A Dissertation On

MODIFIED CFOA AND ITS APPLICATION FOR THE SIMULATION OF INDUCTOR, FILTER AND INVERSE FILTER

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ABSTRACT

The recent trend in electronics has been towards reducing the size of circuits, a trend which culminated in the development of integrated circuits, but it has been proven impractical to achieve a comparable reduction in the size of inductors because of the relation between the size of inductor and the quality factor. Also inductors are seldom used at low frequencies because at very low frequency the size and weight of inductors become exceedingly large and quality factor becomes very low. Fortunately, active circuits can sometimes synthesize the equivalent of an inductor with high quality factor.

In this dissertation, we have presented a modified current feedback operational amplifier (MCFOA), which is more suitable for realizing simulated inductors and active filters. New grounded inductor and floating inductor have been simulated using a minimum number of passive components based on one or two modified current feedback operational amplifiers (MCFOAs) and the result is validated with the implementation of RL high pass filter.

To show the flexibility of the proposed MCFOA, a single input three output (SITO) voltage mode filter and three input single output (TISO) voltage mode filters employing a single MCFOA have been simulated using the PSPICE program.

In communication, control and instrumentation systems an electrical signal may get altered by many linear or nonlinear transformation caused by the signal processors or transmission system. To recover these distortions of the signal, a system is required that has inverse transfer characteristics of the original system. Inverse filter can correct these distortions because it has frequency response, which is the reciprocal of the frequency response of the system that caused the, distortion.

In this dissertation, inverse filters have been designed using both the Modified current feedback operational amplifier (MCFOA) and standard AD844 CFOA. The workability of the proposed inverse filters is demonstrated by PSPICE simulations.

Dedication

I dedicate this thesis

to my family, my teachers and my friends for

supporting me all the way & doing all thee

wonderful things for me.

CHAPTER 1

INTRODUCTION

1.1 INDUCTOR

The recent trend in electronics has been toward reducing the size of circuits, a trend which culminated in the development of integrated circuits. Although it is relatively simple to reduce the dimensions of resistors and capacitors it has proven impractical to achieve a comparable reduction in the size of inductors. Following are the main reasons for this:

[1] Semiconductors, which provide the building material of integrated circuits, do not exhibit ferromagnetism. Hence, both the magnetic material forming the core and the conductors forming the windings of the inductor must be deposited on the semiconductor surface. This arrangement results in inductors of very low inductance L and poor quality factor Q.

[2] The inherent relation between the physical size of an inductor and its quality factor creates a size problem. If the size of inductor is reduced by replacing every linear dimension l by x l, then a new quality factor Q_n will become x^2Q, where Q is the original one. Thus, reduction in size of inductor reduces the quality factor greatly.

[3] Even large inductors are quite lossy. The best attainable Q_L (quality factor of inductor) is about 1,000; by contrast, capacitors with Q_c (quality factor of capacitor) values of 5,000 to 10,000 can be obtained.

[4] For frequencies below 20 Hz, for example, the size and weight of inductors become exceedingly large and quality factor becomes very low. Hence, inductors are seldom used at such low frequencies.

[5] Inductors using ferromagnetic materials are basically nonlinear elements. Hence, unless the amplitude of the signal which they handle is kept small and direct currents are avoided, they generate harmonic distortion.

[6] Inductors tend to act as small antennas, radiating as well as picking up electromagnetic waves. This can result in undesirable noise and coupling of signals in circuits containing inductors.

Thus, conventional wire-wound inductors are bulky and costly. There also exist fundamental limitations on the realization of inductances for microminiature and integrated circuit applications. As a result, there has been increasing interest in the realization of active RC filters requiring no inductance. The advantages in using active RC filters are the reduction in size, weight, cost, and power consumption, and an increase in the system reliability in comparison to discrete versions.

In Electronics, Instrumentation and control various signal processing functions such as active filters, Oscillator design , Phase shifters, cancellation of parasitic element etc require the desirability of Spiral Inductors but the use of inductors in analog circuits has many disadvantages like usage of space, weight, cost ,Quality factor, tunability mentioned as above. One among them is that the analog filters using inductor works well at high frequencies, however, in low frequency applications that is in the frequency range (0- 20 kHz), the inductors cannot be used for the reasons like the size and weight of the inductors are large and physically bulky and their characteristics are quite non ideal and Inductors are impossible to fabricate in monolithic form and are incompatible with any of the modem techniques for assembling electronic systems.

Aside from the large areas that are potentially consumed, another serious problem with spiral inductors is their relatively large loss, i.e., their quality factor value Q is about ten if the frequency of operation is of the order of gigahertz. If the operating frequency of the spiral inductors is decreased, their Q also reduces proportionally. The restrictions of the Q of the spiral inductors can be attributed to the resistive losses in a metal spiral and substrate losses. Also, the value of Q slightly depends on the shape of the inductor whether square or octagonal. The series resistance of the metal has the greatest influence on the total loss of the spiral inductors. Fortunately, active circuits can sometimes synthesize the equivalent of an inductor with high-Q. For this purpose, a great number of active circuits for realizing grounded inductors were reported in the literature [2]–[13]. Also, many simulated floating inductors using at least two active devices were developed in [14]–[22]. So, it is required to design the analog filters without the use of inductors. One of the possible methods is the use of inductor less filter. It should be mentioned that a circuit employing a minimum number of active and passive components is advantageous from the points of view of very large-scale integration (VLSI) implementation, area, power consumption, and cost. Also, it is well known that, in IC processes, the realization of grounded passive components, especially grounded capacitors, are more convenient than floating ones [23], [24]. Hence, the circuit designer should avoid using floating passive components and passive component-matching constraints as much as possible. In this dissertation, we present a modified current-feedback operational amplifier (MCFOA), which is more suitable for realizing simulated inductors and active filters. The proposed MCFOA is based on a composite CC reported in [41].

1.2 FILTERS

Electronic filters are electronic circuits which perform signal processing functions, specifically to remove unwanted frequency components from the signal and/or to enhance wanted ones. Electronic filters can be:

- passive or active
- analog or digital
- highpass, lowpass, bandpass, band reject (notch), or all-pass.
- discrete-time (sampled) or continuous-time
- linear or non-linear
- Infinite impulse response (IIR type) or finite impulse response (FIR type).

The most common types of electronic filters are linear filters, regardless of other aspects of their design. The oldest forms of electronic filters are passive analog linear filters, constructed using only resistors and capacitors or resistors and inductors. These are known as RC and RL single pole filters respectively. More complex multipole LC filters have also existed for many years and the operation of such filters is well understood with many books having been written about them.

Hybrid filters have also been made, typically involving combinations of analog amplifiers with mechanical resonators or delay lines. Other devices such as CCD delay lines have also been used as discrete time filters. With the availability of digital signal processing, active digital filters have become common.

