M. Tech (VLSI AND EMBEDDED SYSTEMS) Bharti Gahlot 2018

INVERSE-SHADOW FILTERS EMPLOYING CURRENT FEEDBACK OPERATIONAL AMPLIFIER

A DISSERTATION

SUBMITTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS
FOR THE AWARD OF THE DEGREE
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IN
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Submitted by:

Bharti Gahlot

(2k16/vls/05)

Under the supervision of

Prof. Rajeshwari Pandey

Dept. of Electronics & Communications, DTU, Delhi

Department of Electronic &Communications Engineering
DELHI TECHNOLOGICAL UNIVERSITY
(Formerly Delhi College of Engineering)
Bawana Road, Delhi-110042
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Department of Electronic &Communication Engineering
DELHI TECHNOLOGICAL UNIVERSITY
(Formerly Delhi College of Engineering)
Bawana Road, Delhi-110042

CANDIDATE'S DECLARATION

I, Bharti Gahlot , Roll No(s). 2k16/vls/05 student of M. Tech. (VLSI and Embedded Systems), hereby declare that the project Dissertation titled "Inverse-Shadow Filters employing current feedback operational amplifier" which is submitted by me to the Department of Electronics and Communications Engineering, Delhi Technological University, Delhi in partial fulfillment of the requirement for the award of the degree of Master of Technology, is original and not copied from any source without proper citation. This work has not previously formed the basis for the award of any Degree, Diploma Associateship, Fellowship or other similar title or recognition.

Place: Delhi

Date:

Bharti Gahlot
2k16/vls/05
M.tech(VLSI)

Department of Electronic &Communication Engineering

DELHI TECHNOLOGICAL UNIVERSITY

(Formerly Delhi College of Engineering)

Bawana Road, Delhi-110042

CERTIFICATE

I hereby certify that the Project Dissertation titled "Implementation of Inverse Shadow Filters" which is submitted by Bharti Gahlot, Roll No. 2k16/vls/05 [VLSI and Embedded Systems], Delhi Technological University, Delhi in partial fulfillment of the requirement for the award of the degree of Master of Technology, is a record of the project work carried out by the students under my supervision. To the best of my knowledge this work has not been submitted in part or full for any Degree or Diploma to this University or elsewhere.

Place: Delhi

Date:

Prof. Rajeshwari Pandey

SUPERVISOR

Associate Dean (UG)

Dept. of Electronics and Communications

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Place: Delhi

Date:

Bharti Gahlot

2k16/vls/05

M.tech(VLSI)

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LIST OF ABBREVIATIONS

Abbreviation	Full Form
CFOA	Current Feedback Operational Amplifier
CM	Current Mode
VM	Voltage Mode
DSP	Digital Signal Processing
ASP	Analog Signal Processing
VLSI	Very Large Scale Integration
CMOS	Complementary Metal Oxide Semiconductor
ADC	Analog to Digital Convertor
DAC	Digital to Analog Convertor
IC	Integrated Circuit
AC	Alternating Current
DC	Direct Current
CC	Current Conveyor
CCII	2nd generation Current Conveyor
CCCII	2nd generation Current Controlled Current Conveyor
CDTA	Current Differencing Transconductance Amplifier
CFTA	Current Followr Transconductnce Amplifers
VOA	Voltage Operational Amplifier
VDTA	voltage difference transconductance amplifier

ZCCITA	Z-copy current inverter transconductance amplifier
Op-amp	Operational amplifier
HPF	High-pass Filter
LPF	Low-pass Filter
BRF	Band reject Filter
BPF	Band-pass Filter
IAP	Inverse All-pass
DBTA	Differntially Bufferd Transconductnce Amplifers
CFTAs	Current Follower Transconductnce Amplifers
CCTA	Curent Conveyor Transconductnce Amplifers
DDCC	Differential Difference Current Conveyor
CB	Complementary Bipolar
MTC	Mixed Translinear Cell
FTFN	Four Terminal Floating Nullor

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ABSTRACT

New configuration of voltage-mode electronically tunable shadow inverse lowpass, shadow inverse highpass ,shadow inverse bandpass, and shadow inverse bandreject filters is proposed .The 4-terminal current-feedback operational-amplifier(CFOA) and passive elements are used to build the proposed filters circuits .Inverse filters are mostly used in correcting the distortions of signals in various instrumentation ,control and communication systems caused by transmission or signal processing systems. A shadow filter configuration have an external amplifier added in feedback loop of the basic inverse filter and gain of this amplifier can be used to control to tune the characteristics of the resulting filter. The parameters of the filters can be modified over a large range without disturbing the passive and active components of the filter itself by acting only from outside filter. Unlike current-mode shadow filters which mostly requires a summing junction, this voltage mode realization does not require summing amplifier. The passive and active sensitivities of proposed filters parameters is found to be low .The CFOA used for pspice simulation is commercial available current feedback amplifiers AD844 macromodel with ± 15 V voltage supply . The characteristics and validity of proposed biquad filters are verified by SPICE simulation.

Chapter

1. INTRODUCTION

With growing opportunities , there is more emphasis on very large scale integration(VLSI) in various range of application used in everyday lives such as DVD players ,cell phones ,personal computer ,digital cameras .Thus creating a challenge of realizing complex mixed-signal systems composing of both analog and digital subsystems in smaller area consuming low power and capable of working at high frequency .More emphasis is given on digital signal processing (DSP) because evolution in IC technology has given compact and efficient implemenatation of various types of digital signal processing applications on silicon chip and also the programming of DSP-based designs is easily realized and have minimum dependency for realization of most functions digitally on analog designers .However, natural signals usually encountered are analog, for instance video, audio ,sound ,picture and so on . In fact Amplification, rectification ,oscillators ,contionous-time filtering ,encoding(digital or analog conversion) at the input and equivalently decoding (analog or digital conversion) and smoothing using filters at the receiving end are impossible to be performed without analog signal processing.As a consequence the success of digital technique cannot make the analog design techniques to become outdated; rather it can only create inexhaustible opportunities and analog circuit designers faces new challenges .

1.1. Voltage and current mode signal processing

Conventionally ,voltage mode(VM) approach dominate analog systems ,where the current signals are being transferred into voltage domain. With the reduction of threshold voltage and power supply voltages ,the performance of voltage-mode circuits gets highly affected. Current-mode control was formally introduced to the power electronics world in 1978 [1]. It was quickly accepted as the most rugged way to control power supplies . Current mode approach implies that information being operated is in currents instead of traditional voltage mode ,which uses voltage.

Limited b.w at large close loop gain of voltage-mode operational amplifier due to its constant gain-b.w product leads to increase in the complexity of its design and not able to fulfill the requirement of low power consumption ,high b.w and low supply voltage simultaneously .So drift towards current mode approach is increasing. Since here the attention is on the currents in branch instead of voltage at particular node ,CM circuits are less influenced with the reduction of power supply as compared to their equivalent parts .As an outcome from the publication of limitless literature on CM circuits in recent past ,significant ASP and signal generating circuit are proposed .Due to immense growth in IC technology in last two decade ,CM analog techniques have quiet been exploited by circuit designers .Claiming to several advantages of CM mode such as large bandwidth , high operational speed ,superior signal linearity .A further advantage is of low voltage operation seeing as the design of CM circuits operate in the processing of current which could be obtained with lower voltage swing.

A well-known current mode circuit is current feedback operational amplifier (CFOA) having congenital advantage comparing to the traditional voltage operational amplifier cause of constant b.w with respect to closed loop gain unity gain, high linearity and better frequency performance and high slew-rate. It is also commercially available in integrated circuit form, namely AD844.2 stages of CFOA are current conveyor(CCII) and voltage follower which can be implemented using (CCII). The basic concept of current-conveying and its embodiment as current conveyorI was introduced by Smith and Sedra in 1968 [2]. Subsequently, this was referred to as the first generation current conveyor (CCI) and later on, in 1970, they introduced a more versatile current conveyor, named as second generation current conveyor (CCII) [3]. CC is a three terminal device, when connected with the other circuits externally can perform many analog signal processing purposes.

1.2. Inverse filter

There are frequent circumstances where altercation of electric signal through a linear or non linear transformation before processed or transmission and passed via a system executing the inverse of initial transformation such as , division and multiplication [63], pre-emphasis and de-emphasis in FM systems [62], logarithmic and exponential amplifiers occur. The literature survey of the inverse filter suggest that numerous type of digital inverse filter has been designed but analog inverse filter remain unexplored in this area because of limited availability of analog inverse filter circuit design [4,5]. However recent research trends that the area is now gaining a new interest.

In communication [6] acoustic systems, speech processing and control systems also, inverse active filters have significant role, in which they are utilized to amend distortion of signal initiated by signal processing or transmitting circuits [4], [8]. Modification is attained by utilizing inverse filter with freq response which is reciprocal of freq response of circut which is foundation of distortion. Ever since 2nd order are incredibly utilized in comunication systems in past decade, realisations of 2nd order inverse active filters likewise obtained momentous interest [9], [10].

1.3. Shadow filter

The authors Y. Lakys and A. Fabre In 2010 first projected a shadow filter [11].Shadow filters are novel class of analog filters of 2nd order. The scheme of this type of filter is established on the information that by acting on the external amplifier's external gain there will be shifting the characteristic frequency of the 2nd order filter. In this type of filter, by working on the external amplifier's gain a alteration in the center frequency of the band-pass filter will be done while at the same time the quality factor will settle its value to preserve a constant bandwidth. This attribute is accomplished with no modification in the passive and active components of the filter. This advance is very helpful in designing reconfigurable filters whose the center frequency can be varied over a broad frequency range. Furthermore, the hop amid 2 consecutive bands will be carried out extremely swiftly during the signal transmission. Adding up to that is the electronic fine tuning of the filter factors using this method can be very striking to compensate for the non ideal property. If the gain of the external amplifier is restricted by a DC current or voltage, the distortion in analog signal will be saved which results from switches as the switching is in the dc domain. Different from other tuning techniques in such a way like switched capacitors technique where the tuning is done via the switching frequency and also here the analog signal experience from the switching distortion. The scheme of the shadow filter can be extensive to 2 interesting characteristics depending on which output terminal is feeding the feedback amplifier. The 2 appealing characteristics are tuning of bandwidth(b.w) with tuning/fixing the center frequency. Moreover there can be tuning of the center frequency and bandwidth with fixing the quality factor.

1.4. Motivation

Active filters have inescapable use in communication, automatic control and instrumentation. The filter's application have large ranges such as, band pass filter of high frequency are used for selection of channels at the telephone central offices, data acquisition system require low pass filters, in audio amplifier high pass filter are used for signal separation, for suppressing the wave traps also known as interfering signals band notch filters are used .The simplicity and usefulness makes the study of second order active networks important. Oftentimes implementation of higher order filter can be done by cascaded network of second filters ,if required. The standard 2nd order general equation of biquad filter is given by:

$$\mathrm{T.F} = \frac{\mathrm{N(s)}}{\mathrm{s}^2 + \mathrm{s}\frac{\omega_0}{Q} + \omega_0^2} \tag{1.1}$$

Where

ω_0=undamped natural frequency

Q= quality factor

N(s)=real polynomial with possible order of 0,1or 2

Filter constituting of active component rather than passive component are requisite due to size of inductors and poor performance at lower frequency which restrict their performance**.** Diverse varieties of novel second order circuits have been issued in the literature.Then, for shortest possible response delay, it is mandatory to use active element. In other respects, it was demonstrated [12, 13] that frequency response of current mode circuits are superior that voltage mode circuits. Also regrettably there is commercial inaccessibility of the active elements VDTA, CDTA ,ZC-CITA and CCCII+ active which were used to implement shadow and inverse filter previously [11,21,22,30,60]. For all this reason, the active filter realized using CFOA. In the next module via characterization of this circuit for this application to show that it is impeccably suitable for frequency-agile active filters. Even then the circuit proposed is in voltage mode because as we will see in next module, the summing circuits for summing of feedback signals are required in shadow/agile filters in current mode. But in the proposed circuit, no such summing circuit is used in voltage mode realization.

In the field of DSP for inverse digital filtering, several techniques are already presented[14]; despite that for implementation of continuous-time analog inverse filters, very few circuits/methods are studied[4-5,9-11]. Nevertheless, a shadow filtering based inverse filter have not been stated till now to our information and same is the base of our proposed circuit.

1.5. Objective

In perspective of the above assessment, the disquisition has concentrated on the underneath objectives to be achieved for the propounded circuits:

- Literature survey to distinguish research gaps
- Study of different inverse filters circuits
- Study various techniques of shadow filters
- Design and development of propounded circuits
- Implementation and simulation of application

The essential focus of this dissertation was to design topologies that can perform inverse low pass, inverse band-pass filter, inverse band-reject filter and inverse high pass filter using shadow filter techniques. These employ CFOAs, passive elements such as resistors and capacitors. By means of work on the gain of external amplifier, the parameters of filters can be varied without modification into passive and active components of filter. Approach in this way is very beneficial into the creation of reconfigurable fiilters where over wide range of freq the centre frequency can be varied. Furthermore, during signal transmission hop between two successive bands can be easily carried. Additionally, filter parameter's electronic fine tuning using this method to offset the non-ideal effects can be easily achieved. To meet the requisite objectives, the proposed designs will be done with the help of PSPICE software to test their functionality.

1.6. Organization of Thesis

The entire thesis is partition into following chapters:

Chapter 1: This chapter venture into concise presentation in analog circuit design, problem in implementing CM and VM processing, brief overview of inverse filter and overview of shadow filters. The motivation and objectives of the dissertation are also discussed in this chapter.

Chapter 2: An extensive survey of topologies of shadow filter proposed by various authors have been discussed and compared within this chapter

Chapter 3: This chapter features reviews of a number of schemes published by numerous authors on inverse filter techniques have been included here.

Chapter 4: This division explore into CFOA structure, mathematic description of its inputs and outputs and its basic operation.

Chapter 5: This chapter comprises of fundamentals of shadow filter and its numerous schemes of implementations.

Chapter 6: The attributes of inverse filter schemes and its techniques have been included in this chapter.

Chapter 7: A novel structures of inverse shadow filters are propounded and various performance parameters are calculated.

Chapter

2. LITERATURE SURVEY OF SHADOW FILTER

2.1. Introduction

Linear circuits such as frequency filters have wide application in the area of electronics. These are also used as basic building blocks in analog signal processing. In the last decade, an enormous number of active building units for analog signal processing were presented. Nevertheless, there is still the requirement for progress in the development of new active elements that can offer new potentials and better parameters. Also the need for high performance low voltage circuits encourages the analog designers to look for new circuits structures and new techniques. There are two such techniques which are known as shadow filter and inverse filters. In this thesis combination of these techniques have been done to gain more advantages in application of active filters. A brief introduction and literature survey of them is described in this chapter.

2.2. Shadow filters

In 2010 [11] Y. Lakys and A. Fabre firstly presented the principle of shadow filter (also known as frequency-agile filter). Shadow filters are the recently addition to the family of second order filters. Its principle is found on altering characteristic frequency of the 2^{nd} order filter by controlling external amplifier’s gain. As expressed in [11] voltage-mode shadow filter is constructed by a 2nd order filter with 2 outputs namely, bandpass and lowpass, a voltage amplifier in feedback and a summing circuit for summing different feedback signals. Its Fig. 2.1is shown below:

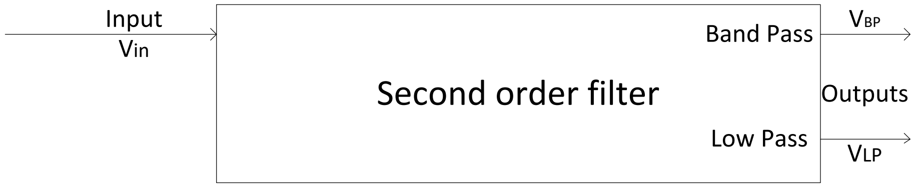

Fig. 2.1 Basic 2^{nd} order filter

$$F_{BP}(s) = \frac{V_{BP}}{V_{IN}} = \frac{a's}{1 + as + bs^2} \quad (2.1)$$

$$F_{LP}(s) = \frac{V_{LP}}{V_{IN}} = \frac{d'}{1 + as + bs^2} \quad (2.2)$$

Where a,b,a′,d′ constants are both real and positive.

