs. peak to peak input current]

Fig. 4.2: Calculated and Simulated THD of the Translinear loops of Fig. 4.1: (—) Calculated results. (o) Simulated results of Fig. 1(a), (+) Simulated results of Fig. 1(b)
4.5 Conclusion

The nonlinear behavior of the FTFN is an important source of errors that occur between the theoretical analysis and the experimental results. The FTFN was practically implemented using two CCII+(AD844). The basic circuit of the CCII− is the translinear loop which is also widely used in designing current-mode integrated circuits. Thus the nonlinear characteristics of the translinear loops is explored to determine this nonlinearity. A compact expression for the current-voltage characteristics of the translinear loops has been developed.

The expression represents the normalized input voltage as a function of the normalized input current. Using this expression, closed-form expressions for the amplitudes of the harmonic components of the voltage resulting from a sinusoidal input current have been obtained. SPICE simulation results, in excellent agreement with calculated results, confirm the validity of the obtained expressions. Extension of the analysis presented here to obtain expressions for the amplitudes of the intermodulation components of the input voltage resulting from a multisinusoidal input current is mathematically straightforward.
CHAPTER 5

CONCLUSION AND FUTURE WORK

5.1 Introduction

In this thesis, the FTFN and its applications were discussed. The uses of FTFN in designing universal active filters and sinusoidal oscillators were investigated. In this regard, several current mode and voltage mode universal filters have been designed. Also, several sinusoidal oscillators based on a single FTFN or two FTFNs have been proposed.
In this chapter, the thesis work is summarized and some conclusions and comments regarding this work are drawn. Then, recommendations and suggestions for future work will be presented.

5.2 Conclusions

The different possible realizations of the FTFN based on the currently available active devices have been discussed indicating their associated advantages and disadvantages. The two-CCII+ based FTFN realization was selected to implement and test the proposed circuits practically. Power sensing technique applied to an Op Amp was used to realize the FTFN in performing SPICE simulations.

The FTFN wide range of applications was investigated paying more attention to those demonstrating the flexibility and versatility of the FTFN over the Op Amp and the current-conveyor. In this thesis, the use of FTFN in designing universal active filters has been enriched by proposing several active filters. Two (MISO) and two (SIMO) current mode universal filters were proposed. The advantages and disadvantages of each filter were discussed. These filters were compared with new CCII+, OTA-Op Amp counterparts. This work showed that the FTFN is very strong candidate in designing current mode
universal active filters because the FTFN based filters offer some advantages. For example, the proposed filters use lower number of active and passive components. The FTFN current mode based filters are associated with high impedance output currents suitable for cascading. Also, some of the filters based on the FTFN use only grounded passive elements which is appropriate for integration.

Also, a new (MISO) voltage mode universal filter was proposed. In contrast with the filter suggested in the literature, this filter offers independent control of the parameters $\omega_c$ and $\omega_o/Q_o$. This filter was compared with CCII+ and CFA counterparts. It was found that designing voltage mode universal filter requires less number of active elements than those required by CCII+ and still offers almost the same advantages. Moreover, it was shown that the FTFN based filter using the same number of active and passive elements enjoy almost similar features of CFA based filter counterparts.

The use of the FTFN in designing sinusoidal oscillators was investigated. Several oscillators based on a single FTFN which can provide output current associated with high impedance were proposed. Two of them were selected to be tested practically and to be compared with the already known oscillators. The comparison shows that the proposed oscillators enjoy attractive features which are not provided by the already known oscillators. For example, one of the proposed oscillator combines all of following attractive features: use of grounded capacitors, independent control of the frequency and
condition of oscillation via grounded elements, high impedance output current, quadrature current outputs.

Furthermore, a new oscillator based on two FTFNs was proposed. The proposed oscillator enjoys the following features: providing two output currents with variable phase shifts, offering grounded resistor control of the frequency of oscillation and the amplitude of the output currents, using grounded capacitors. Thus, the proposed oscillator indicates that the FTFN is very efficient in designing current mode oscillators. Finally, the nonlinear characteristics, which contributes to the oscillators amplitude limitation, of the FTFN based on the translinear loop, were explored.