1.2.1 Classification of Filters by Technology

1.2.1.1 Passive filters

Passive implementations of linear filters are based on combinations of resistors (R), inductors (L) and capacitors (C). These types are collectively known as passive filters, because they do not depend upon an external power supply. Inductors block high frequency signals and conduct low-frequency signals, while capacitors do the reverse. A filter in which the signal passes through an inductor, or in which a capacitor provides a path to ground, presents less attenuation to low frequency signals than high-frequency signals and is a lowpass filter. If the signal passes through a capacitor, or has a path to ground through an inductor, then the filter presents less attenuation to high-frequency signals than low-frequency signals and is a highpass filter. Resistors on their own have no frequency-selective properties, but are added to inductors and capacitors to determine the time constants of the circuit, and therefore the frequencies to which it responds.

The inductors and capacitors are the reactive elements of the filter. The number of elements determines the order of the filter. In this context, an LC tuned circuit being used in a bandpass or band-stop filter is considered a single element even though it consists of two components.

At high frequencies (above about 100 megahertz), sometimes the inductors consist of single loops or strips of sheet metal, and the capacitors consist of adjacent strips of metal. These inductive or capacitive pieces of metal are called stubs.

(a) Single element types

A lowpass electronic filter is realised by an RC circuit. The simplest passive filters consist of a single reactive element. These are constructed of RC, RL, LC or RLC elements.

The quality or "Q" factor is a measure that is sometimes used to describe simple bandpass or bandstop filters. A filter is said to have a high *Q* if it selects or rejects a range of frequencies that is narrow in comparison to the centre frequency. *Q* may be defined for bandpass and bandreject filters as the ratio of centre frequency divided by 3dB bandwidth. It is not commonly employed with higher order filters where other parameters are of more concern, and for high-pass or low-pass filters Q is not normally related to bandwidth.

- **L filter:**

  An L filter consists of two reactive elements, one in series and one in parallel.

- **T and π filters:**

  Three element filters can have a 'T' or 'π' topology and in geometries, a lowpass, highpass, bandpass, or bandstop characteristic is possible. The components can be chosen symmetric or not, depending on the required frequency characteristics. The highpass T filter in the illustration, has a very low impedance at high frequencies, and a very high impedance at low frequencies. That means that it can be inserted in a transmission line, resulting in the high frequencies being passed and low frequencies being reflected. Likewise, for the illustrated lowpass π filter, the circuit can be connected to a transmission line, transmitting low frequencies and reflecting high frequencies.

(b) Multiple element types

Multiple element filters are usually constructed as a ladder network. These can be seen as a continuation of the L, T and π designs of filters. More elements are needed when it is desired to improve some parameter of the filter such as stop-band rejection or slope of transition from pass band to stop band.

1.2.1.2 Active filters

Active filters are implemented using a combination of passive and active (amplifying) components, and require an outside power source. Operational amplifiers are frequently used in active filter designs. These can have high Q, and can achieve resonance without the use of inductors. However, their upper frequency limit is limited by the bandwidth of the amplifiers used.

1.2.1.3 Digital filters

A finite impulse response filter Digital signal processing allows the inexpensive construction of a wide variety of filters. The signal is sampled and an analog to digital converter turns the signal into a stream of numbers. A computer program running on a CPU or a specialized DSP (or less often running on a hardware implementation of the algorithm) calculates an output number stream. This output can be converted to a signal by passing it through a digital to analog converter. There are problems with noise introduced by the conversions, but these can be controlled and limited for many useful filters. Due to the sampling involved, the input signal must be of limited frequency content or aliasing will occur.

1.2.1.4 Quartz filters and piezoelectric

In the late 1930s, engineers realized that small mechanical systems made of rigid materials such as quartz would acoustically resonate at radio frequencies, i.e. from audible frequencies (sound) up to several hundred megahertz. Some early resonators were made of steel, but quartz quickly became favoured. The biggest advantage of quartz is that it is piezoelectric. This means that quartz resonators can directly convert their own mechanical motion into electrical signals. Quartz also has a very low coefficient of thermal expansion which means that quartz resonators can produce stable frequencies over a wide temperature range. Quartz crystal filters have much higher quality factors than LCR filters. When higher stabilities are required, the crystals and their driving circuits may be mounted in a "crystal oven" to control the temperature. For very narrow band filters, sometimes several crystals are operated in series.

Engineers realized that a large number of crystals could be collapsed into a single component, by mounting comb-shaped evaporations of metal on a quartz crystal. In this scheme, a "tapped delay line" reinforces the desired frequencies as the sound waves flow across the surface of the quartz crystal. The tapped delay line has become a general scheme of making high Q filters in many different ways.

1.2.2 Classification of Filters by Transfer Function

The transfer function of a filter is the ratio of the output signal to that of the input signal as a function of the complex frequency. The transfer function of all linear time invariant filters generally shares certain characteristics.

Since the filters are constructed of discrete components, their transfer function will be the ratio of two polynomials in, i.e. a rational function of numerator and denominator. The order of the transfer function will be the highest power of encountered in either the numerator or the denominator.

The polynomials of the transfer function will all have real coefficients. Therefore, the poles and zeroes of the transfer function will either be real or occur in complex conjugate pairs. Since the filters are assumed to be stable, the real part of all poles (i.e. zeroes of the denominators) will be negative, i.e. they will lie in the left half plane in complex frequency space.

The proper construction of a transfer function involves the Laplace transform, and therefore it is needed to assume null initial conditions, because And when f(0) = 0 we can get rid of the constants and use the usual expression

An alternative to transfer functions is to give the behavior of the filter as a convolution. The convolution theorem, which holds for Laplace transforms, guarantees equivalence with transfer functions.

Filters may be specified by family and passband. A filter's family is specified by certain design criteria which give general rules for specifying the transfer function of the filter. Some common filter families and their particular design criteria are:

- **Butterworth filter** - no gain ripple in pass band and stop band, slow cutoff.
- **Chebyshev filter (Type I)** - no gain ripple in stop band, moderate cutoff.
- **Chebyshev filter (Type II)** - no gain ripple in pass band, moderate cutoff.
- **Bessel filter** - no group delay ripple, no gain ripple in both bands, slow gain cutoff.
- **Elliptic filter** - gain ripple in pass and stop band, fast cutoff.
- **Gaussian filter** - no ripple in response to step function.

Generally, each family of filters can be specified to a particular order. The higher the order, the more the filter will approach the "ideal" filter. The ideal filter has full transmission in the pass band, and complete attenuation in the stop band, and the transition between the two bands is abrupt (often called brick-wall).

Each family can be used to specify a particular pass band in which frequencies are transmitted, while frequencies in the stop band (i.e. outside the pass band) are more or less attenuated.

- **Lowpass filter -** Low frequencies are passed, high frequencies are attenuated.
- **Highpass filter** - High frequencies are passed, Low frequencies are attenuated.
- **Bandpass filter** - Only frequencies in a frequency band are passed.
- **Bandstop filter** - Only frequencies in a frequency band are attenuated.
- **All pass filter** - All frequencies are passed, but the phase of the output is modified.

The family and passband of a filter completely specify the transfer function of a filter. The transfer function completely specifies the behaviour of a linear filter, but not the particular technology used to implement it. In other words, there are a number of different ways of achieving a particular transfer function when designing a circuit.