The adjustability of external amplifier A's gain allows to modify the parameters such as characteristic frequency, quality factor or b.w without changing the main structure of 2nd order filter [11,1 2]. This filter can also be implemented current-mode circuit. Although summing circuit is simpler to implement in voltage-mode. In telecommunications or for requirement of additional trimming due to their perspective modification cause of non-ideal effects [5,11], shadow filter have found many application.

2.2.1. Review of shadow filter Implementations

In [11], the authors first presented the technique of shadow filter. A current-controlled current-conveyor (CCCII) based shadow filter in current mode is presented. It realizes bandpass filter having variable center frequency and quality factor using 4 CCCIIs so that constant b.w is found.

The method to control the quality factor and b.w orthogonally without disturbing the central frequency presented in [11] was further extended by the authors in [60]. The proposed current-mode bandpass filter consist of uses two Z-copy current inverter transconductance amplifier (ZCCITA) and two current amplifiers. Still, to get bandpasss output a current mirror is needed.

In [15], the authors additionally extended idea proposed by the same authors in [11] by generalizing the shadow filter to n-class shadow filter. Here a CCCII-based shadow filter was presented whose center frequency was $f_{oA} = f_o\,(1 - Ad\,')^{n/2}$.

The author in [16] gave a feasible solution in voltage-mode shadow filter to the case of constant value of the product of the quality factor and center frequency by taking feedback from highpass output and the case of constant center frequency with variable quality product by taking feedback from the bandpass filter. Furthermore, the author produced a problem about the possibility of having constant Q with a variable center frequency. There was no real experimental implementation was presented.

In [17]-[20], practical application of shadow filter of different classes based on current-mode CCCII were presented. Current-mode shadow filters using the voltage difference transconductance amplifier (VDTA) and the current difference transconductance amplifier (CDTA) of different classes were presented in [21] and [22].

In [23-25], employing commercially available integrated circuit current feedback operational-amplifiers (CFOA) diverse variety of voltage-mode shadow filters were published. Instead of traditional one input and multi-outputs, here 2^{nd} order shadow filters have multi input and single output. Also, the summing function is done internally and due to grounded active element input or passive component, a summing input is produced.

2.3. Inverse filters

Inverse-filters are essential in frequent applications in communication, signal processing, control and instrumentation systems. Few instances such as to observe the signal's relative cleanness [6] and to do the function of de-emphasis in FM systems [62].

There are frequent circumstances where altercation of electric signal through a linear or non linear transformation before processed or transmission and passed via a system executing the initial transformation's transformation such as , division and multiplication [8], de-emphasis and pre-emphasis in FM systems [62], logarithmic and exponential amplifiers occur. Despite the fact that the problem of inverse filtering in digital filtering can be easily done, continuous time circuit didn't have general method for inverse filter till 1997.In 1997, A. Leuciuc proposed for the first time a scheme to achieve the inverse transfer characteristic for nonlinear resistive circuits and the inverse transfer function for linear dynamic systems. This method is presented using nullors as basic building blocks.

2.3.1. Review of inverse filter Implementations

Inverse filter technique was first introduced in [4] in1997.For linear dynamic system and non-linear resistive circuits, author first put forward inverse transfer function and inverse transfer characteristic by using 2 port active circuit element called nullor. This technique is described by employing Sallen key highpass biquad filter and half-wave rectifier in inverse form.

The authors proposed [5] to transform RC filter employing VM op-amp into CM 4-port floating nullor(FTFN by Senani in 1987) was proposed. Technique in [4] and RC-CR dual transformation procedure (because of which this method limits to planar circuit) is employed.

To remove the limitation of [5], a alternative procedure is described in [10] by acquiring CM inverse filter based on FTFN from VM filter. This makes it applicable to non-planar circuit also. The cascadilibity charactersic of Friend-Delyannis biquad is further checked.

Different type of inverse filter based employing active element FTFN with the help of CFOA are published in [26-29].In[26],author used one FTFN to derive inverse-LPF,BPF,BRF,HRF using approach of Senani[7]. In [27, 28] every circuit realizes one particular inverse filter transfer function. But by only choosing relevant topology admittance, [29] could produce inverse-LP,HP,BP,BR transfer function. Since FTFN is not commercially available, it can be synthesized using two CCII-s or two CCII+s (or two CFOAs which consists of a CCII+ and a voltage buffer). Such a synthesis was first proposed by Senani in [7]. Authors of [30, 31] implemented inverse all-pass (IAP) filters employing current difference transconductance amplifier (CDTA) and current conveyors, correspondingly.

2.4. Concluding remarks

In this chapter brief introduction and literature survey of two techniques known as shadow filter and inverse filter is being described. These techniques are together used in this thesis for the designing of different types of filters as shown in subsequent chapters.

Chapter

3. REALIZATION OF CFOA

3.1. Introduction

Numerous active devices for ASP have recently been published. To name few examples of these topologies published in literature, for instance, differntially bufferd transconductnce amplifers (DBTAs) [31], curent differncing transconductnce amplifers (CDTAs) [32], current followr transconductnce amplifers (CFTAs) [33], curent conveyor transconductnce amplifers (CCTAs) [34-38], differntial differnce curent conveyor (DDCC) [39-40] and so on. Regrettably among most of them are produced via simulation programs whose active elements consist of transistor model of the either bipolar or CMOS technology having questionable practical feasibility.Experimental verification by means of their on-chip fabrication is costly and time-consuming [41].

Amongst the above mentioned active components, an interesting active component called the current feedback operational amplifier (CFOA) is present. It is particularly appropriate for a group of ASPs [42-44]. Operation of this topology is possible in either current and voltage modes, also offering flexibility allowing a wide range of circuit designs. Furthermore, it offers beneficial features, for instance high slew rate, independence from parasitic capcitances, wide b.w and uncomplicated implementation [45- 48].

Current feedback op-amps (CFOA) attracted the interest of the researchers and designers of analog circuit once it was recognized that the designed amplifiers exhibits properties which for the most part are noteworthy different from the properties shown by renowned voltage operational amplifier (VOA) based implementations. In VOA, there are inevitable conflict of gain-band-width product whereas circuit employing CFOA can produce gain varying constant bandwidth topologies.Additionally, cause of greatly elevated slew-rates of the order of numerous hundred to thousand V/μs (which can be as large as 9,000 V/μs for modern CFOAs) comparing to a very diffident 0.5 V/μs for the widely available commercial μA741 IC of VOA, CFOAs could implement circuits which are able to work over a large extent of frequency ranges than those achievable with VOAs.

3.2. CFOA-Current Feedback Operational Amplifier

The CFOA is a four terminal device as shown in Fig. 3.1. The input and output terminals are represented by X, Y and O, Z, respectively.

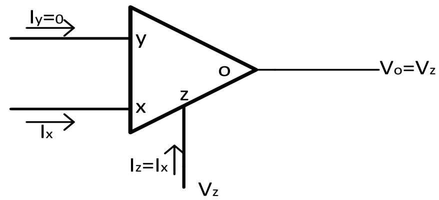

Fig 3.1 Current Feedback Operational Amplifier (CFOA) [49]

The hybrid matrix describing the behaving of CFOA[50] is :

$$\begin{bmatrix} Iy \\ Vx \\ Iz \\ V_0 \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} V_Y \\ Ix \\ V_Z \\ I_0 \end{bmatrix} \qquad (3.1)$$

As revealed in Fig.3.1 and above matrix equation 3.1, the potential voltage Vy which applied at port Y is being conveyed to the terminal X, denoted by Vx. Furthermore, when a current source is applied at port X, identical current is obtained at terminal Z. The output voltage Vo acquired at terminal O corresponds to voltage Vz when applied at terminal Z. In CFOA, both current mirror deed and voltage buffering deed occur. The impedance at input and output terminal of voltage follower should be infinite and zero respectively so as to accomplish the ideal feature of CFOA. Hence resistance at port Y ought to be infinite and input resistance at port X ought to be zero ideally. In the same way, resistance at the compensating port Z ought to be infinite seeing that it is the output excitation intended for the 2nd voltage follower which is appear between terminal O and Z and output port for the current follower appear amid terminals Z and X. Terminal X and Y are inverting and non-inverting terminal correspondingly for the amplifier.

The interior representation of CFOA[51] is revealed in Fig. 3.2. The output of the input follower is conected to Inverting port of CFOA , as a result its input resistance become equal to the buffer's output impedance and have low rate of impedance. Non-inverting terminal of the amplifier operate as an input terminal for the input follower and hence making the terminal to have high impedance.

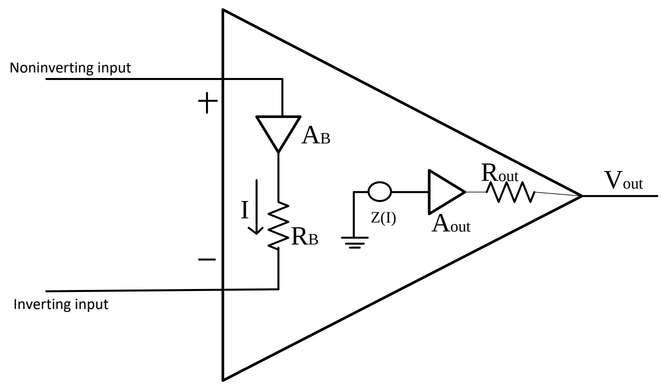

Fig. 3.2: Current Feedback Operational Amplifier (CFOA) model [51]

The gain of input buffer is represented by A_B which is nearly unity. R_B is representing the output impedance of the input follower. The resistance of output buffer corresponds to Rout and Aout corresponds to gain of the output buffer, which is as well close to unity. At higher values of frequencies, the output impedance of the input follower affects the stability of the CFOA. To accomplish further accuracy of amplifier, using closed feedback loop the value of the trans-impedance is usually made very high [51].

3.2.1. The CFOA including externally-available compensation Pin: AD844

Even though due to CFOA's popularity numerous integrated circuit manufactures have manufactured them as IC, their IC usability can be divided into two types. One of them falls into the category of CFOAs having same pin-compatibility as compared to VOAs without having externally available pin for compensation. In contrast, the other kinds have the alternate choice of including a pin for compassion (usually numbered as pin no.5) which is externally available at the same time preserving pin-compatibility with VOAs. Later ones are recognized as AD844 in analog devices. The substitution of AD844 with conventional op-amp is possible except that the current feedback structural design have resulted in much enhanced ac performance, large linearity and an extraordinary clean pulse response. Typical illustration of implementation suggested by the manufacturers include Flash ADC input amplifers, Video buffers, rapid speed curent DAC interfaces and cable drivrs and pulse amplifers.

3.2.2. General description of AD844

AD844 as of analog devices is a high speed monolithic operational amplifier has been fabricated with process called junction-isolated complementary bipolar (CB).It unite high b.w and very rapid large signal response with brilliant dc performance. It normally give elevated slew-rates of the order of 2,000 V/μs that is, rate of chang of peak outpt can go over 2000 V/μs for a complete 20 V outpt step voltage . Even though it has been optimized to use in applications requiring current to voltage conversion and like amplifier called inverting amplifier, it can have satisfactory application in non-inverting and others. This type of op amp yields a closed-loop b.w namely dominated mainly by the resistor of feedback and nearly free from the gain of closed-loop. The inpt bias curents and offset voltge of the AD844 are laser- trimed to lessen DC errors so that drift in the ofset voltge is usually 1 μV/C and drift in bias current drift in the order of 9 nA/C.

The internal architecture of AD844 as of analog device is formed by a translinear 2nd generation Current conveyor (CCII+) formed by a 4 transistor-mixed translinear cell and subsequently a voltage buffer (translinear) as shown in Fig 3.3[52]

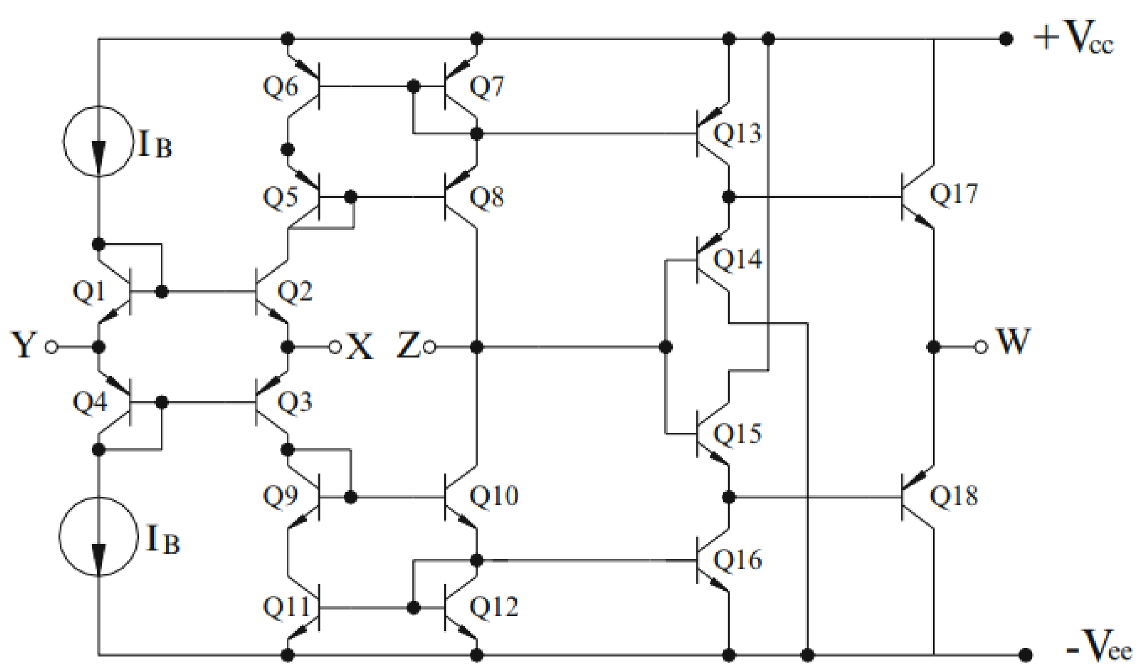

Fig.3.3 Bipolar realization of AD844-CFOA Current feedback operational amplifier [52]

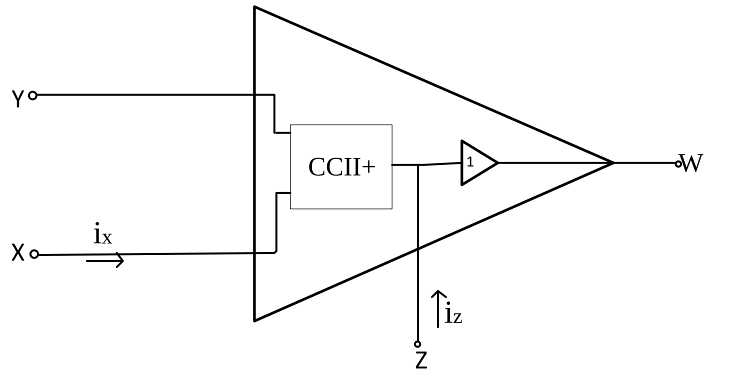

Fig. 3.4 A symbolic diagram of the interior architecture of AD844 CFOA[52]

Its simplified simplified diagram showing this identification is shown in Fig. 3.4.

Transistors Q1 to Q4 in the interior architecture of CFOA as shown in Fig.3.3 form an mixed translinear cell(MTC),where currents in collector of transistors Q3 and Q2 are sense due to 2 modified wilson n-p-n and pnp currents mirrors. This current mirrors are formed by transistors Q9to Q12 and Q5 to Q8 respectively .These 2 invariable(constant) current sources, both equivalent to I_B ,which forces identical emitter currents in transtor Q4 and Q1 in such a manner that when voltage Vy is applied at the terminal Y makes Iy=0.MTC creates reproduction of currents Ix at the Z terminal thus yield Iz=Ix. By seeing above Fig 3.3, one can derived that if Ix=0,Vx=Vu then current at Z terminal become zero. To find the value of currents in x and z terminal when current at x port is not zero, accurate study of the design is done in [53].This is done by employing the exponential relationship amid base-emitter voltages and current in collector for the transistors from Q1 to Q4 which give following equation:

$$I_z = I_x = -2I_B \mathrm{Sinh}\left(\frac{V_y - V_x}{V_T}\right) \tag{3.2}$$

from above , for (Ix < < 2*I_B) an estimated relation between Vy, Vx and rx can be calculated as shown:

$$V_x \approx V_y + r_x i_x \qquad \text{where } r_x = \frac{V_T}{2I_B} \tag{3.3}$$

After terminating Z port with externally applied impedance Z_{load}, voltage named Vz at port Z is formed. After passing this voltage via voltage follower, identical voltage is obtained which is named Vw .This follower is consist of a different MTC made from transistors Q13 to Q18 where transistor Q13 and transistor Q18 supply the DC biasing currents. The characteristic of final stage of CFOA can be described by equation similar to eq 3.3 giving Vw=Vz.