To verify the theoretical analysis of the proposed circuits, the proposed circuits were either tested experimentally or simulated using SPICE simulation. Both experimental results and SPICE simulation were in good agreement with the theory presented. Although the experimental results of the proposed circuits were obtained using discrete CCII+ realization of the FTFN, the error between these results and the theoretical analysis is within the engineering limits. Thus if a real FTFN is used in implementing the proposed circuits, then it is expected that practical performance of the FTFN based circuits will be superior to the CCII+, OTA or CFA counterparts.
5.3 Future Work

Since nothing is perfect and the science and engineering is continually improving, this work can be improved and further extended in many directions:

1) Designing and implementing of fully integrated FTFN is highly recommended as the FTFN proved to be a very versatile and flexible active device. Also, it will be very useful if an additional output current and voltage buffer are added in the integration to approach the dream of universal element as was discussed in chapter 1.

2) Designing universal filter circuits which can offer some advantages over the proposed filters in terms of reducing the passive and active components as well as the matching/cancellation requirements for realizing the filters.

3) Designing universal (SIMO) voltage mode filter based on the FTFN which can compete with other devices counterparts.

4) Designing new sinusoidal oscillators which can offer the same or more advantages of the proposed circuits but using less number of passive elements.

5) Further investigation of the advantages of using more than a single FTFN in designing sinusoidal oscillators. For example, the design of current mode of multiphase oscillator based on the FTFN is still open to investigation.

6) Introducing the FTFN into the design of function generators.

7) Incorporating the nonlinear characteristics of the FTFN in designing the oscillators.
References


[27] Stevenson, J., “Use of Reciprocity and Duality to Generate Equivalent Active RC Networks”, Proceedings of ISCAS, 1985, pp.821-822.


[45] Sun, Yichuang and Fidler, J.K., "Versatile Active Biquad Based on Second-
No 1, pp 91-98


[48] Sanchez-Sinencio, E., Geiger, R.L. and Navarez-Lozano, H., "Generation of
Continuous-time Two Integrator Loop OTA Filter Structures". IEEE Transactions on

[49] Wu, J and Xie, C.-Y., "New Multifunction Active Filter Using OTAs". International

[50] Sun, Y and Fidler, J.K., "Novel OTA-C Realization of Biquadratic Transfer

[51] Siddiqi, M.A. and Ahmed, M.T., "Direct Form Active-R synthesis Techniques and
635

[52] Soderstrand, M. A. Watt, V.H.C., Gee, K.B., and Mcginty, D., "Implementation of
an Active-R Filter Building Block in Semicustom VLSI", International Journal of


National Bureau of Standards, Washington, D C
Vita

- Hussain Abdullah Ahmad Al-Zaher
- Born in Al-Awamiya (Qateef), Saudi Arabia
- Received Bachelor’s degree in Electrical Engineering from King Fahd University of Petroleum and Minerals, Dhahran, Saudi Arabia in 1994.
- Completed Master’s degree requirements at King Fahd University of Petroleum and Minerals, Dhahran, Saudi Arabia in May, 1997.
APPENDIX

(PUBLICATIONS: TWO PAPERS)
Current-Mode Sinusoidal Oscillator Using Single FTFN

MUHAMMAD TAHER ABUELMA'ATTI AND HUSAIN ABDULLAH AL-ZAHER

King Fahd University of Petroleum and Minerals
Dhahran, Saudi Arabia

(Received October 7, 1996; Accepted December 23, 1996)

ABSTRACT

A new current-mode sinusoidal oscillator circuit is presented. The proposed circuit uses single four-terminal floating nullor (FTFN), three capacitors and four resistors. The frequency of oscillation can be controlled using a grounded resistor without disturbing the oscillation conditions. The oscillation conditions can be controlled using a grounded capacitor without disturbing the oscillation frequency. Experimental results are included.

Key Words: current-mode circuits, sinusoidal oscillators

I. Introduction

The four-terminal floating nullor (FTFN) is a more flexible and versatile building block than is the operational amplifier and the current-conveyor (Nordholt, 1982; Huijsing and De Korte, 1977; Higashimura, 1991a). This explains the growing interest in using FTFN in designing current-amplifiers, voltage-to-current converters, gyrators, floating immitances (Nordholt, 1982; Huijsing, 1990; Senani, 1987a, 1987b), and more recently current-mode active-RC filters (Higashimura, 1991a, 1991b; Liu, 1995; Liu and Lee, 1996; Abuelma’atti, 1996) and sinusoidal oscillators (Senani, 1994; Liu and Liao, 1996). In Senani (1994), a procedure was described for transformation of single op-amp oscillators into FTFN-based oscillators. Among the oscillator circuits obtained using this procedure, only two enjoy independent control of the oscillation frequency and conditions (see Fig. 4(a), (c) of Senani (1994)). The first circuit (see Fig. 4(a) of Senani (1994)) has the advantage of using grounded capacitors. However, to obtain its output current, an additional current conveyor or current follower is required. The output current of the second circuit (see Fig. 4(c) of Senani (1994)) can be obtained without additional active elements. However, it uses floating capacitors. In Liu and Liao (1996) a sinusoidal oscillator circuit was presented. This circuit enjoys independent control of the oscillation frequency and conditions and has the advantage of using grounded capacitors. However, to obtain its output currents, additional active elements are required. It is noteworthy here that these three circuits cannot use lossy capacitors because no resistors are connected in parallel with the capacitors. Moreover, the three circuits can not be used to generate low frequencies without using large capacitor values. This is because no difference terms are involved in the expressions for the oscillation frequencies of the three circuits.