1.2.3 Classification of Filters by Topology

Electronic filters can be classified by the technology used to implement them. Filters using passive filter and active filter technology can be further classified by the particular electronic filter topology used to implement them.

Any given filter transfer function may be implemented in any electronic filter topology. Some common circuit topologies are:

- Cauer topology - Passive
- Sallen Key topology - Active
- Multiple Feedback topology - Active
- State Variable Topology - Active
- Biquadratic topology biquad filter - Active

In recent years, the broadband communication systems, such as video applications, wireless telephony and computer networking has been a strong driving force for the IC technology. Filters constitute a small part of a complete communication system. So, they need to be low in power, small in size and they must not limit the performance of the overall system. Filter implementation can be either active or passive. In general, for very high frequencies (> 10 GHz), the distributed passive filter is typically used and for the medium frequency range (between 100 MHz and 1 GHz), lumped passive filter is suitable. But at higher frequency the power consumption and noise contribution of active circuitry would be too large for practical purposes and IC implementation too challenging. For very low frequency applications (< 10 MHz) like video applications, active filters are suitable, in which the prototype element values are too large to be implemented using discrete components. In this dissertation active filters have simulated using modified current feedback operational amplifier which is a step forward to this rapidly growing communication systems field.

1.3 INVERSE FILTER

In communication, control and instrumentation systems an electrical signal get altered by many linear or nonlinear transformation and to recover the distortions of the signal caused by the signal processors or transmission system, A system is required that has inverse transfer characteristics of the original system. So inverse filter can correct this distortion because it has frequency response, which is the reciprocal of the frequency response of the system that caused the, distortion. There are several schemes for performing inverse digital filtering In digital signal processing, but for realising continuous-time analogue inverse filters there are very few methods/circuits.

CHAPTER 2

LITERATURE REVIEW

2.1 LITERATURE REVIEW ON INDUCTOR

The use of inductors in analog circuits has many disadvantages, mentioned in the paper [1]. One among them is that the analog filters using inductor works well at high frequencies, however, in low frequency applications that is in the frequency range (0- 20 kHz), the inductors cannot be used for the reasons [2] These analog filters are based on the op-amp-RC resonator circuit obtained by replacing the inductor L in the LCR resonator by a simulated inductor. Fortunately, active circuits can sometimes synthesize the equivalent of an inductor with high Q. For this purpose, a great number of active circuits for realizing grounded inductors were reported in the literature [3]-[5].

A new simulated inductor employing two resistors, one capacitor, and only a single gain-variable third-generation current conveyor (GVCCIII) is proposed [4]. A circuit consisting of a single unity gain amplifier, two resistors (R1, R2), and a capacitor (C) is presented for realization of a grounded inductor for integrated circuits. The circuit behaves as an inductor with inductance L = CR1R2 in literature [6].

Using the differential voltage controlled current source elements and only one passive grounded-capacitor component, novel lossless grounded and floating inductor simulating networks are reported. A simple method to compensate for the instability in the realised admittance function caused by the negative shunt resistor owing to the nonideality in the active element is also proposed in [7].

A simple circuit for the realization of electronically tunable ideal grounded inductor consists of two current controlled conveyors along with a grounded capacitor. Its application for realization of multifunctional filter is also demonstrated in [8]. Multifunctional filter (low pass, band pass and high pass filters) is designed using proposed grounded inductor realization which provides standard responses without any constrains in terms of matching conditions.

A novel current controllable floating or grounded inductor simulator circuit with minimum elements is proposed. The circuit uses two dual output- second generation current controlled conveyors (DO-CCCIIs) with parasitic resistance at terminal X and one grounded capacitor externally connected. Proposed circuit is suitable for fully integrated circuit design [9]. The non ideality effects such as non ideal gain and parasitic-impedance effects on the low-frequency performance of two new grounded inductor simulators are investigated in great depth. In addition, some low-frequency performance improvement methods for grounded inductor simulators are discussed. Both of the grounded simulated inductors derived from the previously published ones employ a minimum number of grounded passive components and plus-type second-generation current conveyors. Four second order voltage mode universal-filter topologies, three of which are novel, are derived from newly and previously developed simulated inductors as application examples [29].

A novel circuit for realizing floating inductance, floating capacitance, floating frequency dependent negative resistance (FDNR) and grounded to floating admittance converter depending on the passive component selection is proposed. The proposed simulator employs second-generation current conveyors (CCIIs), differential voltage current conveyor (DVCC) and only grounded passive elements . The non-ideal current and voltage gains as well as parasitic impedance effects on the proposed circuit are investigated in literature [40].

New approaches for realizing high quality factor continuous-time asymmetric-slope second order band-pass filters based on concepts of fractional-order filters. Two non-conventional transfer functions and two possible circuits, one based on a floating Frequency Dependent Negative Resistor (FDNR) and one based on a floating inductor both using a fractional capacitor, are proposed in literature [10].

There are many applications in communication, control and instrumentation systems where inverse filters are required to correct the distortions of the signal caused by the signal processors or transmission system. This correction can be done by using an inverse filter, which has frequency response, which is reciprocal of the frequency response of the system that caused the distortion. New configurations for realizing inverse low pass, inverse band pass, and inverse high pass and inverse band reject filters using commercially available current feedback op-amps (CFOA) with an accessible z-terminal such as AD844 are presented in literature [11].

Operational trans-impedance amplifiers, popularly known as current-feedback operational amplifiers (CFOA), offer several advantages over conventional voltage-mode op-amps (VOA)

such as wide and nearly constant bandwidth (BW) for variable gains, very high slew rate, and ease of realizing various functions with the least possible number of passive components Four new single-input–multi-output-type universal bi-quad filters using current-feedback operational amplifiers (CFOAs) are presented in literature [12] , which realize all the five standard filter functions (namely, low-pass, band-pass, high-pass, notch, and all-pass) from the same configuration, provide tunability of the various filter parameters in all the five responses, and offer a resistive or ideally infinite input impedance while employing only four CFOAs and two grounded capacitors as preferred for integrated circuit implementation.

Using a minimum number of passive components, i.e., new grounded and floating inductance simulators, grounded capacitance multipliers, and frequency-dependent negative resistors (FDNRs) based on one/two modified current-feedback operational amplifiers (MCFOAs), are proposed. The type of the simulators depends on the passive element selection used in the structure of the circuit without requiring critical active and passive component-matching conditions cancellation constraints. In order to show the flexibility of the proposed MCFOA, a single-input three-output (SITO) voltage-mode (VM) filter, two three-input single-output (TISO) VM filters, and an SITO current-mode (CM) filter employing a single MCFOA are reported in literature [13].