Therefore, the 4 characteristic equation of Ad844 CFOA describing its properties are as revealed below:

$$iy = 0 \tag{3.4}$$

$$vx = vy \tag{3.5}$$

$$iz = ix \tag{3.6}$$

$$vw = vz \tag{3.7}$$

As revealed in the Fig. 3.3and 3.4 above ,the interior architcture of AD844 composed of a CCII + succeeded by a voltage buffer, this adaptability of AD844 was afterward found to be helpful in enabling to generate a CCII + and CCII- (using two CCII+), because of the pin-by-pin substitution of a voltage operational amplifer(with Z-pin kept open) and additionally as a 4-terminal buildng block on its own.

3.2.3. Conventional Current Conveyor(CC)

Sedra and Smith [2] first proposed CCII, a three-terminal device basic building device in 1968. Yet its real advantages and ground-breaking impact was not apparent at that period. However, class of CM come forward as an important blocks because of characteristic as such of accuracy, large range of frequency range, and versatility in an extensive range of application. It has wider b.w, simple architecture and capacity to be able to work at low voltage. Hence it can be used in numerous application ranging of oscillators design and immitance design, universal filters to differentiator circuits and integrators circuits.

The CC are categorized into 3 generations known as:

First generation current conveyor (CCI)

Second generation current conveyor (CCII)

Third generation current conveyor (CCIII)

Above 3 generations of current conveyors can be explicating with the aid of matrix [54]:

$$\begin{bmatrix} V_x \\ I_Y \\ I_z \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 \\ m & 0 & 0 \\ K & 0 & 0 \end{bmatrix} \begin{bmatrix} I_x \\ V_Y \\ V_z \end{bmatrix} \tag{3.8}$$

Where,

m = 1 for 1st generation of current conveyors

m = 0 for 2nd generation of current conveyors

m = -1 for 3rd generation of current conveyors

with K = 1 designed for positive current conveyors (CC+) and K= -1 designed for negative current conveyors (CC-) [54].

The structure of 3 generations is similar other than their characteristic. The black box representation of CC is revealed in Fig. 3.5 consisting of 3 terminals X, Y and Z.

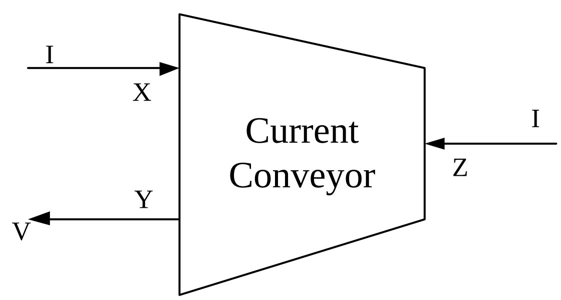

Fig 3.5 Black box representation of CC(Current conveyor)[55]

- Terminal X be a hybrid terminal which for current signal functioning as an input port while for voltage signal, functioning as output terminal.
- Terminal Y is a port for voltage input.
- Terminal Z be a current output port which could either source or sink sinal of current equivalent to the current inserted into terminal X.

If voltage source is injected into terminal Y it follows that identical voltage potential will be emerge on port X. If current source is injected into terminal X it follows that identical current will flow in terminal Y depending on the type of CC.The same current is also transferred to high impedance output terminal Z.

Hence it can be said that CC guarantee the following 2 functionality amid its terminals:

- Between terminals X and Y, Voltage buffer action happens.
- Terminals X and Z, current buffer action happens.

Ideally the impedance at terminal X of CC should be kept as minute as achievable seeing as terminal X output voltage terminal and current input. Likewise terminal Z's output resistance should be devised as large as achievable seeing as it is a current input terminal[6].Furthermore, CC should be able to perform voltage and current action over wide range of frequency .The designing of current mirrors should be done as precise as possible for the voltage and current buffering action to take place to make sure that appropriate mirroring action with minimum offset can occur.

Smith and Sedra [2] in 1968 proposed First generation current conveyor (CCI±) as the foremost basic current mode block as revealed in Fig.3.6. The voltage and the current at terminals Y and X, becomes alike, as a result the impedance level at terminal Y should be low, ideally zero for precise current and voltage conveying [2].

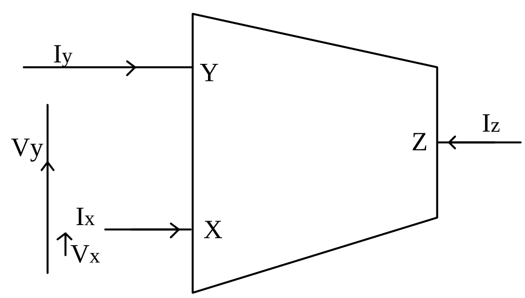

Fig. 3.6 First Generation Current Conveyor [55]

Another block was developed by Sedra and Smith in 1970 as revealed in Fig. 3.7 (recognized as second generation current conveyors (CCII±) [3]. Idyllically in CCII± , no amount of current flows via the Y port. Consequently, the resistance at the Y port is infinite.

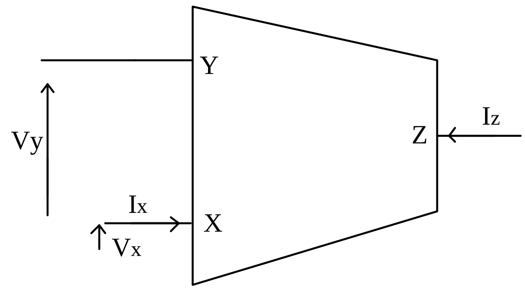

Fig. 3.7 Second Generation Current Conveyor [55]

Third generation current conveyor (CCIII±) was proposed by A. Fabre in 1995 [56] as revealed in Fig. 3.8.In CCIII the current in terminal Y and X are identical, flowing in the opposite directions to each other, that is Iy=-Ix .

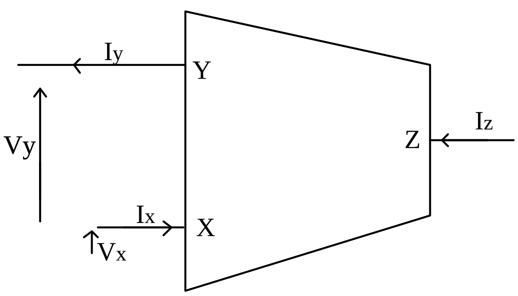

Fig. 3.8 Third Generation Current Conveyor [55]

3.2.4. Second generation current conveyor

One of the most multipurpose CM building component is the second generation current conveyor. From the time its proposal in 1970, there are numerous application of this component. This component is a 3 terminal device with its relationship between port currents and voltages defined by the subsequent hybrid matrix (3.9) as:

$$\begin{bmatrix} Vx \\ Iz \\ Iy \end{bmatrix} = \begin{bmatrix} \beta & 0 & 0 \\ 0 & \alpha & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} Vx \\ Ix \\ Vz \end{bmatrix} \quad (3.9)$$

Where *Y* and *X* are the input terminals and the output terminal is *Z* .

The current and voltage tracking error in CCII is represented by α = 1-εi and β = 1-εv with εv and εi ($|\varepsilon v|$<<1 and $|\varepsilon i|$ <<1), correspondingly. In CCI, Y and X terminals ideally have zero input impedance making it to sink currents, leading to minimizing its flexibility and adaptability.

Consequently there is limitation in the application of Current Conveyor I. This contributing to the inception of 2nd generation CC named as CCII. The Y-terminal is a high resitance port for VM operation and terminal X perform as low resistance terminal for applied input current and output voltage (this terminal is perfect for CM blocks) . Z-port has elevated output resistance for CM structure.

In order to get ideal properties, 2nd generation CCII should be characterized high resitance on terminals *Y* and *Z* and by low impedance on terminal *X*.

In 2nd generation CCII depending on direction of current between terminal Z and X, it is divided into 2 type:

- Positive current conveyor CCII- in this Ix and Iz are having same direction of current as of current mirror.
- Negative current conveyor CCII- in this Ix and Iz are having opposite direction of current as of current mirror.

These 2 CCs are shown below as in Fig. 3.9:

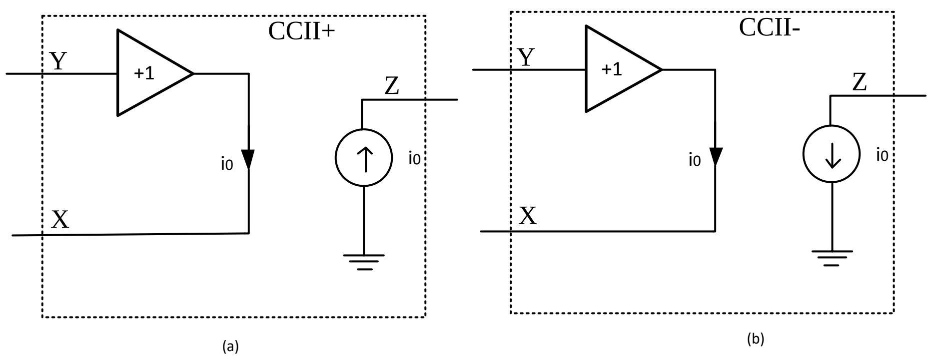

Fig. 3.9(a) Positive current conveyor CCII+(b) negative current conveyor CCII-[55]

The level of resistance at various terminals of CCII+ is revealed in the table 3.1.

CCII Ports	**Impedance Level**
X	Low (ideally zero)
Y	High (ideally infinite)
Z	High (ideally infinite)

Table 3.1 Level of resistance at the ports of CCII[55]

Ideally in the CC, the rate of value of gain of voltage and current is unity. Except in actual practice there are usually some deviations from the ideal characteristics of CC. Fig.3.10 reveal a non-ideal/real model of CCII where β and α (close to unity) parameters represents the non-ideal current and voltage follower characteristics.

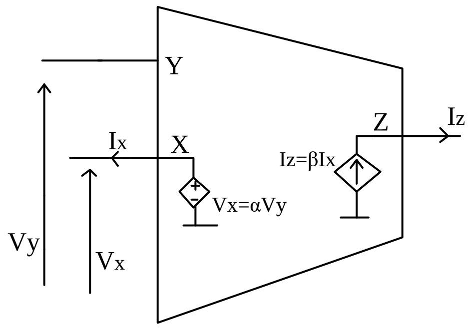

Fig 3.10 Non-ideal or real model of CCII [57]

3.2.5. Voltage Buffer or Follower

Primarily a voltage buffer without distortion or attenuation of the signal, functions as voltage conveyer from a high-impedance source to a low-impedance load. Ideally the input impedance is infinite and output impedance is zero of a voltage follower. These are created by means of high-gain op-amp with negative feedback as shown in Fig. 3.11[58].

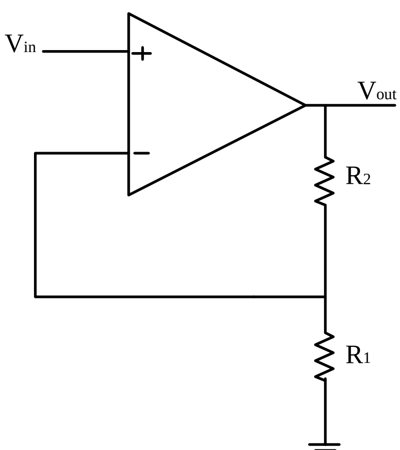

Fig.3.11 Buffer configuration of Op-amp [58]

As already been described in above section, the output stage of the CFOA is voltage buffer. And it conducts the voltage from the large input resistance terminal to the small input resistance ports. Voltage followers are designed commonly with common drain configuration, differential amplifier and flipped voltage follower [59].

3.3. Concluding remarks: In this chapter CFOA structure, mathematic description of its inputs and outputs and its basic operation is discussed. The internal structure of CFOA IC known as AD844 is described. Moreover description of two stages of CFOA which are current conveyor and voltage follower is also mentioned.

Chapter

4. SHADOW FILTERS

4.1. Introduction

The authors Y. Lakys and A. Fabre In 2010 first projected a shadow filter [11].Shadow filters are novel class of analog filters of 2nd order. The scheme of this type of filter is established on the information that by acting on the external amplifier's external gain there will be shifting the characteristic frequency of the 2nd order filter.

Consider a block of 2nd order filter :

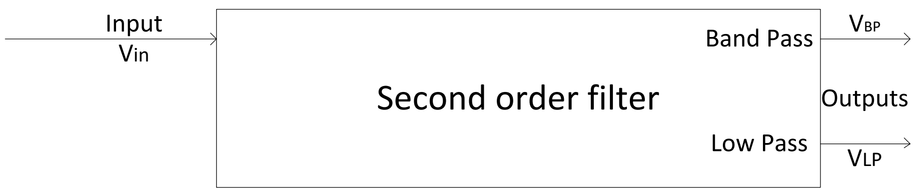

Fig. 4.1 Basic biquad filter [11]

The transfer function is given by:

$$F_{BP}(s) = \frac{V_{BP}}{V_{IN}} = \frac{a's}{1+as+s^2} \quad (4.1)$$

$$F_{LP}(s) = \frac{V_{LP}}{V_{IN}} = \frac{d'}{1+as+s^2} \quad (4.2)$$

Noting that every one of the constant a, b, a' and d' are both real and positive. Now if the output of low pass is being amplified by gain A of external amplifier which is then added to the external input V_{IN} so that the $V_E = V_{IN} - A*V_{LP}$ as revealed in Fig. 4.2:

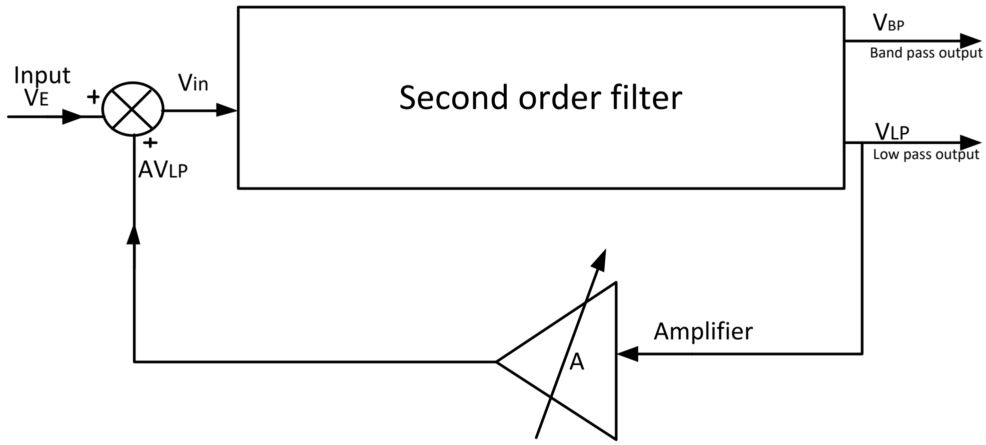

Fig.4.2 Biquad shadow filter [11]

The new transfer function of the bandpass filter is specified by Equation(5.3):

$$\mathrm{F_{BPA}(s)} = \frac{\mathrm{V_{BP}}}{\mathrm{V_E}} = \frac{\frac{a's}{1 - Ad'}}{1 + \frac{as}{1 - Ad'} + \frac{bs^2}{1 - Ad'}} \quad (4.3)$$

The dissimilarities between the shadow filter and the basic 2nd order filter in terms of the various coefficients of filters is shown in table 4.1:

	Basic 2nd order filter(Fig. 4.1)	**Shadow filter(Fig. 4.2)**
Center frequency	$\mathrm{fo} = \frac{1}{2\pi\sqrt{b}}$	$\mathrm{f}_{oA} = \mathrm{f}_o\sqrt{(1 - \mathrm{Ad'})}$
Q-factor	$\mathrm{Q} = \frac{\sqrt{b}}{a}$	$\mathrm{Q}_A = Q\sqrt{(1 - \mathrm{Ad'})}$
Bandwidth	$\mathrm{BW} = \frac{a}{2\pi b}$	$\mathrm{BW}_A = \mathrm{BW} = \frac{a}{2\pi b}$
B.P gain	$G_{BP} = \frac{a'}{a}$	$G_{BPA} = G_{BP}$
L.P gain	$G_{LP} = d'$	$G_{LPA} = \frac{G_{LP}}{\sqrt{(1 - \mathrm{Ad'})}}$

Table 4.1 Filter characteristics of Fig. 4.1 and Fig. 4.2

Its apparent from Table 4.1 that the tuning of characteristic frequency can only be done by the external amplifier's gain A, but provided that the term ($1 - Ad'$) remain positive. Fig. 4.3 illustrate the shadoe filter's characteristic freq. of the shadow filter vs. the amplifier's gain A.