We present a new current-mode sinusoidal oscillator circuit using single FTFN, three capacitors and four resistors. This circuit enjoys the following attractive features:

1. It uses grounded elements for independent control of the frequency of oscillation and the oscillation conditions.
2. It can be used for high and low frequency generation.
3. It employs capacitors where a resistor is in parallel with each capacitor; thus, even lossy capacitors can be used.
4. The output current can be obtained without adding any additional active elements.

II. Proposed Circuit

The proposed circuit is shown in Fig. 1. Assuming that the port relations of the FTFN, shown in Fig. 2, can be expressed as \( I_1 = I_2 = 0 \), \( V_S = \beta V_i \) and \( I_{02} = \alpha I_{01} \), where \( \beta = 1 - \varepsilon \), \( \alpha = 1 - \delta \), \( (|\delta| < 1) \) denotes the voltage tracking error of the FTFN, and \( \alpha = 1 - \delta \), \( (|\delta| < 1) \) denotes the current-tracking error, routine analysis of the circuit shown in Fig. 1 yields the characteristic equation expressed by

\[ s^2 C_2 C_3 + s(C_3 G_3 + C_3 G_2 - \beta C_1 G_4 + \varepsilon C_2 G_4) \]


M.T. Abuelma'atu and H.A. Al-Zahe

Fig. 1. Proposed sinusoidal oscillator circuit.

\[ +G_1G_3 + eG_2G_4 - \beta G_1G_4 = 0. \]  \hspace{1cm} (1)

From Eq. (1), the frequency of oscillation and the oscillation conditions can be expressed as

\[ \omega_0^2 = \frac{G_3G_4}{C_2C_3} + \frac{eG_2G_4 - \beta G_1G_4}{C_2C_3} \]  \hspace{1cm} (2)

and

\[ C_1G_2 + C_2G_3 + eC_2G_4 = \beta C_1G_4. \]  \hspace{1cm} (3)

From Eqs. (2) and (3), it can be seen that the oscillation frequency can be adjusted by controlling the grounded resistor \( R_1 = \frac{1}{G_1} \) such that \( \beta G_1G_4 < G_2G_3 + eG_2G_4 \), without disturbing the oscillation conditions. Generation of low frequencies is possible by adjusting \( R_1 = \frac{1}{G_1} \) such that \( G_1G_4 < G_2G_3 \). Moreover, the oscillation conditions can be adjusted by controlling the grounded capacitor \( C_1 \) without disturbing the oscillation frequency. Thus, the proposed circuit enjoys independent control of the oscillation frequency and the oscillation conditions.

From Eqs. (2) and (3), it is easy to see that, while the frequency and the conditions of oscillation will be insensitive to the current tracking error, they will be slightly affected by the voltage tracking error.

III. Experimental Results

To verify the theoretical analysis, the proposed circuit was used to realize sinusoidal oscillators. At present, there is no commercial version of FTFN. Nevertheless, there are several ways to implement FTFN (Nordholt, 1982; Huijsing and De Korre, 1977; Huijsing, 1990; Senani, 1987a, 1994; Stevenson, 1984) with commercially available integrated circuits. The experimental results reported here were obtained using two second-generation current-conveyor realization of the FTFN proposed in Senani (1987a) and shown in Fig. 3. This realization was adopted because it provides very high impedances at the \( z \)-and \( w \)-terminals. However, current-conveyors are nonideal devices. It is, therefore, essential to investigate the effect of the current-convoyor nonidealities on the performance of the proposed circuit. Figure 4 shows a simplified model for a nonideal second-generation current-conveyor (Svoboda, 1994). Using the model of Fig. 4, the equivalent circuit of the proposed oscillator of Fig. 1 can be represented by the circuit in Fig. 5. Routine analysis of the circuit in Fig. 5 yields the characteristic equation given by

\[ \gamma_1\gamma_2(\gamma_3+\gamma_4)+\gamma_3\gamma_2(\gamma_1+\gamma_4)+\gamma_1(\gamma_1+\gamma_2)(\gamma_3+\gamma_4) \]
Sinusoidal Oscillator Using FTFN

From Eq. (5), it appears that the current-conveyor parasitics, \( R_x, R_y, R_z, \) and \( C_z \), will affect the frequency of oscillation.