The active devices that have been used for the realization of current-mode circuits include current conveyor, current feedback operational amplifier, operational trans-conductance amplifier and four-terminal floating. Recently, attention has concentrated on the use of current operational amplifier (COA) as true current-mode active element in current-mode signal processing circuits. Both input terminals of COA are characterized by low impedance, thereby eliminating response limitations incurred by capacitive time constants. The input terminals are internally grounded leading to circuits that are insensitive to the stray capacitances. The output terminals of COA exhibit high impedance so that COA-based current-mode circuits can easily be cascaded without additional buffers. For ideal operation, the open-loop current gain approaches infinity forcing the input currents to be equal. A current-mode first-order all pass filter configuration is proposed. The presented circuit uses a single current operational amplifier (COA), a resistor and a capacitor, which are of minimum number. High output impedance of the proposed filter enables the circuit to be cascaded without additional buffers. The proposed circuit is insensitive to parasitic input capacitances and input resistances due to internally grounded input terminals of COA. It does not impose any component matching constraint in analog signal processing circuits. Non-ideal effects of COA to the first-order all pass network are also investigated. To demonstrate the performance of the proposed filter a new current-mode quadrature oscillator is given as an application example in literature [14].

The gyrator-based OTA simulated floating inductor can be divided into two categories; 3-OTA and 4-OTA structure which perform identically in the ideal phenomena where all OTA's nonidealties i.e. parasitic elements, effect of finite open-loop bandwidth and noise have been neglected. It has been found that the 4-OTA-based floating OTA counterpart in the practical phenomena cited nonidealties included [15]. Using a complementary BJT process an OTA has been designed by proposing some broad-banding and phase compensation techniques To demonstrate the application in OTA-C continuous-time filter design, a 225 MHz OTA-C low pass filter and a 250 MHz OTA-C bi-quadratic band pass filter are realized and presented in[16]. A current-mode first-order all pass filter configuration is proposed. The presented circuit uses a single current operational amplifier (COA), a resistor and a capacitor, which are of minimum number. High output impedance of the proposed filter enables the circuit to be cascaded without additional buffers. The proposed circuit is insensitive to parasitic input capacitances and input resistances due to internally grounded input terminals of COA. It does not impose any component matching constraint in analog signal processing circuits. Non-ideal effects of COA to the first-order all-pass network are also investigated in literature [17].

2.2 LITERATURE REVIEW ON FILTERS

A new design approach for active RC bi quadratic filter circuits using differentiators is presented. The method exploits current-feedback operational amplifiers as active elements. Compared to conventional op-amps, these amplifiers are characterized by a very high bandwidth and offer more efficient frequency compensation [18]. A new voltage-mode second-order low pass OTA-C filter is proposed. The filter is integrator based and uses only one OTA and two grounded capacitors. A multiple-input operational trans-conductance amplifier (OTA) for the simulation of the proposed filter is also described. The filter has been simulated using a 0.25m CMOS process in [19]. Inductorless active filters using piezoelectric resonators are described. Because of the excellent frequency stability of piezoelectric resonators (quartz crystals in particular), these filters are more suitable for realizing narrow bandwidth responses than are active RC filters which are restricted to relatively low Q applications by the limited tracking capability of the RC components. In addition, because the active filters, the inductors and transformers which are often required in passive piezoelectric filter are eliminated [20]. Simulation of passive component of the inductor using GIC (generalized impedance converter) and the application of the simulated inductor in analog filters has been described in [21]. The analog filters use active component the operational amplifiers, resistors and the simulated inductor. The design of various analog filters is based on the basic LCR resonator circuits. Various types of filters are realized and the frequency response is obtained in [21].

Passive and active analog filters with differential input and differential output are implemented by coupling single ended filters and provide very high common-mode rejection ratios. This makes it possible to place these filters before differential amplifiers, thus improving interference rejection and noise reduction [22].

Resistance capacitance (RC) filters are also low in output thermal noise. The design procedure of second and third order low sensitivity all pole filters, using impedance tapering, has already been published. The component values selected for impedance tapering account for the considerable decrease in output thermal noise. The method of Zurada and Bialko was used to determine output noise spectral density and total rms output noise of filters. Passive elements and operational amplifier are represented by substitute noise models. The noise contribution of each device to the output node is calculated using transfer functions [23]. The linearity of conventional active-RC filters is limited by the operational trans-conductance amplifiers (OTAs) used in the integrators. Trans-conductance capacitance (Gm-C) filters are fast and can be linear- however, they are sensitive to parasitic capacitances [24]. We explore the Gm-assisted OTA-RC technique, which is a way of combining Gm-C and active-RC integrators in a manner that enhances the linearity and speed of the latter, while adding negligible extra noise or power dissipation. Measurements from a fifth-order Chebyshev filter with 20 MHz bandwidth, designed in a 0.18 m CMOS process, demonstrate the efficacy of Gm-assistance in an active-RC integrator. A new MOS-C band-pass-low-pass filter using the current feedback operational amplifier (CFOA) is presented [25]. The filter employs two CFOA's, eight MOS transistors operating in the non saturation region, and two grounded capacitors. The proposed MOS-C filter has the advantage of independent control of Q. [26] novel lossless and lossy grounded parallel inductance simulators are reported .All grounded inductor simulator circuits employing only a single DXCCII and three passive component proposed. The proposed topologies realized all grounded parallel inductance variations .To demonstrate the performance of the presented DXCCII based parallel inductance simulators, we used one of the circuits to construct a third order high-pass filter, a voltage-mode band-pass filter and LC oscillator. The proposed DXCCII and its applications are simulated using CMOS 0.35 mm technology.

The current conveyors (CCIIs) operate linearly under certain conditions. Violation of these conditions causes nonlinear distortion in active filters involving CCIIs. On the other hand, voltage-mode active-RC filters are considerably important in filter design. Considering these facts, a simple formula is derived for maximum input signal amplitude preventing nonlinear distortion in voltage-mode active-RC filters involving current conveyors [27]. New realizations of grounded negative capacitance, using Current Feedback Operational Amplifiers (CFOAs), two resistors and one capacitor has shown in [28]. All the proposed realizations are canonic in the number of passive components and do not require any critical component matching condition [28]. Application examples in capacitive cancellation schemes and resistance-controlled low-frequency quadrature sinusoidal oscillator design are provided.

A circuit can function both as a quadrature oscillator and a universal biquad filter (lowpass, highpass, bandpass). When the circuit functions as a universal biquad filter, the quality factor and pole frequency can be tuned orthogonally via the input bias currents. When it functions as a quadrature oscillator, the oscillation condition and oscillation frequency can be adjusted independently by the input bias currents. The proposed circuit can work as either a quadrature oscillator or a biquad filter without changing the circuit topology. The amplitude of the proposed oscillator can be independently controlled via the input bias currents. The proposed oscillator can be applied to provide amplitude modulated/amplitude shift keyed signals with the above-mentioned major advantages [30].