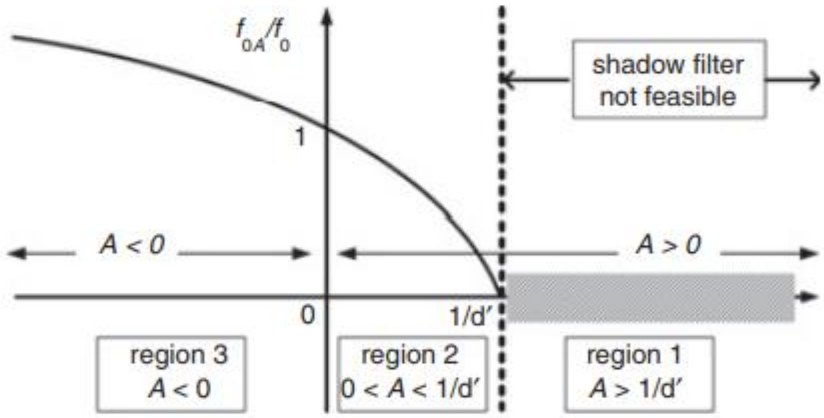

Fig. 4.3 f_{0A}/f_0 variation as a function of gain A[11]

As revealed in Fig. 4.3 the variation of f_{0A}/f_0 with gain A can be divided into 3 sections. The 1st section is not possible for shadow filter. In 3rd section, the shadow filter's characteristic frequency f_{0A} is greater comparing to basic filter's frequency. On the other hand, in 2nd section when amplifier's gain is less than 1/d', f_{0A} will be smaller than f_0.

4.2 Nth Order Shadow Filters:

The authors additionally extended idea proposed by the same authors in [11] by generalizing the shadow filter to n-class shadow filter[15]. Here a CCCII-based shadow filter was presented whose center frequency was $f_{oA} = f_o\,(1 - Ad\,')^{n/2}$.

Below Fig. 4.4 reveal the shadow filter of class-1, where the basic biquad filter is at that instant entitled as clas-0 shadow filter.

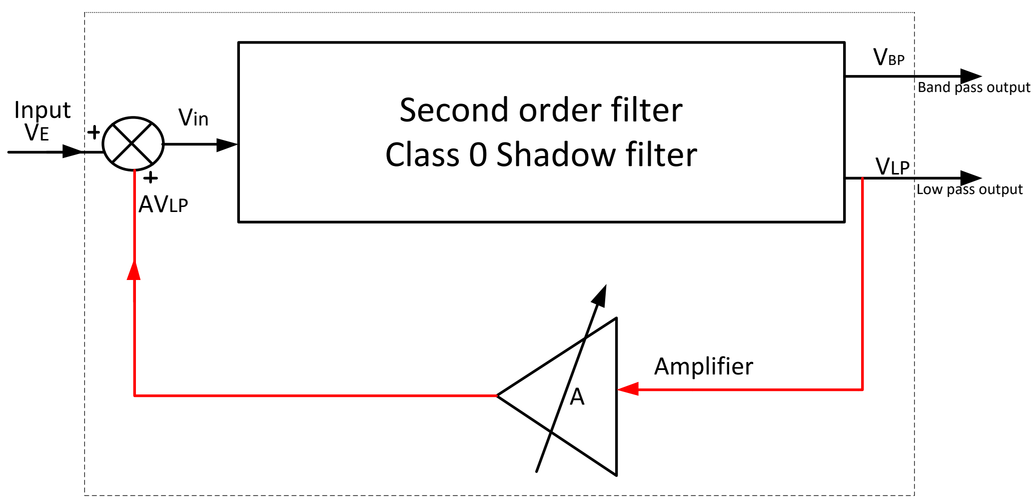

Fig. 4.4: Shadow Filter of Class 1 [1]

Once the amplification of signal of LP is done by number N amplifiers and when added/summed with i/p signals gives realization of shadow filter of Nth order. On behalf of every amp in the feedback, an amp having gain of (1-Ad') must be present in o/p of the LP so as to recompense for reduction in the low pass's gain (GLP) demonstrated in Fig.4.5

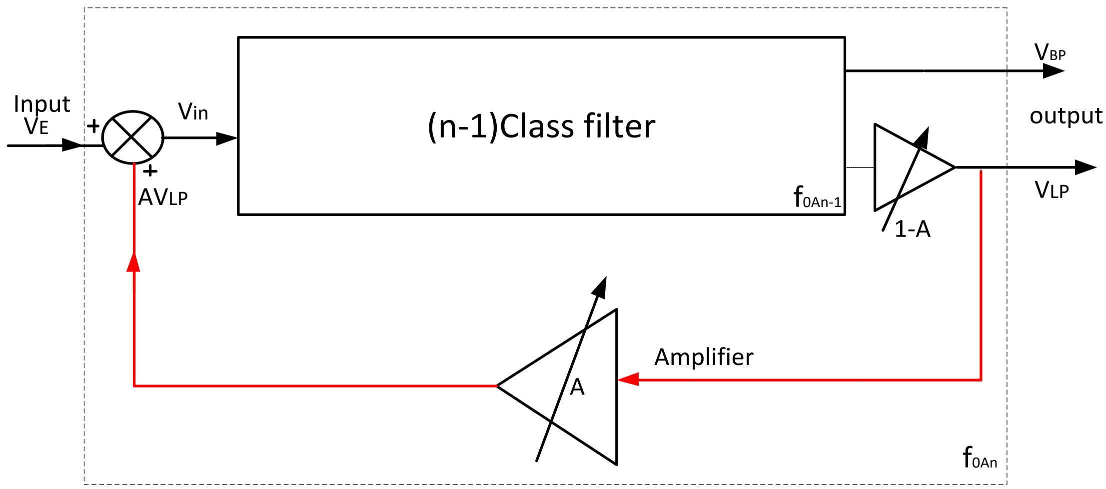

Fig. 4.5 Shadow filter of class-n [15]

Table 4.2 reveals the comparison of the nth order shadow filter parameters opposite to the starting basic filter.

	Basic 2nd order filter(Fig. 4.1)	Nth order Shadow filter(Fig. 4.5)
Center frequency	$\mathrm{fo} = \frac{1}{2\pi\sqrt{b}}$	$\mathrm{f}_{oAn} = \mathrm{f}_o(1-\mathrm{Ad}')^{\frac{n}{2}}$
Q-factor	$\mathrm{Q} = \frac{\sqrt{b}}{a}$	$\mathrm{Q}_{An} = Q(1-\mathrm{Ad}')^{\frac{n}{2}}$
Bandwidth	$\mathrm{BW} = \frac{a}{2\pi b}$	$\mathrm{BW}_{An} = \mathrm{BW}$
B.P gain	$G_{BP} = \frac{a'}{a}$	$G_{BPAn} = G_{BP}$
L.P gain	$G_{LP} = d'$	$G_{LPA} = \frac{G_{LP}}{\sqrt{(1-\mathrm{Ad}')}}$

Table 4.2: Filter Characteristics of shadow filter of class-n and basic 2nd order filter

The variation of of f_{0An}/f_n vs. the external amplifier's gain in support of numerous values of n (the order of the shadow filter) is revealed in Fig. 4.6. As revealed in fig 2, the variation of f_{0An}/f_n with gain A can be divided into 3 sections. In the 1st section, shadow filter is not possible when

A>1. In 3rd section, the shadow filter's characteristic frequency f_{0A} is greater than f_o when gain A is negative and when value of n increases its progress is faster. Whilst for n=2, the evolution was a linear function of A.

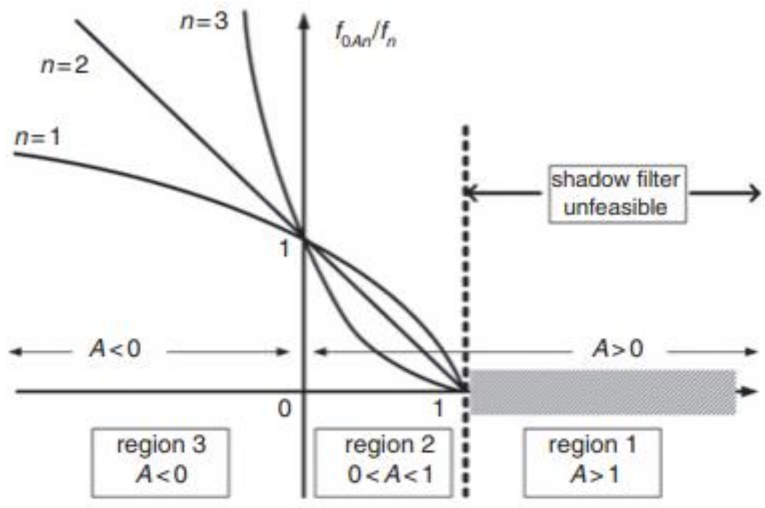

Fig. 4.6 foAn / fn versus A ; for several values of n [15]

4.2.1. Frequency Agile Filters (FAF):

According to the authors [17], the FAF is a reconfgurable filter in which throughout the signal's transmission, the leap betwen 2 bands (from f_1 to f_l) can be accomplished exceedingly quickly.

Presuming that the center frequency is transferable between 2 values: f_{OMin} and f_{OMax}, the range of transferable is well-defined in [17] by way of f_{OMax} and f_{OMin} ratio:

$$n = \frac{foMax}{foMin} \qquad (4.4)$$

The tunable filter in [17] is that type of filter which recompensate on behalf of the deviation in the central freq f_o owing to not having ideal properties and more factors by tuning the f_o around itself. In other words: $f_{0Max} < 2*f_{0Min}$. whereas the reconfigurable filter is that type of filter in which its main freq is estimated so as to be tuned across a wide-ranging frequencies. Implying that: $f_{0Max} > 2*f_{0Min}$.

4.2.2. Expansion of the Shadow Filters Theory to VM circuit

The method suggested in [11] is revealed in Fig. 4.7 with the help of a generalized block diagram [16]. The H(s) and H1(s) are regard as 2^{nd} ordr LP and BP trnsfer functons in [11], correspondingly with identical denominators. As revealed in fig. 1 after amplification of the LP output, it is then aded towards the externaly applied volt Vin.Corresponding LP and BP filter transfer functons which are founded named as H_1'(s) and H'(s) and will preserve their filter function type except for different characteristics.

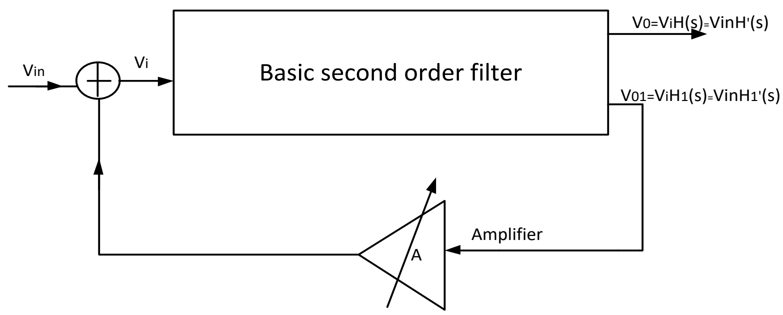

Fig. 4.7 Schematic diagram of the shadow filter [16].

The 2^{nd} order filter block shown in Fig. 4.7 is able to implement every 4 possible transfer function of low pass, high pass, band reject, band pass. The transfer function H(s) and H1(s) can be articulated as (1) and (2) respectively:

$$H(s) = \frac{d + es + fs^2}{s^2 + s\left(\frac{\omega_0}{Q}\right) + \omega_0^2} \tag{4.5}$$

$$H_1(s) = \frac{a + bs + cs^2}{s^2 + s\left(\frac{\omega_0}{Q}\right) + \omega_0^2} \tag{4.6}$$

$H'(s)$ can be determine from Fig. 4.7 as

$$H'(s) = \frac{H(s)}{1 - AH_1(s)} \tag{4.7}$$

Replacing H(s) and $H_1(s)$ in equation 4.7 gives:

$$H'(s) = \frac{d + es + fs^2}{s^2(1 - Ac) + s\left[\left(\frac{\omega_0}{Q}\right) - Ab\right] + \omega_0^2 - Aa} \tag{4.8}$$

Above can also be written as

$$H'(s) = \frac{d' + e's + f's^2}{s^2 + s\left[\left(\frac{\omega_0'}{Q'}\right)\right] + {\omega_0'}^2} \quad (4.9)$$

Where

$$(d', e', f') = \frac{d,e,f}{(1-Ac)} \quad (4.10)$$

$$\omega_0' = \sqrt{\left(\frac{(\omega_0^2 - Aa)}{(1 - Ac)}\right)} \quad (4.11)$$

And

$$Q' = \sqrt{\frac{(1 - Ac)(\omega_0^2 - Aa)}{(\omega_0 - AbQ)}} \quad (4.12)$$

The coefficients ω_o’ and Q_o’ are representing the quaity fctor and the characteristic freq ofshadow filter respectively .

Now just as in [11] the following cases are considered.

Case1: If LP’s output is taken for the feedback then with the choices b = 0 and c = 0 equation (4.11) and (4.12) will result in

$$\omega_0' = \omega_0\sqrt{\left(1 - \frac{Aa}{\omega_0^2}\right)} \quad (4.13)$$

$$Q' = Q\sqrt{\left(1 - \frac{Aa}{\omega_0^2}\right)} \quad (4.14)$$

above shows that the quality factor and centre frequency transform proportionately except that the b.w $\Delta\omega' = \omega_0'/Q_0'$ remains the similar as the original b.w $\Delta\omega = \omega_0/Q_0$.

Case 2: By applying a = 0 and b = 0, the transfer function of $H_1(s)$ come out to be high pass biquad filter

$$\omega_0' = \frac{\omega_0}{\sqrt{(1 - Ac)}} \quad (4.15)$$

$$Q' = Q\sqrt{(1 - Ac)} \quad (4.16)$$

For this case, the product of ω_o' and Q_o' remains constant and lower Q factorat higher freqs and vice-versa could be founded.

Case 3: If bandpass's output is taken for the feedback then parameter are taken as c = 0 and a = 0 providing:

$$\omega_0{}' = \omega_0 \tag{4.17}$$

And

$$Q' = \frac{Q}{\left(1 - \frac{AbQ}{\omega_0}\right)} \tag{4.18}$$

Here we obtain a fixed center frequency with varying Q factor with fixed center frequency. The resultant b.w is given by:

$$\Delta\omega' = \Delta\omega\left(1 - \frac{Ab}{\omega_0}\right) \tag{4.19}$$

The sensitivities with respect to external gain A of ω_o' and Q_o' by equation (4.13) and (4.14) is calculated to examine the effect of gain A and is found out to be:

$$S_A^{\omega_0{}'} = S_A^{Q'} = \frac{aA}{2\left(1 - \frac{aA}{\omega_0^2}\right)} \tag{4.20}$$

As can be seen from above equation, A is having negative value which implies that sensitivities of ω_o' and Q_o' would not cause whichever influence towards the shadoe fiter's main parameters.

4.2.3 New extension of the Shadow Filters

The authors in 2016 [23] proposed a innovative method for the shadow filter as shown in Fig.4.8 which can be thought as a new extension to the shadow filters theory.