The experimental results obtained using the values \( R_x=R_y=R_z=10 \, \Omega \), \( C_z=C_3=100 \, \text{pF} \), \( C_1=200 \, \text{pF} \) and \( R_1=10 \, \text{K} \) for the passive components with 10\% tolerances are shown in Fig. 6, and a typical output waveform at an oscillation frequency of 139.2 KHz is shown in Fig. 7. Shown also in Fig. 6 are the calculated oscillation frequencies using Eqs. (2) and (5). For \( R_1=50 \, \text{K} \), the calculated frequency of oscillation, using Eq. (2), is 142 KHz, the measured frequency is 126 KHz and the calculated frequency of oscillation using Eq. (5) is 136 KHz. It appears that the measured and calculated results are in a fairly good agreement. The difference between the measured and calculated results can be attributed to the parasitics of the current conveyors used to implement FTFN and the stray capacitances resulting from breadboard implementation.

IV. Conclusion

A current-mode sinusoidal oscillator has been presented. The proposed circuit enjoys the following

\[
\omega_0 = \frac{(G_3 + G_4 + G_p) (G_2(G_1 + G_z) + G_3(G_1 + G_z + G_z)) + (G_3 + G_z) (G_2(G_1 + G_z) + G_3(G_1 + G_z + G_z)) + \frac{1}{2} G_3 G_2 G_3 - \frac{1}{2} G_2 G_3 (G_1 + G_z)}{C_1 (C_2 (G_3 + G_4 + G_p) + C_3 G_2) + C_3 (C_2 (G_1 + G_z) + C_3 (G_1 + G_z) + G_3) + \frac{1}{2} C_3 G_2) + (C_1 + C_2) (C_3 (G_2 + G_4) + C_3 (G_1 + G_z) + G_2)}
\]

Fig. 5. Equivalent circuit of the oscillator circuit in Fig. 1.

Fig. 6. Variation of the frequency of oscillation of the circuit in Fig.
\( 1. \ R_x=R_y=R_z=10 \, \text{K} \), \( C_z=C_3=100 \, \text{pF} \), \( C_1=200 \, \text{pF} \).
--- Calculated (5). ------- Calculated (2). ⬠ Measured.
M.T. Abuelma'atti and H.A. Al-Zaher

Fig. 7. Typical output wave form measured across 100Ω resistor connected to node W of the FTFN oscillator circuit in Fig. 1. $R_1=1 \, \text{MΩ}, R_2=R_3=10 \, \text{KΩ}, C_2=C_3=100 \, \text{PF}, C_1=200 \, \text{PF}.$

advantages:

(1) Each capacitor is connected in parallel with a resistor; thus, lossy capacitors can be used.

(2) Independent control of the oscillation frequency and the conditions is possible using grounded elements.

(3) It can be used to generate low and high oscillation frequencies.

(4) The output current can be obtained without adding any additional active elements.

References


使用單一FTFN的電流模式弦波振盪器

MHAMMAD TAHER ABUELMA’ATTI AND HUSAIN ABDULLAH AL-ZAHER

King Fahd University of Petroleum and Minerals
Dhahran, Saudi Arabia

摘要

本文提出一種新的電流模式弦波振盪器電路。此電路使用一個四端浮接電流抵銷器（four terminal floating nullor, FTFN），三個電容器，以及兩個電阻器。電路的振盪頻率可以由一個接地的電阻器來控制而不影響到電路的振盪條件。電路的振盪條件則由一個接地的電容器來控制也不會影響到電路的振盪頻率。實驗結果呈現在本論文內。
Universal three input and one output current-mode filter without external passive elements

M.T. Abuelma'atti and H.A. Alzaher

Indexing terms: Current-mode circuits, Active filters

A new current-mode universal filter is proposed. The filter uses only operational amplifiers and operational transconductance amplifiers (OTAs) and can realise lowpass, highpass, bandpass, notch and allpass responses without changing circuit topology. The parameters \( \omega_0 \) and \( Q \) can be electronically tuned by adjusting the bias currents of the OTAs. The proposed circuit has low sensitivity.