Four new topologies, all separately able to realize eight main types of inductors employing only grounded passive components, are presented. For the topologies, ideal and non-ideal impedance functions are given and an application example is shown [31]. Novel universal voltage mode filter with only two current conveyers with only three inputs and one output is shown in [33]. A novel floating inductance simulator, based on second generation current conveyors, is presented. The circuit shows the capability of regulation and, in theory, cancellation of the undesired inductance series resistance, with consequent increase in the low frequency operating range. Simulation results are included to prove the effectiveness of this solution [34]. The filters are classified in two classes. The class I filters have floating passive elements, whereas class II filters have all resistors and capacitors grounded. Each class of filters includes two subclasses based on the filter's capability to have independent control of the filter quality factor or not. The effects of non-ideal second generation current conveyors are briefly discussed [35]. A new three-input single-output voltage-mode universal bi-quadratic filter with high-input impedance using only three plus-type second-generation current conveyors (CCIIs) is presented. The proposed configuration uses only two capacitors and two resistors and can realize all the standard filter functions, that is, high-pass, band-pass, low-pass, notch, and all-pass filters without changing the passive elements [36]. A new current-conveyor-based (CC) biquad, equivalent to the well known KHN circuit, is introduced. The proposed circuit employs exactly the same number (five) of CCs and resistors (six/ seven) along with two grounded capacitors as in the two CC biquads [37].

Recently, it has been shown by **Paul Dey Patranabis** that a floating NIC can be realized with two second generation current conveyors (CC), without constraints. In this communication it is shown that more useful floating generalized positive immittance converter/inverter elements too can be realized with only two second generation current conveyors [38].Voltage mode filters using two current conveyers are presented in [39]. The low-sensitivity realization of band-pass (BP) active resistance-capacitance (RC) filters using "uniform modified leap frog"(UMLF) structure. Sensitivities with respect to filter passive components are investigated [41].

New balanced input balanced output switched capacitor (SC) band-pass low-pass filter using the current feedback operational amplifier (CFOA) is presented [42]. The proposed SC bi-quad filter is based on new balanced input balanced output lossless and lossy integrators. The introduced filter has the advantage of independent control of Q, the balanced operation.

2.3 LITERATURE REVIEW ON INVERSE FILTERS

In [43], a general procedure is presented by **A. Leuciuc** for obtaining the inverse transfer function for linear dynamic systems and the inverse transfer characteristic for non-linear resistive circuits by using nullors (a nullor is a two-port network element characterised by $v_1 = 0$, $i_1 = 0$ for the input port and v_2 = arbitrary, i_2 = arbitrary for the output port) as basic building blocks. The method has been explained through examples of a Sallen Key highpass filter and inverse of a half wave rectifier. In [44], a technique for transforming current-mode four terminal floating nullor (FTFN) based inverse filter from the voltage-mode filter is given by **B. Chipipop and W. Surakampontorn**. The realisation procedure utilises network theory concepts related to nullors and RC:CR (capacitor-resistor) dual transformation. Due to the use of dual transformation, this approach can only be applied to planar circuit. **H. Y. Wang** and **C. T. Lee** presented another easier procedure, by the use of adjoint transformation for deriving current-mode FTFN-based inverse filter from the voltage-mode filter and it is applicable to nonplanar circuits [45].

All the above procedure in [43] [44] [45] can be utilized for obtaining single-input single-output inverse filters. In [46] and [47] various inverse current-mode and voltage-mode filters are proposed by **M. T. Abuelma'atti, S. S. Gupta, D. R. Bhaskar, R. Senani** and **A. K. Singh**, respectively. But, each circuit presented in [46, 47] has one inverse filter function. In recent literature (Soliman 1996 [48]; Lidgey and Hayatleh 1997 [49]; **Senani** 1998 [50]; **Ananda Mohan** 2003 [51]; Biolek, Senani, Biolkova and Kolka 2008), there is an increased attention on realising various analogue signal processing functions using CM elements such as second-generation current conveyors (CCII+ ; a building block having three terminals x, y and z and having terminal voltages and currents characterised by $i_y = 0$, $v_x = v_y$ and $i_z = +i_x$) and current feedback op-amps (CFOA) a building block having four terminals x, y, z and w which is internally a cascade of CCII+ and a unity-gain voltage follower and, therefore, has its terminal voltages and currents characterised by $i_y = 0$, $v_x = v_y$, $i_z = i_x$, and $v_x = v_z$. Use of CFOAs in circuit design results in attractive features of nearly constant bandwidth for moderate values of gain, higher slew rates (typically in excess of 2000V/m s), relatively higher operational frequency range and possibility of designing various signal processing/signal generation circuits with the least possible number of external passive components without requiring any component-matching conditions or realisation constraints.

Moreover, the use of CFOA is particularly appealing as it can be used to implement both varieties of second generation current conveyor (CCII) (CCII+ as well as CCII-) and also an FTFN (which requires two CFOAs (**Abuelmaatti** 2000 [47])) and is also commercially available as a four-terminal building block such as AD844 from Analog Devices Inc. Motivated by these advantages of CFOAs, four inverse filter configurations using CFOAs (one each for inverse LP, inverse BP, inverse HP and inverse BR) were recently proposed in **Gupta, Bhaskar, Senani** and **Singh** (2009) [48].

A number of new configurations that realise inverse LP, BP, HP and BR filters are presented using both CFOAs and MCFOAs which offer properties superior to the previously published inverse filters of references (Leuciuc 1997; Chipipop and Surakampontorn 1999; Wang and Lee 1999; Abuelmaatti 2000) as well as the recently proposed circuits of Gupta et.al (2009).

CHAPTER 3

CURRENT FEEDBACK OPERATIONAL AMPLIFIER

This chapter describes the basics of CFOA as well as Modified CFOA. Here, CMOS realization of Modified CFOA [52] is discussed and its simulation results are presented. This simulated Modified CFOA is used for designing of inverse filter proposed in this work.

3.1 BASICS OF CFOA

The current feedback operational amplifier otherwise known as CFOA or CFA is a type of electronic amplifier whose inverting input is sensitive to current, rather than to voltage as in a conventional voltage-feedback operational amplifier (VFA). The CFOA was invented by **David Nelson** at Comlinear Corporation.

The circuit symbol of the current feedback operational amplifier (CFOA) is shown in Figure 3.1. It is a four terminal device where Y and X are input, W and Z are output terminals.

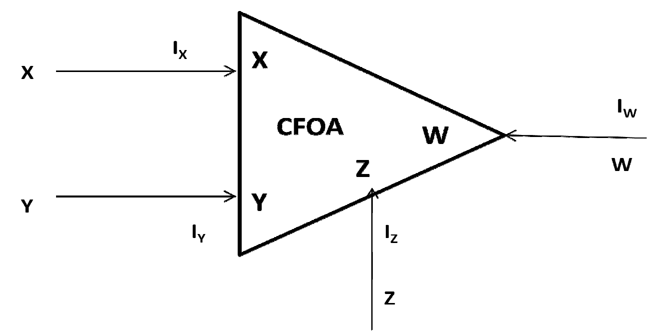


Fig. 3.1: Symbol of CFOA

It is characterized with the following matrix equation

$$\begin{bmatrix} I_Z \\ I_Y \\ V_X \\ V_W \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} I_X \\ V_Y \\ V_Z \end{bmatrix} \tag{3.1}$$

3.1.1 Operation of CFOA

Referring to the schematic shown below, the first section forms the input stage and error amplifier. The inverting input (node where emitters of Q1 & Q2 are connected) is low impedance and hence, sensitive to changes in current. Resistors R1–R4 set up the quiescent bias conditions and is chosen such that the collector currents of Q1 & Q2 are the same. In most designs, active biasing circuitry is used instead of passive resistive biasing, and the non-inverting input may also be modified to become low impedance like the inverting input in order to minimise offsets.