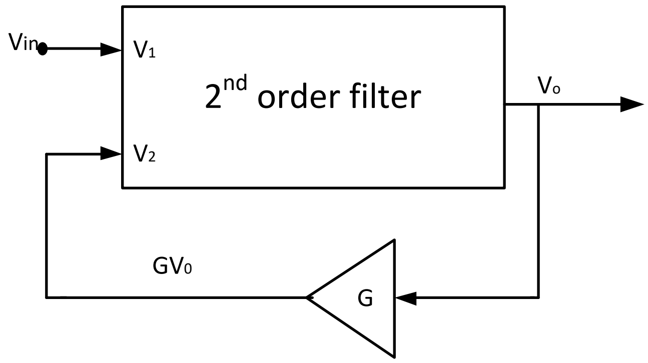

Fig. 4.8: Novel type of class-1 shadow filter[23]

Theoretical approach: Taking a second order filter of dual-input single-output as revealed in Fig.4.9

Fig. 4.9: Dual-Inputs Single Output biquad Filter[23]

Presuming a, b, G, A and B, are both real and positive, the transfer functions is given by :

$$V_o = \frac{AaV_1 + BsV_2}{as^2 + bs + 1} \tag{4.21}$$

$$V_o = \frac{AsV_{in} + GBV_o}{as^2 + bs + 1} \tag{4.22}$$

$$V_o \left[1 - \frac{GB}{as^2 + bs + 1}\right] = \frac{Aas}{as^2 + bs + 1} \tag{4.23}$$

$$V_o = \frac{AsV_{in}}{as^2 + bs + 1 - GB} \tag{4.24}$$

$$V_o = \frac{\frac{AsV_{in}}{1-GB}}{\frac{a}{1-GB}s^2 + \frac{b}{1-GB}s + 1} \tag{4.25}$$

After rearrangement, above equation can be written as:

$$V_o = \frac{sAV_{in}}{as^2+bs+1-GB} \tag{4.26}$$

The transfer function in equation (4.26) is 2nd order VM BPF with center frequency, b.w and gain given by

$$\omega_0 = \sqrt{\left(\frac{(1 - GB)}{a}\right)} \tag{4.27}$$

$$GAIN_{BP} = \frac{aA}{b} \tag{4.28}$$

$$B.W = \frac{b}{a} \tag{4.29}$$

After assessing above equations it can be concluded that the center frequency can be electronically controlled by regulating the external voltage amplifier’s gain with no disturbance in the b.w and the BP filter’s gain .

4.3. Concluding remarks: In this chapter shadow filter technique and its numerous schemes of implementations have been described. In this dissertation above scheme is applied for inverse filter technique. Here the basic second order filter will be inverse filter and above mentioned technique of shadow filter will be applied to it. Brief introduction of inverse filter is done in the subsequent chapter.

Chapter

5. INVERSE FILTER

5.1. Introduction

Inverse-filters are essential in frequent applications in communication, signal processing, control and instrumentation systems. Few instances such as to observe the signal's relative cleanness [61] and to do the function of de-emphasis in FM systems [62].

There are frequent circumstances where altercation of electric signal through a linear or non linear transformation before processed or transmission and passed via a system executing the inverse of initial transformation such as , division and multiplication [63], pre-emphasis and de-emphasis in FM systems [62], logarithmic and exponential amplifiers occur. Despite the fact that the problem of inverse filtering in digital filtering can be easily done, continuous time circuit didn't have general method for inverse filter till 1997.In 1997, A. Leuciuc proposed for the first time a scheme to achieve the invrse trnsfer charcteristic for nonlinear-resistive circuits and inverse transfer function intended for linear dynamic systems. This method is presented using nullors in main role.

Nullor is one of the prime active circuit elements. The pair of nullator-norator regarded by way of a case of limiting whichever floatng controled sourc [64].This pair is employed as prototype numerous activ electrnic systems, for instance: operational mirrored amplifier,op-amp and CCIIs. It is not easy to implement infnite gain controled sourc whichever in preparation or in comp simulatons.

5.1.1. Basic structure: The subsequent method is applied only to voltage output and voltage driven circuits. Analogous outcomes could be effortlessly derived in the situation of current-output and current-driven circuts. Fundamental to achieve the circuit's inverse-system including a nullor is revelaed in Fig. 5.1.Two cases are considered, separately, on behalf of linear-dynamic circuts and on behalf of nonlinear-resistive circuts.

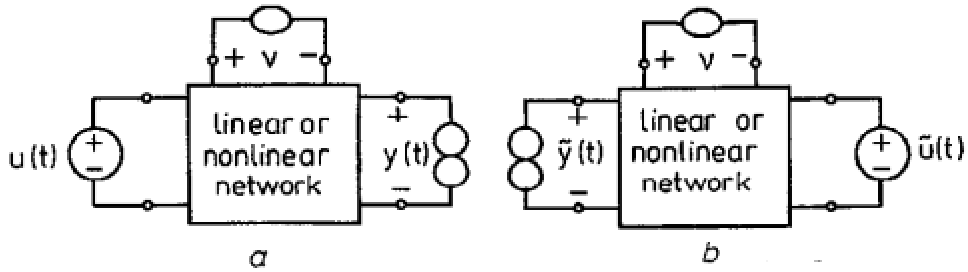

Fig. 5.1(a) Orignal circut (b) its inverse [4]

A. Linear-dynamic: For the basic original circuit, the voltag at the nulator terminal revealed as:

$$V(s) = H_1\,(S)*U(S) + H_2(s)*Y(s) \qquad (5.1)$$

And for the inverse one it is

$$V(s) = H_1(S)*\acute{Y}\,(S) + H_2(s)*\acute{U}(s) \qquad (5.2)$$

Where the transfer function of right and left ports are $H_2(s)$ and $H_1(s)$ and correspondingly. Because of presence of nullator, the voltage is placed .Therefore i/p-o/p transfer function aimed at the above 2 circutsgiven by:

$$\mathrm{H(s)} = \frac{Y(s)}{U(s)} = -\frac{\mathrm{H_1(s)}}{\mathrm{H_2(s)}} \qquad (5.3)$$

$$\hat{\mathrm{H}}(\mathrm{s}) = \frac{\acute{Y}\,(s)}{\acute{U}(s)} = -\frac{\mathrm{H_2(s)}}{\mathrm{H_1(s)}} = \frac{1}{\mathrm{H(s)}} \qquad (5.4)$$

B. Nonlinear-resistive: Voltge atnullator’s terminal meant for fundamental and inverse circuit respectively is

$$v = f(u,y) = 0 \qquad (5.5)$$

$$w = f(g,u) = 0 \qquad (5.6)$$

It is apparent that Fig. 1(b) circuit has an i/o-o/p characteristic (U, j) alike to the i/o-o/p charcteristic (U, y) of intial circut.

Verification in circuit revealed overhead does not guarantee operation aimed at inverse transformation of intial circut into Fig. 1b is done, in the view that if G = y, then j = U. To accomplish this it is requires that in situation of linear-dynamic circuts, the transfer function H(s) be minimum phase (no poles in right half of s-plane) and into situation of nonlinear-resistive circuts, explicit transfr charcteristic y(u) selected as invertible.

5.2. Inverse filter procedure to transform VM op-amp based filter into (FTFN)-based inverse filter

The author in [5] suggested a process to transform VM RC filter based on operational amplifier into a CM inverse filter based on four terminal floating nullor (FTFN).

5.2.1 Procedure of transformation: An nulor representative of an FTFN is revealed in Fig 5.2(a). Its port relations are characterized by

$$I_1 = I_2 = 0 \tag{5.7}$$

$$V_1 = V_2 \tag{5.8}$$

$$I_{o1} = I_{o2} \tag{5.9}$$

The o/p impdances of ports W also Z is generally arbitrary. But FTFN in Fig. 5.2(b) has high o/p impdance of Z terminal and low the o/p impdance of W terminal.

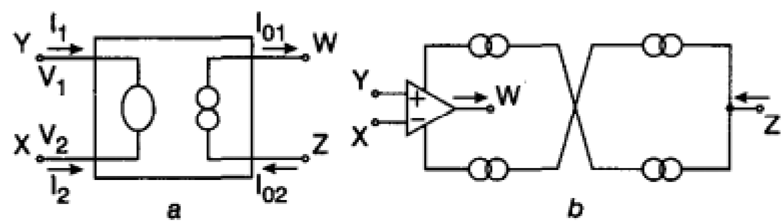

Fig. 5.2(a) and (b) Nullor model of FTFN[5]

The process to achieve the CM FTFNbased inverse filter is as follows:

(i) First step is to replacement of the op-amps in the RC circuit based on op-amp with nullator/ nortor pairs

(ii) Then drivng sources and nortors are exchanged with each other [4]

(iii) The RC:CR duality transformation is carried out[62]

(iv) Replacement of the nullors with FTFNs, accordingly realizing the inverse filter.

Here the transfer function should be minimum phase function that is, no poles in right-half s-plane[4]. Because of the application of dual-transformation, such procedure could be used for planar circut.

5.2.2. Extension of the procedure to realize CM FTFN-based inverse filters Using nullors

The authors in [10] proposed a convenient procedure to achieve CM FTFN based inverse filters from the acute VM filters by employing nullors. All of the resulting filters possesses an inverse transfer function similar to that of the original VM filter without the requirement for any matching conditions.

5.2.3. Transformation method: Since an operational amplifier could be represented by nulor by grounding 1 port of nortor, substituting CCII with nulator or nortor-pair by 1 the nulator's port linking 1 port of nortor, FTFN is equivalent to an ideal nullor. CM inverse filter can be achieved when taking into account the active RC nullorbased VM filter with examples of CCII ,operational amplifier otherwise filter created with FTFN by by means of given method:

(i) In the circuit, substitute every active devices by their equivalent nullor/nrator-pairs

(ii) Exchange outputted norator andd driving source with each other[7]

(iii) After interchanging the norator with nulator, utilize the adjoint transformation.

(iv) Replacement of the nullors by FTFNs, accordingly realizing the inverse filter.

At that time seeing that commercial FTFN were not available, they were synthesized using two CCII-s or two CCII+s. This type of synthesis was first suggested by Senani in [7]. Thus, the inverse filter in both [5,10] were realizable with two CFOAS.

In literature due to increase interest in CM elements cause of their numerous advantages, there is an increased attention on realizing inverse filter using them

5.4. Concluding remarks: In this chapter basic structure of inverse filter has been described. Various processes to transform VM RC filter into a CM inverse filter using FTFN and nullors have been mentioned. In this dissertation above scheme is applied along with shadow filter technique to propose an inverse shadow filter as will be covered in subsequent chapter.

Chapter

6. PROPOSED INVERSE SHADOW FILTERS

6.1 Introduction

Active biquad filters be able to be categorized based on their topology: fixed and variable topology type. The variable topology is moreover divided into VM, CM and MM types. The biquad circuits working in mixed-mode can be operated in voltage, current, trans-impedance, and transadmittance modes. One of the most widely used analog filters topology is variable topology type VM filters. By applying shadow filter technique along with inverse filter method as described in previous chapters, inverse shadow filters of low pass, high pass, band pass and band reject filter type have been proposed. Their structures and workings have been described in subsequent sections.

6.2 Proposed configuration of Inverse Shadow Low Pass Filter

Figure 6.1 reveals proposed inverse-shadow low-pass filter which is achieved after modification of Fig. 1(a) [28].Here CFOA1 to CFOA3 form an inverse low pass filter and CFOA4 is applied as an external amplifier in feedback with the help of resistances R_a and R_b to get an inverse shadow low pass filter. Supposing idyllic CFOAs, repetitive analyis generates transfer function of equation 6.24 designed to get particular-i/o particular-o/p with sub-circuit which is built by means of CFOA1-4, R0-3, C1-2 and R_a and $R_b$resistance for feedback.

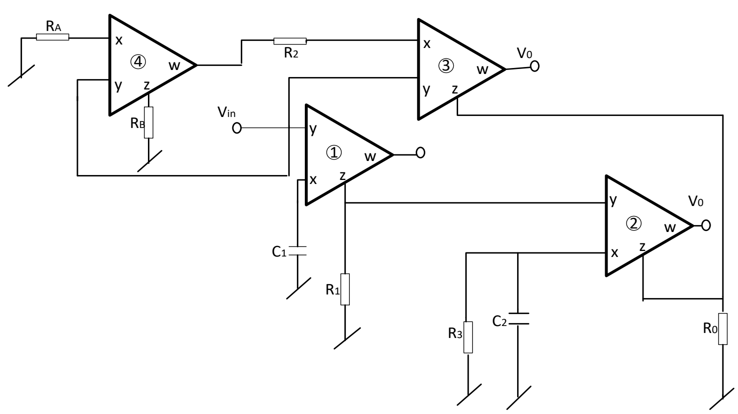

Fig 6.1 Inverse shadow low pass filter

Supposing idyllic CFOAs, repetitive examination generate transfer function of eqution given by:

$$Vin\left(s^2 + s\left(\frac{1}{C_2R_3}\right) + \frac{1}{C_1C_2R_1R_2}\right) - \frac{sV1}{C_1R_1} = Vo\left(\frac{R_2}{R_o}\right)\left(\frac{1}{C_1C_2R_1R_2}\right) \tag{6.1}$$

In Fig. 6.1 the sub-circuit built about CFOA4, Ra and Rb is a votage amp having gain revealed by below eqution:

$$G = \frac{V1}{Vin} = -\frac{R_b}{R_a} \tag{6.2}$$

Combining equation 6.1 and 6.2 the transfer function of above inverse shadow low pass filter is given by:

$$\frac{Vo}{Vin} = \frac{1}{\frac{\left(\frac{R2}{Ro}\right)\left(\frac{1}{C_1C_2R_1R_2}\right)}{s^2 + s\left(\frac{G}{C_1R_1} + \frac{1}{C_2R_3}\right) + \frac{1}{C_1C_2R_1R_2}}} \tag{6.3}$$

The characteristic frequency, b.w gain and quality factor are given by:

$$\omega_0 = \sqrt{\left(\frac{1}{C_1C_2R_1R_2}\right)} \tag{6.4}$$

$$B.W = \left(\frac{G}{C_1R_1} + \frac{1}{C_2R_3}\right) \tag{6.5}$$

$$GAIN_{LP} = \frac{R_2}{R_0} \tag{6.6}$$

$$Q = \frac{\omega_0}{B.W} = \sqrt{\left(\frac{1}{C_1C_2R_1R_2}\right)}\frac{1}{\left(\frac{G}{C_1R_1} + \frac{1}{C_2R_3}\right)} \tag{6.7}$$

Inspection of above equation shows that the tuning of quality factor Q can be made by adjusting the gain G external sub-circuits built around CFOA4 using the resistances R_A and R_B without disturbing the characteristic frequency.

6.2.1. Sensitivity study: Passive senstivities of parameter ω_0 and $\frac{\omega_0}{Q_0}$ aimed at proposed configuration could be define by:

$$S^{\omega_0}_{C_1,C_2,R_1,R_2} = -\frac{1}{2} \tag{6.7}$$

$$S^{\frac{\omega_0}{Q_0}}_{C_1,C_2,R_1,R_3} = -1 \tag{6.8}$$

Therefore, all the active as well as passive sensitivities are no higher than unity.

Effect of gain G is studied, senstivities $\frac{\omega_0}{Q_0}$ and ω_0 w.r.t G are calculated and conveyed:

$$S_G^{\frac{\omega_0}{Q_0}} = 1 \tag{6.9}$$

$$S_G^{\omega_0} = 0 \tag{6.10}$$

As seen from above all senstivities are in lesser amount than unity accordingly accidental alteration of G would not be rendering several influence on the coefficients of the inverse shadoe LP filter.