Introduction: Universal voltage-mode active filters using the operational transconductance amplifier (OTA) have many advantages.
such as simplicity, integrability and programmability [1-5]. These resistorless realizations invariably demand the use of externally connected capacitors. Conversely, the voltage-mode operational amplifier-based active-R capacitorless realisations are attractive for monolithic integration and can offer wider signal bandwidths [6, 7]. However, current-mode OTA-capacitor and active-R based filters have many advantages compared with their voltage-mode counterparts [8, 9]. Therefore, a capacitorless-resistorless current-mode active filter using OTAs and operational amplifiers would be attractive for its integrability, programmability and wide frequency range of operation [10].

In this Letter, a new current-mode capacitorless-resistorless active filter using only OTAs and operational amplifiers is presented. The circuit has three inputs and one output and can realise lowpass, highpass, bandpass, notch and allpass responses without changing circuit topology.

![Fig. 1 Proposed universal current-mode filter](image)

Proposed circuit: The proposed circuit is shown in Fig. 1. Using standard notation, the OTA can be characterised by $i = g_m(v_i - v_c)$, where $g_m = I_{cm}/2V_T$ is the transconductance of the 4th OTA, $I_{cm}$ is the auxiliary bias current, $V_T$ is the thermal voltage and $v_i$ and $v_c$ are the input voltages of the OTA. Assuming internally compensated operational amplifiers with open-loop gain of $A = B/V$, where $B$ is the gain-bandwidth product of the 8th operational amplifier, routine analysis yields the current transfer function given by

$$I_{out} = g_{m1} s^2 g_{m2} I_1 s^2 g_{m3} I_2 + s B g_{m1} g_{m2} I_1 + B I_2 g_{m1} g_{m3} I_1$$

$$= s g_{m1} g_{m2} g_{m3} I_1 + s B g_{m1} g_{m2} I_2 + B I_2 g_{m1} g_{m3} I_1$$

$$= s g_{m1} g_{m2} g_{m3} I_1 + s B g_{m1} g_{m2} I_2 + B I_2 g_{m1} g_{m3} I_1$$

From eqn. 1, it can be seen that:
(i) the lowpass response can be realised with $I_1 = I_2 = 0$
(ii) the highpass response can be realised with $I_1 = I_2 = 0$
(iii) the bandpass response can be realised with $I_1 = I_2 = 0$
(iv) the notch response can be realised with $I_1 = I_2 = 0$
(v) the allpass response can be realised with $I_1 = I_2 = 0$, and $g_m = g_m = g_m$

From eqn. 1 the parameters $a_0$ and $a_0/Q_0$ can be expressed as

$$a_0 = B g_{m1} g_{m2} g_{m3} I_1$$

$$a_0/Q_0 = B g_{m1} g_{m2} g_{m3} I_1$$

From eqn. 1 it can also be seen that the lowpass DC gain and the high frequency gain of the highpass and the bandpass gain at $a_0$ equal $g_m$. Thus the lowpass, highpass and bandpass gains can be electronically tuned by adjusting the auxiliary bias current $I_{cm}$.

![Fig. 2 Simulated results](image)

Simulation results: The universal filter circuit of Fig. 1 was simulated using the PSPICE circuit simulation program. The operational amplifier LF156 and the OTA macromodel [11] were used in simulation. The multiple output OTA was simulated using parallel-connected single-ended OTAs. With a DC supply of ±10V, the results obtained for the lowpass and the notch responses are shown in Fig. 2.

Conclusion: A new universal second-order current-mode filter has been presented. The proposed filter uses only operational amplifiers and operational transconductance amplifiers. No external passive elements are used. The proposed filter enjoys the following advantages:

(i) current control of the gain and the parameters $a_0$ and $a_0/Q_0$ of the filters
(ii) independent control of the parameter $a_0$ without disturbing the parameter $a_0/Q_0$ and the gain
(iii) independent control of the parameter $a_0/Q_0$ without disturbing the parameter $a_0$ and the gain
(iv) independent control of the gain without disturbing the parameters $a_0$ and $a_0/Q_0$
(v) low sensitivities of the parameters \( \omega_n \) and \( \omega_c/Q \) to the gain-bandwidth products of the operational amplifiers and the transconductances of the operational transconductance amplifiers.
(vi) realisation of all the standard filter functions: lowpass, highpass, bandpass, notch and allpass
(vii) insensitivity to temperature variations.

© IEE 1997

11 December 1996

M.T. Abuelma'atti and H.A. Alzaher (King Fahd University of Petroleum and Minerals, Box 203, Dhahran 31261, Saudi Arabia)

References