With no signal applied, due to the current mirrors Q3/Q4 & Q5/Q6, the collector currents of Q4 and Q6 will be equal in magnitude if the collector 3ecurrents of Q1 and Q2 are also equal in magnitude. Thus, no current will flow into the buffer's input (or equivalently no voltage will be present at the buffer's input). In practice, due to device mismatches the collector currents are unequal and this results in the difference flowing into the buffer's input resulting in an offset at its output. This is corrected by adjusting the input bias or adding offset nulling circuitry. The second section (Q3–Q6) forms an I-to-V converter. Any change in the collector currents of Q1 and Q2 (as a result of a signal at the non-inverting input) appears as an equivalent change in the voltage at the junction of the collectors of Q4 and Q6. C_s is a stability capacitor to ensure that the circuit remains stable for all operating conditions. Due to the wide open-loop bandwidth of a CFA, there is a high risk of the circuit breaking into oscillations. C_s ensures that frequencies where oscillations might start are attenuated, especially when running with a low closed-loop gain.

The output stage is a buffer which provides current gain. It has a voltage gain of unity (+1 in the schematic).

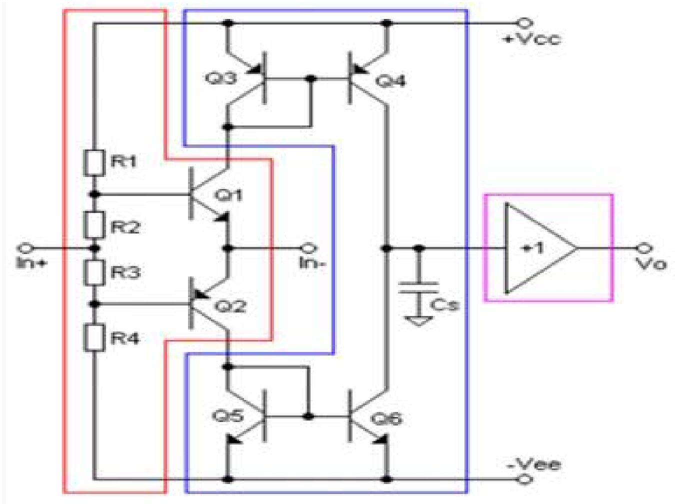


Fig. 3.2: Representative schematic of a current feedback operational amplifier

In simple configurations, such as linear amplifiers, a CFOA can be used in place of a VFA with no circuit modifications, but in other cases, such as integrators, a different circuit design is required. The classic four-resistor differential amplifier configuration also works with a CFOA, but the common mode rejection ratio is poorer than that from a VFA. Internally compensated, VFA bandwidth is dominated by an internal dominant pole compensation capacitor, resulting in a constant gain/bandwidth limitation. CFOAs, in contrast, have no dominant pole capacitor and therefore can operate much more closely to their maximum frequency at higher gain. Stated another way, the gain/bandwidth dependence of VFA has been broken.

In VFAs, dynamic performance is limited by the gain-bandwidth product and the slew rate. CFOA use a circuit topology that emphasizes current-mode operation, which is inherently much faster than voltage-mode operation because it is less prone to the effect of stray node capacitances. When fabricated using high-speed complementary bipolar processes, CFOAs can be orders of magnitude faster than VFAs. With CFOAs, the amplifier gain may be controlled independently of bandwidth. This constitutes the major advantages of CFOAs over conventional VFA topologies.

Disadvantages of CFOAs include poorer input offset voltage and input bias current characteristics. Additionally, the DC loop gains are generally smaller by about three decimal orders of magnitude. Given their substantially greater bandwidths, they also tend to be noisier.

3.2 PROPOSED MODIFIED CFOA (MCFOA)

Differential CFOA is a four terminal device characterized by the matrix equation:

$$\begin{bmatrix} I_Z \\ I_Y \\ V_X \\ V_W \end{bmatrix} = \begin{bmatrix} \alpha_1 & 0 & 0 & 0 \\ 0 & -\alpha_2 & 0 & 0 \\ 0 & 0 & \beta_1 & 0 \\ 0 & 0 & 0 & \beta_2 \end{bmatrix} \begin{bmatrix} I_X \\ I_W \\ V_Y \\ V_Z \end{bmatrix} \tag{3.2}$$

As it can be seen from above equation, the MCFOA is different from the conventional current-feedback operational amplifier (CFOA) because the W terminal current of the MCFOA is copied to the Y terminal in the opposite direction. However, it is well known that the Y-terminal current of the conventional CFOA is equal to zero.

Fig. 3.3 and 3.4 show the symbol and construction using commercially active devices of modified CFOA (MCFOA) respectively.

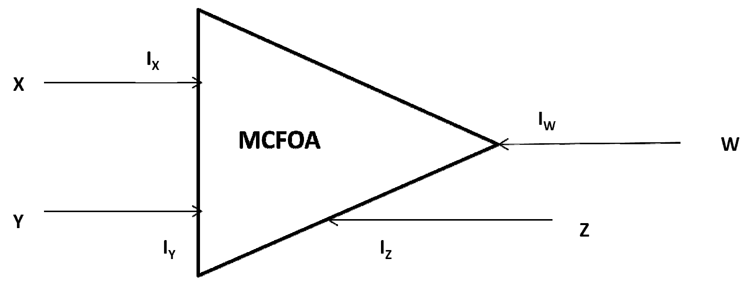


Fig. 3.3: Symbolic representation of the MCFOA

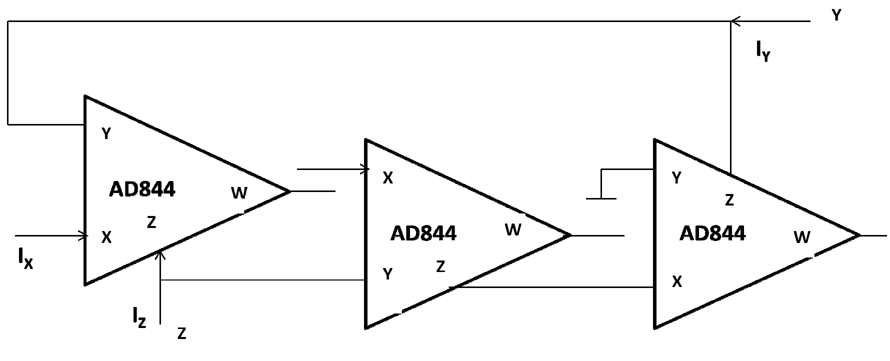


Fig. 3.4: MCFOA construction using commercially available active devices

3.3 CMOS REALIZATION OF MODIFIED CFOA

CMOS realization of modified CFOA proposed in [13] is shown in Fig. 3.5. The output resistances of the transistors M_9, M_{10}, M_{19}, and M_{20} in the MCFOA of Fig. 2 are assumed to be equal to r_0 and, similarly, the output resistances of the transistors M_{11} , M_{12} , M_{19} , and M_{20} of the MCFOA are equal to r'_0 . Thus, the resistances seen at terminals Y, Z, X, and W are respectively calculated as