6.2.2. Simulation results of proposed filter

To verify the working of the proposed inverse shadow LPF, it is simulated using commercially available AD844. For this purpose the simulation is done using PSPICE software. Pspice simulated implementation of this filter is shown below:

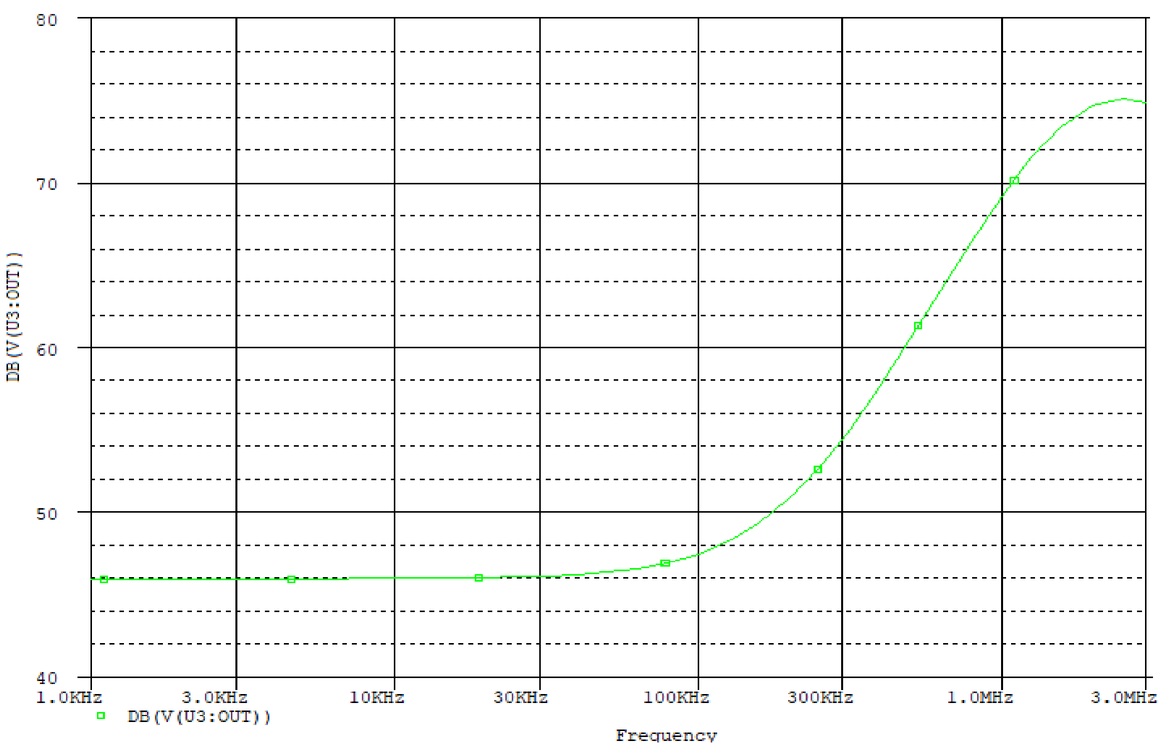

Fig. 6.2 Simulated LPF: magnitude response of proposed shadow inverse lowpass filter

Tuning of quality factor by adjusting gain G while maintaining constant characteristic frequency: The circut in Fig. 6.1 is executed experimentally where gain G is worked for tuning quality factor. Fig. 6.2 shows comparison between magnitude waveform having different quality factor while maintaining same characteristic frequency of the LPF. As seen from Fig.6.2 the characteristic frequency remain constant of value 15.9KHz

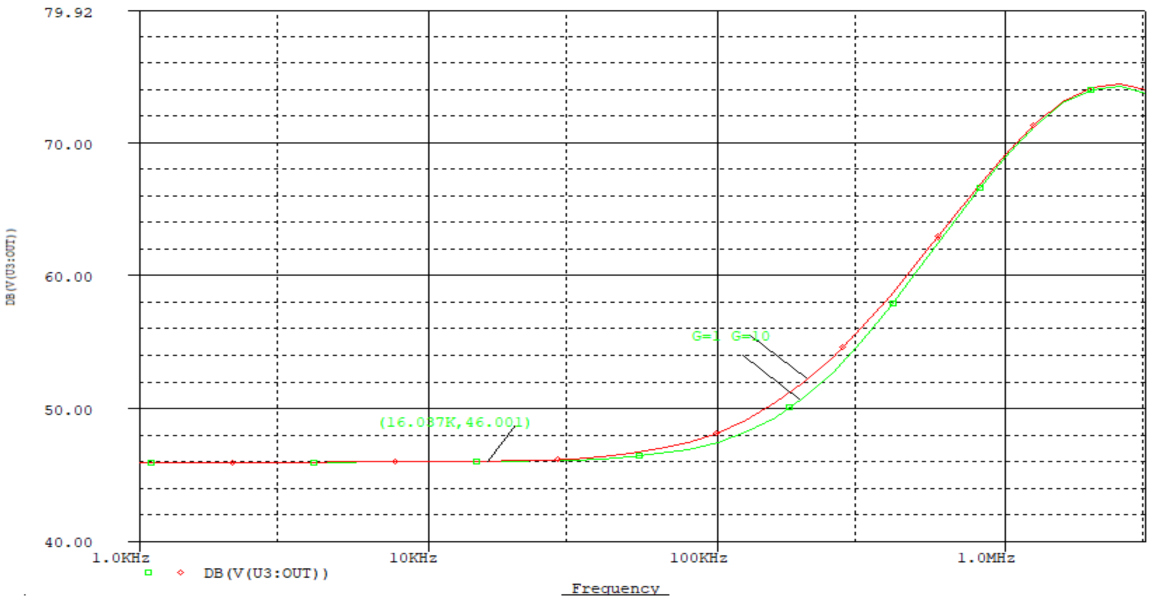

Fig. 6.3 Comparison between 2 gain and phase waveforms having different quality of LPF of Fig.6.1 with gain set to 1 and 10

In Fig. 6.1 by setting gain to different values of 1,20,35,50 the magnitude waveforms of inverse-shadow low-pass fiter is revealed below:

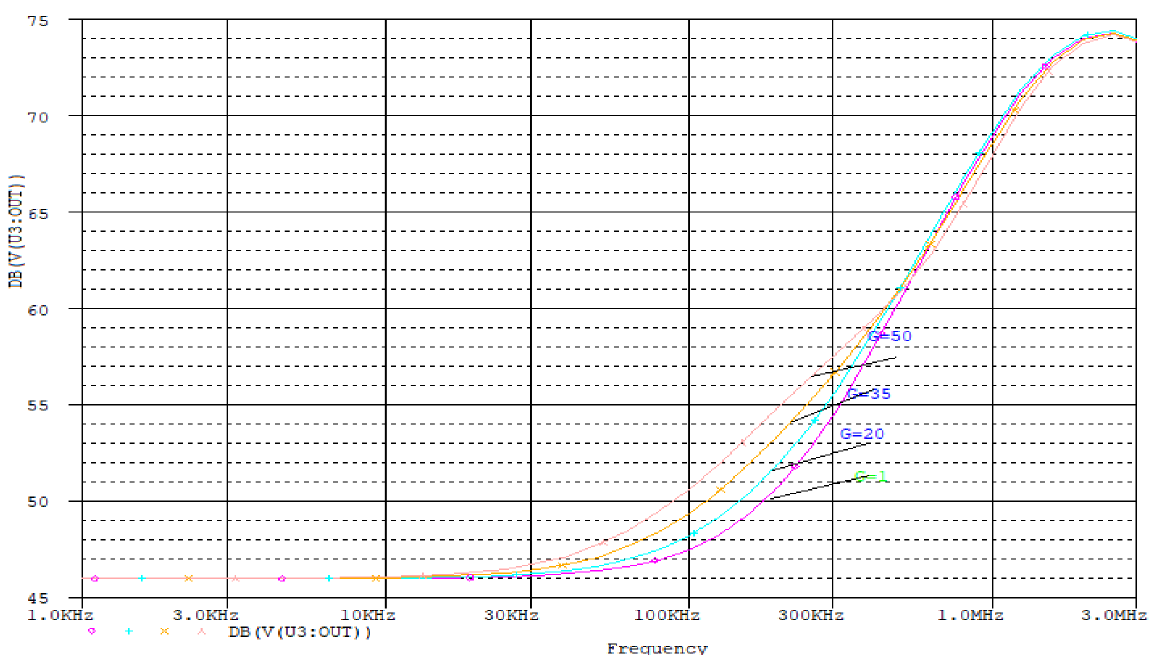

Fig. 6.4 magnitude response of inverse-shadow low Pass filter

6.3. Proposed configuration of Inverse Shadow High Pass Filter

Figure 6.5 reveals proposed inverse-shadow High-pass filter which is achived after modification of Fig. 1(c) [28]. Here CFOA1 to CFOA3 form an inverse high pass filter and CFOA4 is applied as an external amplifier in feedback with the help of resistances R_a and R_b to get an inverse shadow high pass filter. Supposing idyllic CFOAs, repetitive analyis generates transfer function of equation 6.13 designed to get particular-i/o particular-o/p with sub-circuit which is built by CFOA1-4, R0-3, C1-2 and R_a and R_b resistance for feedback

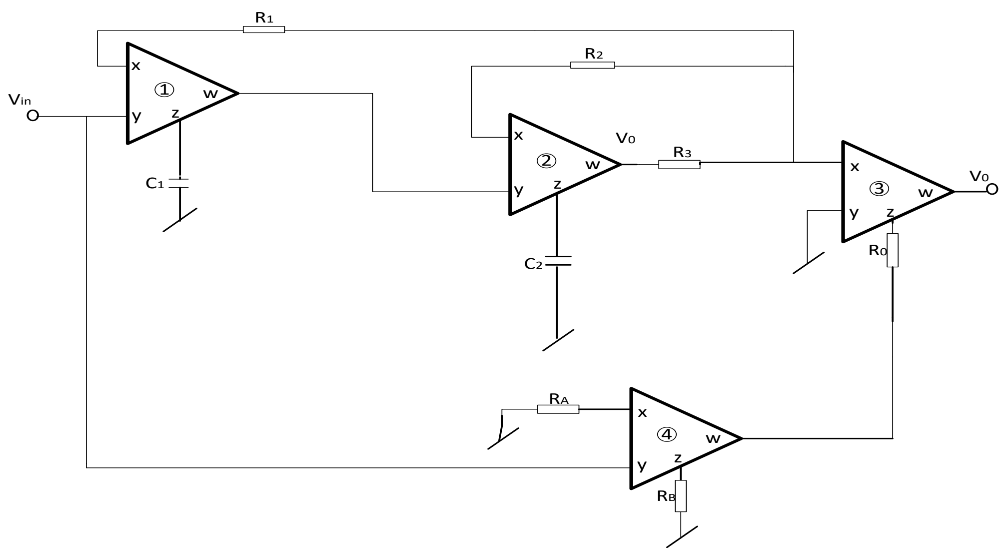

Fig.6.5 Inverse shadow High pass filter

Supposing idyllic CFOAs, repetitive examination generate transfer function of eqution given by:

$$Vin\left(s^2 + s\left(\frac{1}{C_1R_2}\right)+\frac{1}{C_1C_2R_2R_3}\right) - \frac{s^2(R_1)\,V1}{Ro} = s^2(R_1)Vo \qquad (6.11)$$

In Fig. 6.5 the sub-circut built about CFOA4, Ra and Rb is a voltage amp having gain revealed by below eqution

$$G = \frac{V1}{Vin} = -\frac{R_b}{R_a} \qquad (6.12)$$

Combining equation 6.11 and 6.12 the transfer function of above inverse shadow low pass filter is given by:

$$\frac{Vo}{Vin} = \frac{-1}{\frac{\left(\frac{s^2}{Ro\left(\frac{G}{Ro}+\frac{1}{R1}\right)}\right)}{s^2 + \frac{s}{C_1R_1R_2\left(\frac{G}{Ro}+\frac{1}{R1}\right)} + \frac{1}{C_1C_2R_1R_2R_3\left(\frac{G}{Ro}+\frac{1}{R1}\right)}}} \tag{6.13}$$

The characteristic frequency, b.w gain and quality factor are given by:

$$\omega_0 = \sqrt{\left(\frac{1}{C_1C_2R_1R_2R_3\left(\frac{G}{Ro}+\frac{1}{R1}\right)}\right)} \tag{6.14}$$

$$B.W = \frac{1}{C_1R_1R_2\left(\frac{G}{Ro}+\frac{1}{R1}\right)} \tag{6.15}$$

$$GAIN_{LP} = \frac{1}{RoM} \tag{6.16}$$

$$Q = \frac{\omega_0}{B.W} = \sqrt{\left(\frac{1}{C_1C_2R_1R_2}\right)}\left(\frac{1}{C_1R_1R_2\left(\frac{G}{Ro}+\frac{1}{R1}\right)}\right) \tag{6.17}$$

Inspection of above equation shows that the tuning of center frequency can be made by adjusting the gain G external sub-circuits built around CFOA4 using the resistances R_A and R_B.

6.3.1. Sensitivity study: Passive senstivities of parameter ω_0 , $\frac{\omega_0}{Q_0}$ andQ_0 aimed at proposed configuration could be define by:

$$S^{\omega_0}_{C_1,C_2,R_1,R_2,R_3,R_0} = -\frac{1}{2}, S^{\omega_0}_{R_1} = -\frac{1}{2}, S^{\omega_0}_{R_0} = -\frac{1}{2} \tag{6.18}$$

$$S^{\frac{\omega_0}{Q_0}}_{C_1,R_1,R_2} = -1, S^{\frac{\omega_0}{Q_0}}_{R_1} = 0, S^{\frac{\omega_0}{Q_0}}_{R_0} = 1 \tag{6.19}$$

$$S^{Q_0}_{C_2,R_3,R_0} = -\frac{1}{2}, S^{Q_0}_{C_1,R_2} = \frac{1}{2}, S^{Q_0}_{R_1} = 0 \tag{6.20}$$

Therefore, all the active as well as passive sensitivities are no higher than unity.

Effect of gain G is studied; sensitivities $\frac{\omega_0}{Q_0}$ and ω_0 w.r.t G are calculated and conveyed:

$$S^{\omega_0}_{G} = -\frac{1}{2}, S^{\frac{\omega_0}{Q_0}}_{G} = 1, S^{Q_0}_{G} = \frac{1}{2} \tag{6.21}$$

As seen from above all senstivities are in lesser amount than unity accordingly accidental alteration of G would not be rendering several influence on coefficient of the inverse shadow HP flter.

6.3.2. Simulation results

To verify the working of the proposed inverse shadow HPF, it is simulated using commercially available AD844. For this purpose the simulation is done using PSPICE software. Pspice simulated implementation of this filter is shown below:

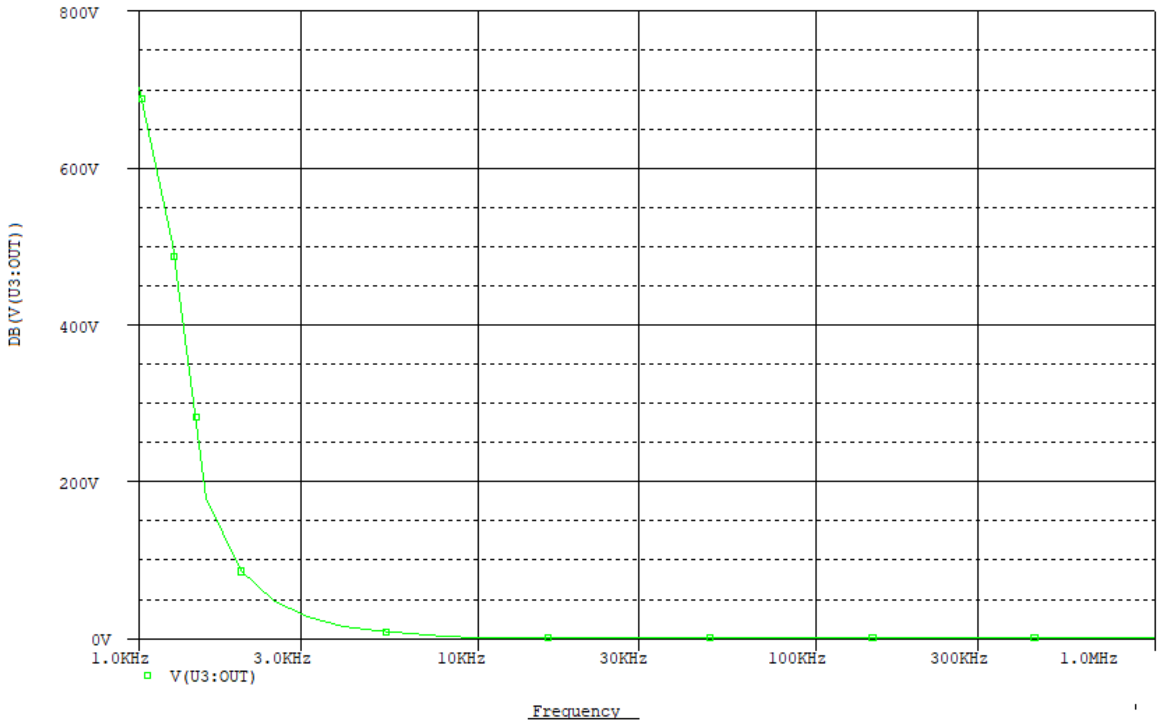

Fig. 6.6 Simulated HPF: magnitude response of proposed shadow inverse Highpass filter

Tuning of center frequency by adjusting gain G: The circuit in Fig. 6.5 is implemented experimentally where G is used to tune the center frequency. Fig. 6.7 shows comparison between 2 magnitude waveform having varying center frequency.

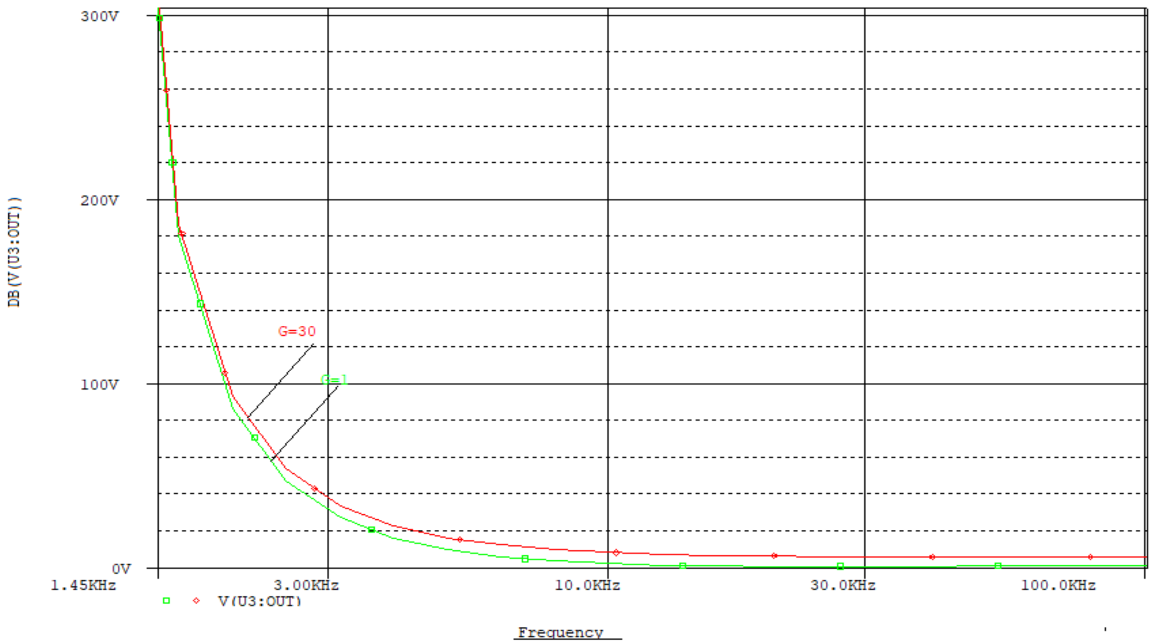

Fig. 6.7 Comparison between 2 gain waveforms having different center frequency of HPF of Fig. 6.5 with gain set to 1 and 30

In Fig. 6.5 by setting gain to different values of 1,2,4,6 the gain waveforms of inverse shadow high-pass filter is revealed below:

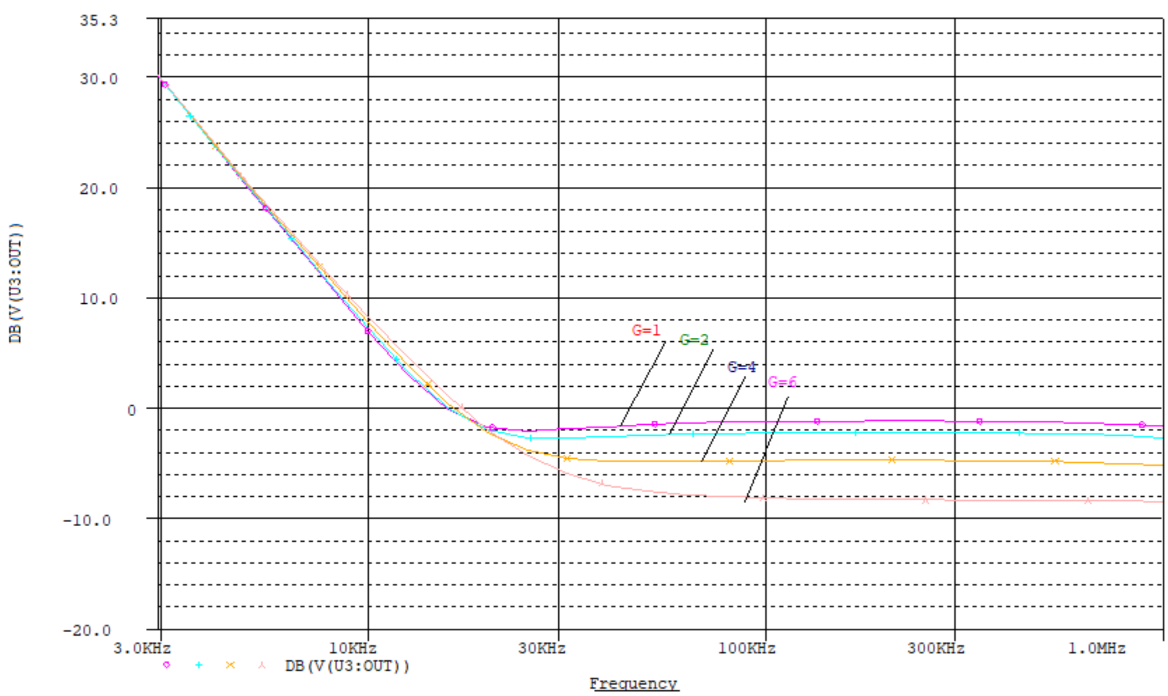

Fig. 6.8 Measured gain freq charcteristics of inverse-shadow HPF response by setting characteristic frequency

6.4. Proposed configuration of Inverse Shadow Band Pass Filter

Figure 6.9 reveals proposed inverse-shadow band-pass filter which is achieved after modification Fig. 1(b) [28]. Here CFOA1 to CFOA3 form an inverse band pass filter and CFOA4 is applied as an external amplifier in feedback with the help of resistances R_a and R_b to get an inverse shadow band pass filter. Supposing idyllic CFOAs, repetitive analyis generates transfer function of equation (6.24) designed to get particular-i/o particular-o/p with sub-circuit which is built by means of CFOA1-4, R0-3, C1-2 and $R_a$and $R_b$resistance for feedback.

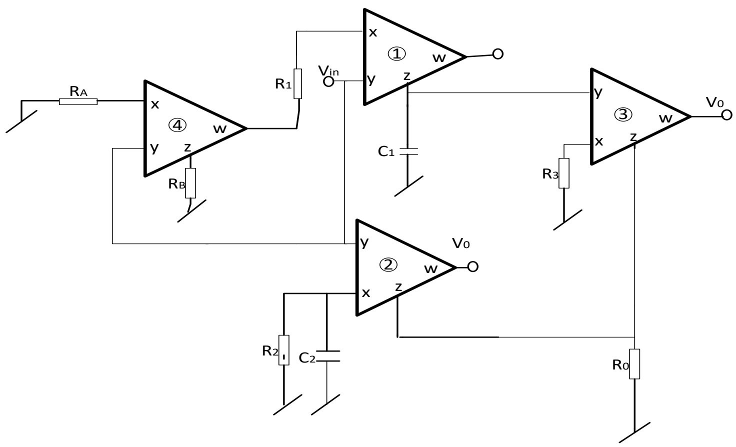

Fig. 6.9 Inverse shadow band pass filter

Supposing idyllic CFOAs, repetitive examination generate transfer function of eqution given by:

$$Vin\left(s^2 + s\left(\frac{1}{C_2R_2}\right)+\frac{1}{C_1C_2R_1R_3}\right) - \frac{V1}{C_1C_2R_1R_3} = Vo\left(\frac{R_2}{R_o}\right)\left(\frac{s}{C_2R_2}\right) \quad (6.22)$$

In Fig. 6.9 the sub-circut built about CFOA4, Ra and Rb is a votage amp having gain revealed by below eqution:

$$G = \frac{V1}{Vin} = -\frac{R_b}{R_a} \quad (6.23)$$

Combining equation 6.22 and 6.23 the transfer function of above inverse shadow low pass filter is given by:

$$\frac{Vo}{Vin} = \frac{1}{\frac{\left(\left(\frac{s}{C_2R_2}\right)\left(\frac{R2}{Ro}\right)\right)}{s^2 + \frac{s}{C_2R_2}+\frac{(1+G)}{C_1C_2R_1R_3}}} \quad (6.24)$$

The characteristic frequency, b.w, gain and quality factor are given by:

$$\omega_0 = \sqrt{\left(\frac{(1+G)}{C_1 C_2 R_1 R_3}\right)} \quad (6.25)$$

$$B.W = \frac{1}{C_2 R_2} \quad (6.26)$$

$$GAIN_{LP} = \frac{R_2}{R_0} \quad (6.27)$$

$$Q = \frac{\omega_0}{B.W} = \sqrt{\left(\frac{(1+G)}{C_1 C_2 R_1 R_3}\right)}(C_2 R_2) \quad (6.28)$$

Inspection of above equation shows that the tuning of center frequency of inverse shadow bandpass filer can be made by adjusting the gain G of external sub-circuits built around CFOA4 using the resistances R_A and R_B without disturbing the bandwidth.

6.4.1. Sensitivity Study: Passive senstivities of parameter ω_0 and $\frac{\omega_0}{Q_0}$ aimed at proposed configuration could be define by:

$$S_{C_2,R_2}^{\frac{\omega_0}{Q_0}} = -1 \quad (6.29)$$

$$S_{C_1,C_2,R_1,R_3}^{\omega_0} = -\frac{1}{2} \quad (6.30)$$

Therefore, all the active as well as passive sensitivities are no higher than unity.

Effect of gain G is studied; the sensitivities $\frac{\omega_0}{Q_0}$ and ω_0 w.r.t G are calculated and conveyed:

$$S_G^{\frac{\omega_0}{Q_0}} = 0 \quad (6.31)$$

$$S_G^{\omega_0} = \frac{1}{2} \quad (6.32)$$

As seen from above all senstivities are in lesser amount than unity accordingly accidental alteration of G would not be rendering several influence on coefficients of the inverse shadow BP filter.

6.4.2. Simulation results

To verify the working of the proposed inverse shadow BPF, it is simulated using commercially available AD844. For this purpose the simulation is done using PSPICE software. Pspice simulated implementation of this filter is shown below:

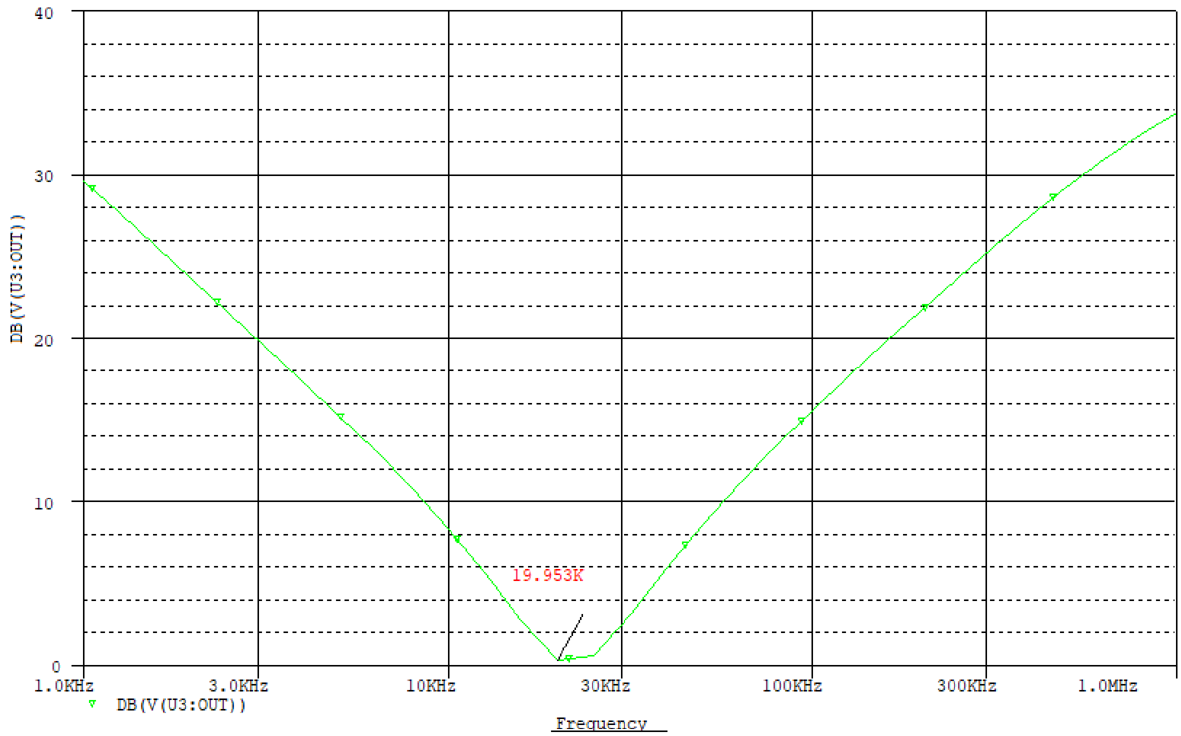

Fig. 6.10 Simulated BPF: magnitude response of proposed shadow inverse bandpass filter

Tuning of center frequency by adjusting gain G while maintaining constant bandwidth: the circuit in Fig. 6.9 is executed experimentally where gain G is worked for tuning of center frequency. Fig. 6.11 shows comparison between 2 magnitude waveform having varying center frequency while maintaining same b.w of the BPF.