$$R_Y = \frac{r_{03}r_{024}}{r_{08}+r_{024}} \tag{3.3}$$

$$R_Z = \frac{r_{03}r_{018}}{r_{03}+r_{018}} \tag{3.4}$$

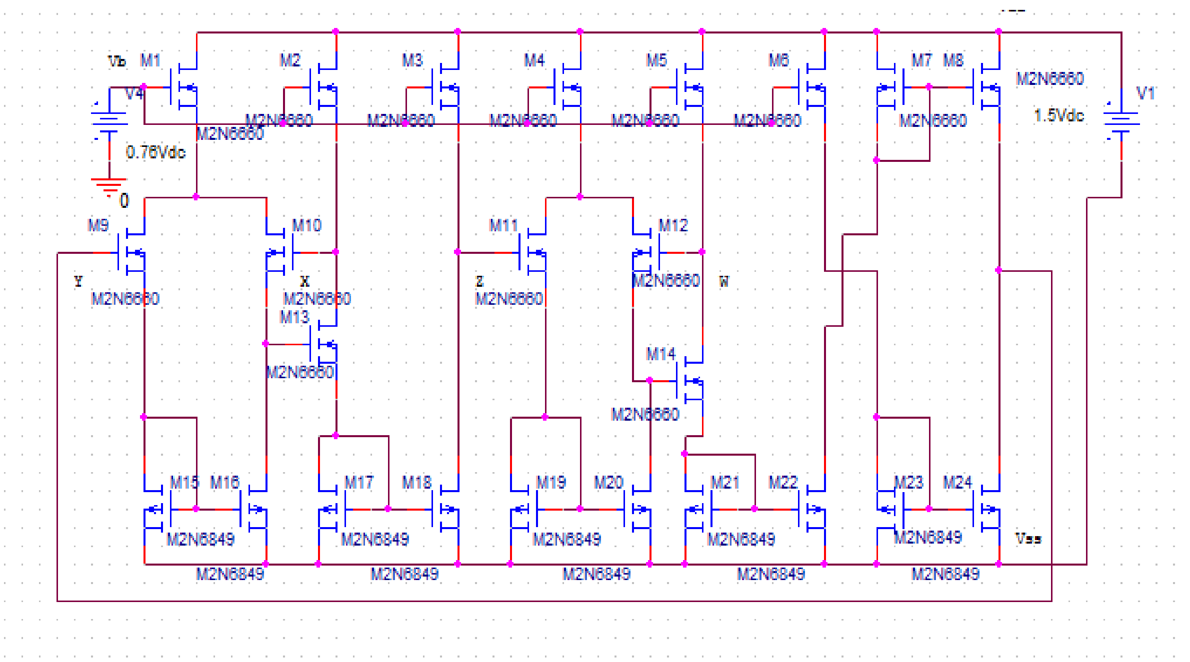


Fig. 3.5: Developed CMOS structure for the MCFOA

$$R_X = \left(\frac{r_{013}+\frac{r_{017}}{1+g_{m17}r_{017}}}{1+r_{013}g_{m13}\left(1+\frac{r_0}{2}g_{m10}\right)}\right)||r_{02} \tag{3.5}$$

$$\cong \frac{2}{g_{m13}g_{m10}r_0}$$

$$R_W = \left(\frac{r_{014}+\frac{r_{021}}{1+g_{m21}r_{021}}}{1+r_{014}g_{m14}\left(1+\frac{r'_0}{2}g_{m12}\right)}\right)||r_{05} \tag{3.6}$$

$$\cong \frac{2}{g_{m12}g_{m14}r'_0}$$

Here, r_{0i} and g_{mi} are the output resistance and transconductance of the i^{th} CMOS transistor, respectively. In (3.3) and (3.6), it is assumed that $g_{m17}r_{017} >> 1$ and $g_{m21}r_{021} >> 1$, respectively. From (3.3)–(3.6), it can be seen that, while the terminals Y and Z have very high resistances, the terminals X and W exhibit low resistances due to the feedback loops composed of the transistors (M_{10}, M_{13}) and (M_{12}, M_{14}), respectively.

The proposed MCFOA can also be compared with a recently reported gain-variable third-generation current conveyor (GVCCIII) [2] and will exhibit its superiority as follows. The GVCCIII is defined by the following matrix equation:

$$\begin{bmatrix} V_X \\ I_Y \\ I_Z \end{bmatrix} = \begin{bmatrix} \beta & 0 & 0 \\ 0 & -\Upsilon & 0 \\ 0 & \pm\alpha & 0 \end{bmatrix} \begin{bmatrix} V_Y \\ I_X \\ V_Z \end{bmatrix} \qquad (3.7)$$

The + and - signs of above equation are used for the plus-type GVCCIII (GVCCIII+) and minus-type GVCCIII (GVCCIII-), respectively. It is observed from equation that the GVCCIII is a three terminal active device while the MCFOA is a four-terminal device. In other words, both the GVCCIII and MCFOA have terminals X, Y, and Z, but the MCFOA also has the terminal W. Also, the current gains and in a GVCCIII are constant and ideally equal to unity, but the voltage gain can be controlled by selecting appropriate external resistors. Contrary to variable voltage gain of the GVCCIII, both of the voltage gains and of the MCFOA are constant and ideally equal to unity. On the other hand, using a single GVCCIII or MCFOA as well as one capacitor and two resistors, a grounded inductor realization is possible. Nevertheless, inductor simulator with GVCCIII employs floating passive components and needs a resistive element matching condition [2].

3.4 SIMULATION RESULTS OF MODIFIED CFOA

For simulation CMOS implementation of modified CFOA proposed in [13] is used. The SPICE simulation is performed using 0.25µm, Level 3, TSMC CMOS process parameters provided by MOSIS (AGILENT) and supply voltages taken are $V_{DD} = -V_{SS} = 1.25V$ and Biasing voltage $V_B = .8V$. Transistors aspect ratios are reported in Table 3.1.

Table 3.1: Aspect ratio of the CMOS transistor used in the MCFOA of Figure 3.5

PMOS Transistors	**$W_{(\mu m)}/L_{(\mu m)}$**
M_1, M_4 and M_9, M_{10}, M_{11} and M_{12}	1.0/0.25
M_2, M_3, M_5, M_6, M_7 and M_8	2.0/0.25
M_{13} and M_{14}	4.0/0.25
NMOS Transistors	**$W_{(\mu m)}/L_{(\mu m)}$**
M_{15}, M_{16}, M_{17}, M_{18}, M_{19}, M_{20}, M_{21}, M_{22}, M_{23} and M_{24}	0.5/0.25

The commercial current feedback amplifiers AD844 macro model with ± 12 V voltage supply is used to realize the CFOA in figure 3.1.