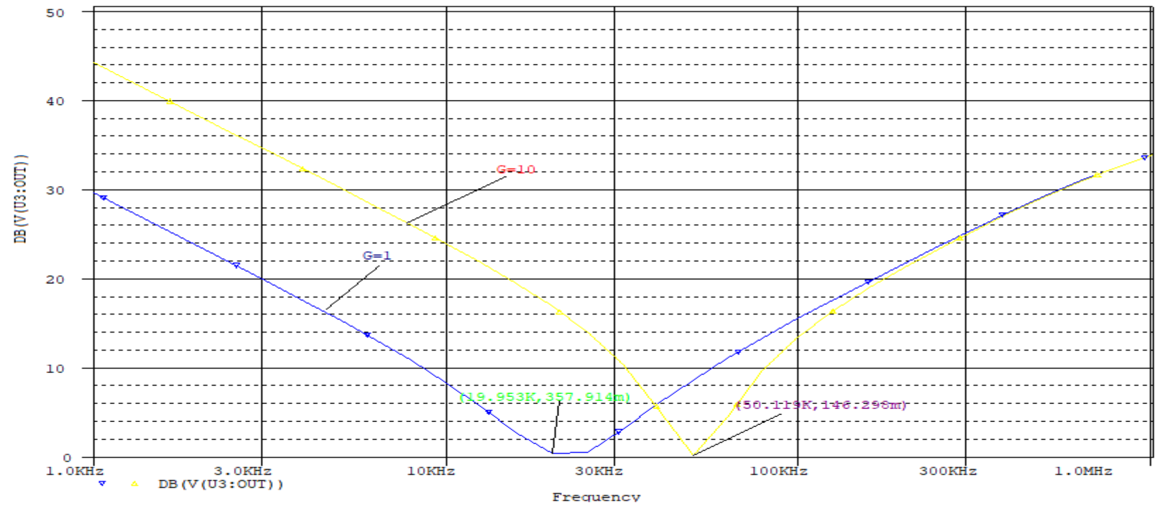

Fig. 6.11 Comparison between 2 gain waveforms having different center frequency of BPF of Fig.6.9 with gain set to 1 and 10

In Fig. 6.9 by setting gain to different values of 1,4,7,10,the gain waveforms of inverse shadow band-pass filter is revealed below:

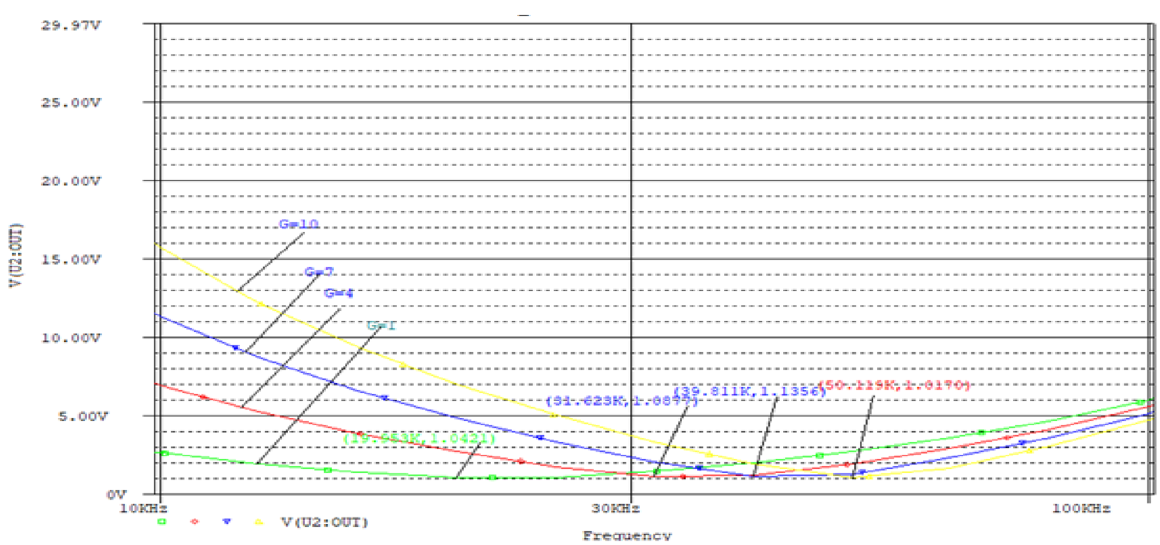

Fig. 6.12 Measured magnitude freq charcteristics of inverse shadow BPF response by setting center frequency

6.5. Proposed configuration of Inverse Shadow Band Notch Filter

Figure 6.13 reveals proposed inverse-shadow band-notch filter which is achieved after modification of Fig.1(d)[28]. Here CFOA1 to CFOA3 form an inverse band notch filter and CFOA4 is applied as an external amplifier in feedback with the help of resistances R_a and R_b to get an inverse shadow band notch filter. Supposing idyllic CFOAs, repetitive analyis generates transfer function of Equation (6.35) designed to get particular-i/p particular-o/p with sub-circuit which is built by means of CFOA1-4, R1-4, C1-2 and RA and RB resistance for feedback.

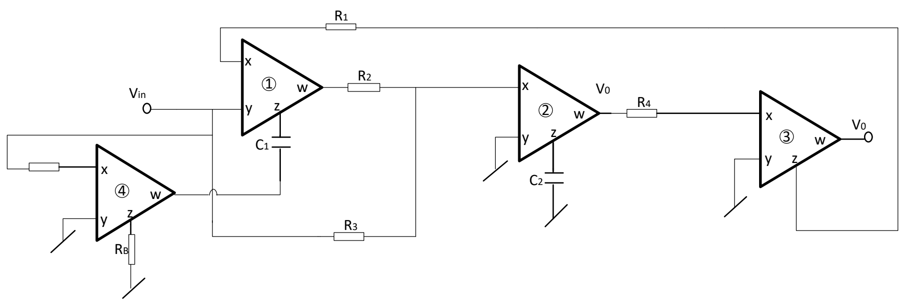

Fig. 6.13 Inverse shadow band reject filters

Supposing idyllic CFOAs, repetitive examination generate transfer function of eqution given by:

$$Vin\left(s^2+\frac{sR_1}{C_2R_4R_3}+\frac{1}{C_1C_2R_2R_4}\right)+\frac{sR_1GV1}{C_2R_4\,R_2}=Vo\left(s^2+\left(\frac{1}{C_1C_2R_2R_4}\right)\right) \quad (6.33)$$

In Fig. 6.13 the sub-circut built about CFOA4, Ra ,Rb is a voltge amp having gain revealed by below eqution:

$$G=\frac{V1}{Vin}=\frac{R_b}{R_a} \quad (6.34)$$

Combining equation 7.1 and 7.2 the transfer function of above inverse shadow band reject filter is given by:

$$\frac{Vo}{Vin}=\frac{1}{\dfrac{s^2+\left(\frac{1}{C_1C_2R_2R_4}\right)}{s^2+\frac{sR_1}{C_2R_4}\left(\frac{1}{R_3}+\frac{G}{R_2}\right)+\frac{1}{C_1C_2R_2R_4}}} \quad (6.35)$$

The characteristic frequency, b.w, gain and quality factor are given by:

$$\omega_0 = \sqrt{\left(\frac{1}{C_1 C_2 R_2 R_4}\right)} \tag{6.36}$$

$$B.W = \frac{R_1}{C_2 R_4}\left(\frac{1}{R_3} + \frac{G}{R_2}\right) \tag{6.37}$$

$$GAIN_{BR} = 1 \tag{6.38}$$

$$Q = \frac{\omega_0}{B.W} = \sqrt{\left(\frac{C_2 R_4}{C_1 R_2}\right)}\frac{1}{R_1\left(\frac{1}{R_3}+\frac{G}{R_2}\right)} \tag{6.39}$$

Inspection of above equation shows that the tuning of bandwidth of inverse shadow bandreject filer could be prepared while altering gain,G of external sub-circuits built arond CFOA4 by means of resitances R_A and R_B without disturbing center frequency.

6.5.1. Sensitivity Study:

Passive senstivities of parameter ω_0 and $\frac{\omega_0}{Q_0}$ aimed at proposed configuration could define by:

$$S^{\omega_0}_{C_1,C_2,R_2,R_4} = -\frac{1}{2} \tag{6.40}$$

$$S^{\frac{\omega_0}{Q_0}}_{C_2,R_2,R_3,R_4} = -1 \tag{6.41}$$

$$S^{\frac{\omega_0}{Q_0}}_{R_1} = 1 \tag{6.42}$$

Therefore, all the active as well as passive sensitivities are no higher than unity.

Effect of gain G is studied; sensitivities $\frac{\omega_0}{Q_0}$ and ω_0 w.r.t G are calculated and conveyed:

$$S^{\omega_0}_{G} = 0, S^{\frac{\omega_0}{Q_0}}_{G} = 1 \tag{6.43}$$

As seen from above all sensitivties are in lesser amount than unity accordingly accidental alteration of G would not be rendering several influence on coefficients of the inverse shadow BR filter.

6.5.2. Simulation results

To verify the working of the proposed inverse shadow BRF, it is simulated using commercially available AD844. For this purpose the simulation is done using PSPICE software. Pspice simulated implementation of this filter is shown below:

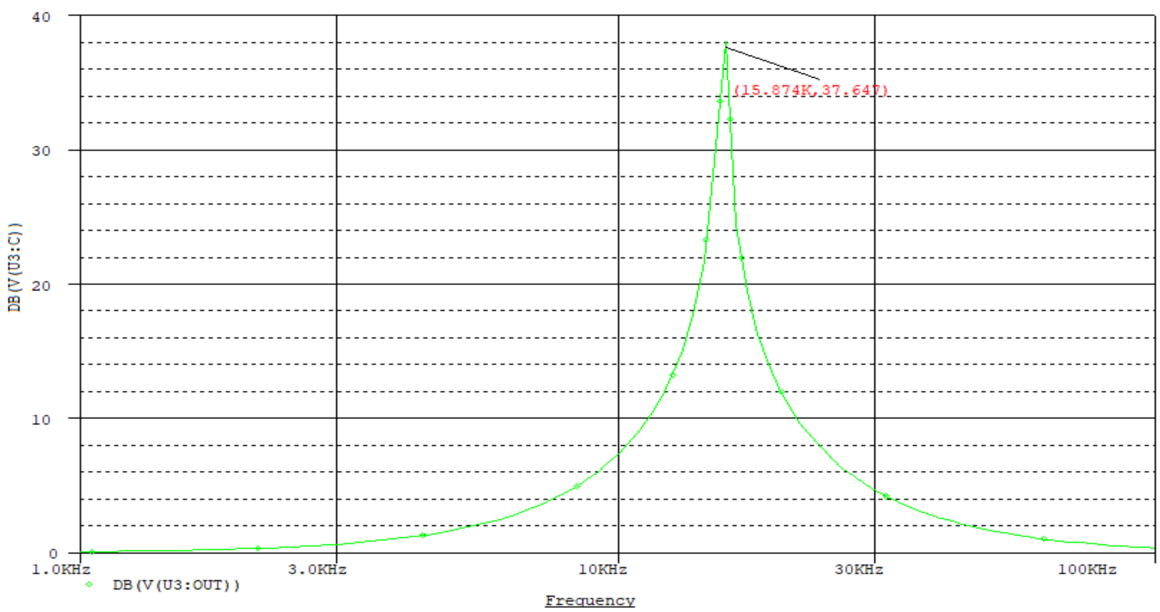

Fig. 6.14 Simulated BRF: magnitude response of proposed shadow inverse lowpass filter

Tuning by adjusting gain G while maintaining constant center frequency: the circut of Fig.6.13 is executed experimentally whre gain G is worked for tuning of bandwidth. Fig. 6.15 shows comparison between 2 magnitude waveform having varying b.w while maintaining same characteristic frequency of the BRF. As seen from Fig. 6.15 the characteristic frequency remain constant of value 15.9KHz.

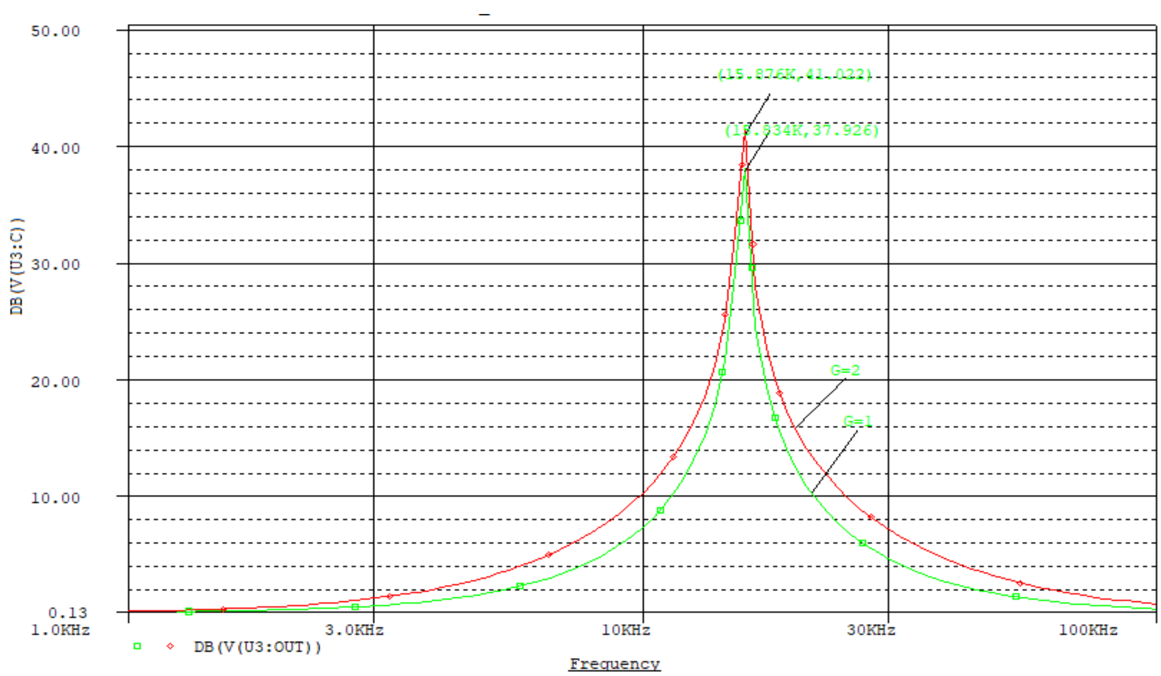

Fig. 6.15 Comparison between 2 gain waveforms having different b.w of BRF of Fig.6.13 with gain set to 1 and 2

In Fig. 6.16 by setting gain to different values of 1,4,7,10,the gain waveforms of inverse shadow band-reject filter is revealed below:

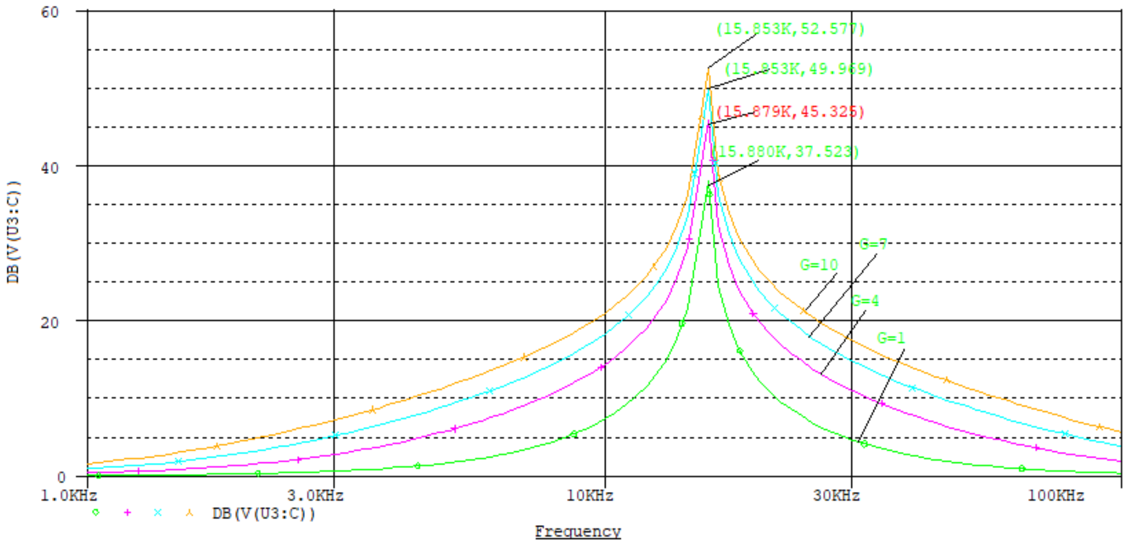

Fig. 6.16 Measured magnitude frequency characteristics of inverse shadow BRF response by working bandwidth.

Chapter

7. CONCLUSION AND FUTURE WORK

This dissertation presents realization of a new inverse shadow LPF, BPF, HPF and BRF. Inside introductory chapter a concise summery on evolution of current mode signal processing along with brief introduction on inverse and shadow filtering is presented which guide to motivation intended for the work commenced in this thesis. Comprehensive objective of the work is submitted thereafter. The theory of shadow filter and inverse filter has been revised and examined in chapter 4 and 5 respectively. A new extnsion to shadow filter theory and inverse filter have been presented in literature requiring no call in lieu of externally attached summer circuit in inverse shadow filters designing in contrast to Fabre's desig. Furthermore, the scheme of the shadow filter in addition to inverse filter have been combined to embrace distinctive filter configrations to acquire tunable center freq as it is enlighten in the LP and HP filters and also for tunable bandwidth (b.w) alongwith constant center freq as in the band pass filter. A novel offered design based on technique of shadow filter is set up to tune bandwidth of BR filter with unvarying center frequency. The workability of all novel circuits have been proved by SPICE simulations founded upon AD844-type CFOAs.

The future work includes different applications based on inverse shadow filter. Design of a inverse shadow high pass filter having the ability to tune quality factor and center frequency independently. Work can be done on the designing of a universal inverse shadow filter.

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