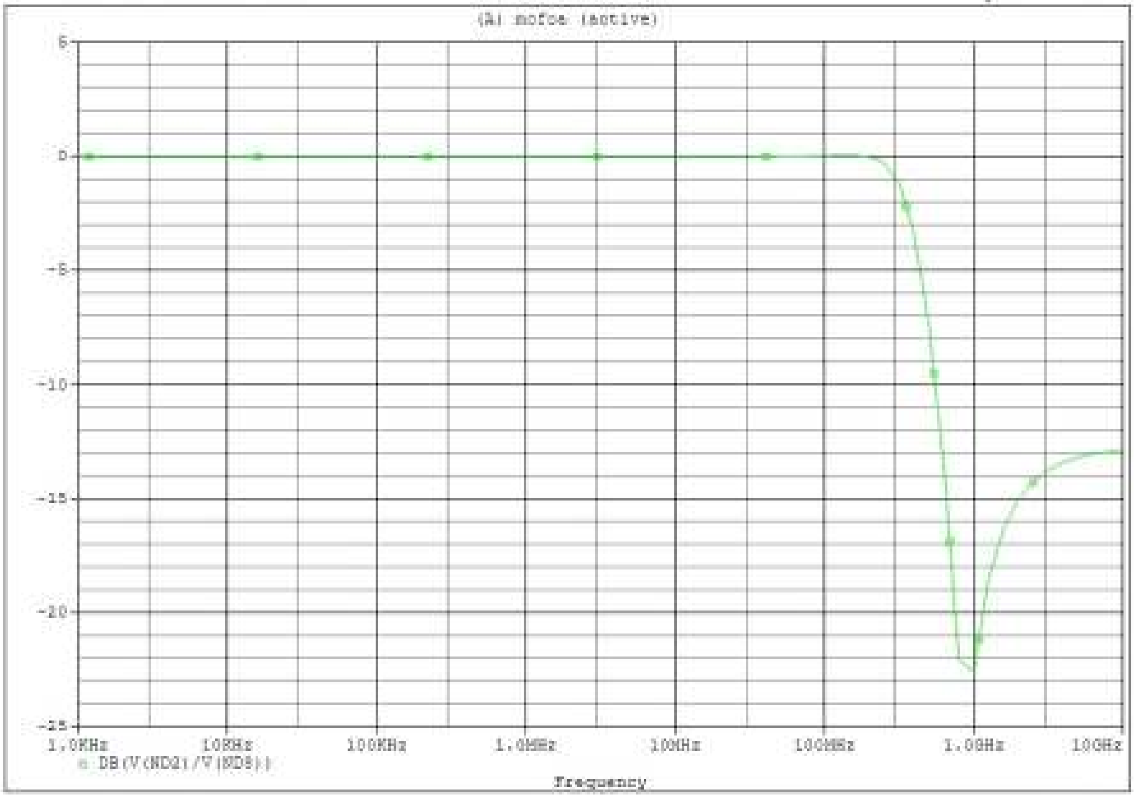


Fig. 3.6: Variation of the voltage gains of the MCFOA against frequency

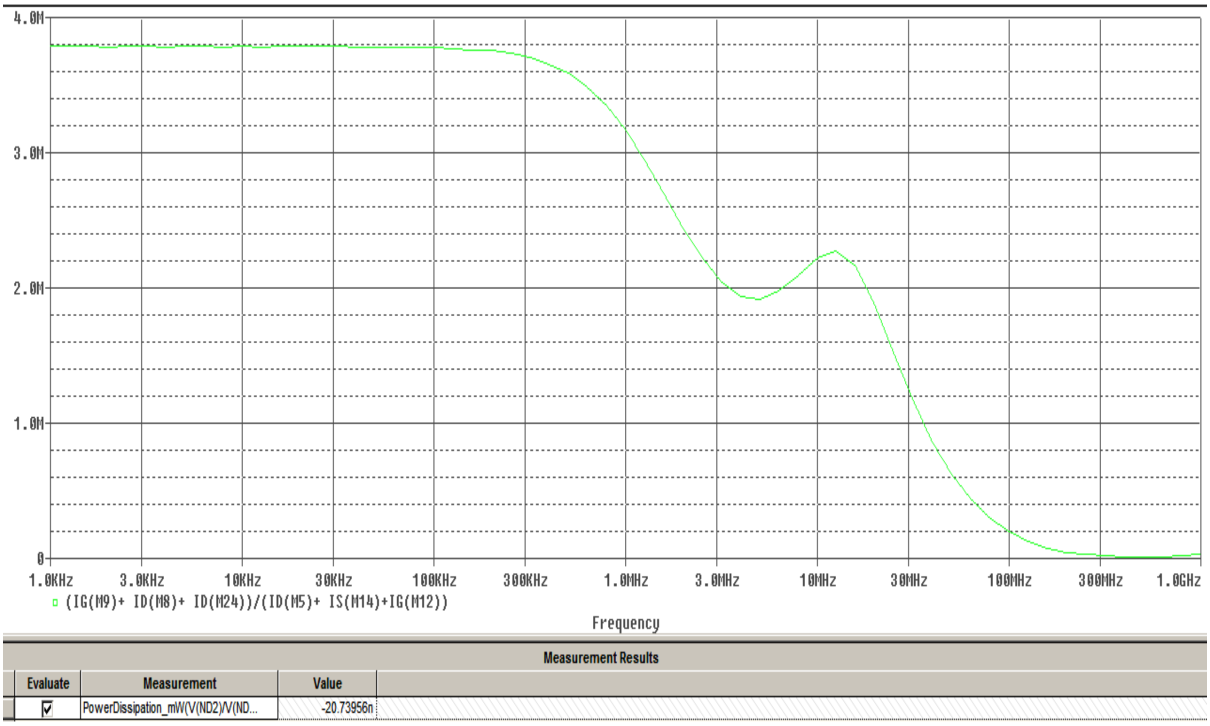


Fig. 3.7: Variation of the current gains of the MCFOA against frequency

The variation of the voltage gains of the MCFOA against frequency is shown in figure 3.6 and bandwidth of the voltage gains obtained is 1.75 GHz. The total power dissipation of the MCFOA is calculated as 768.52 mw.

The Variation of the current gains of the MCFOA against frequency is shown in figure 3.7 and bandwidth of the current gains obtained is 1.65 GHz. The total power dissipation of MCFOA is calculated as 20.73 mw.

CHAPTER 4

SIMULATED INDUCTOR AND FILTERS USING THE PROPOSED MCFOA

The simulated inductors proposed in [3]–[5] are composed of a single active device, i.e. a modified inverting second generation current conveyor (MICCII) in [3], a CCII in [4], and a first generation current conveyor (CCI) in [5]. However, they use a floating capacitor as well as floating resistors and suffer from component matching requirements. The grounded synthetic inductors reported in [6]–[12] use two active devices, and the circuit introduced in [13] employs three active components. Furthermore, the circuits presented in [6] [7] [13] employ a floating passive component. Fortunately, proposed simulator circuits with a single MCFOA [13] employ only grounded passive elements which are canonical in the number and require no critical component matching conditions which in turn make them suitable for fully IC technology.

4.1 GROUNDED INDUCTOR USING MCFOA

The proposed grounded simulator circuits using a single MCFOA are shown in figure 4.1. The straight forward analysis of the networks gives the following input admittances for the circuit of figure 4.1:

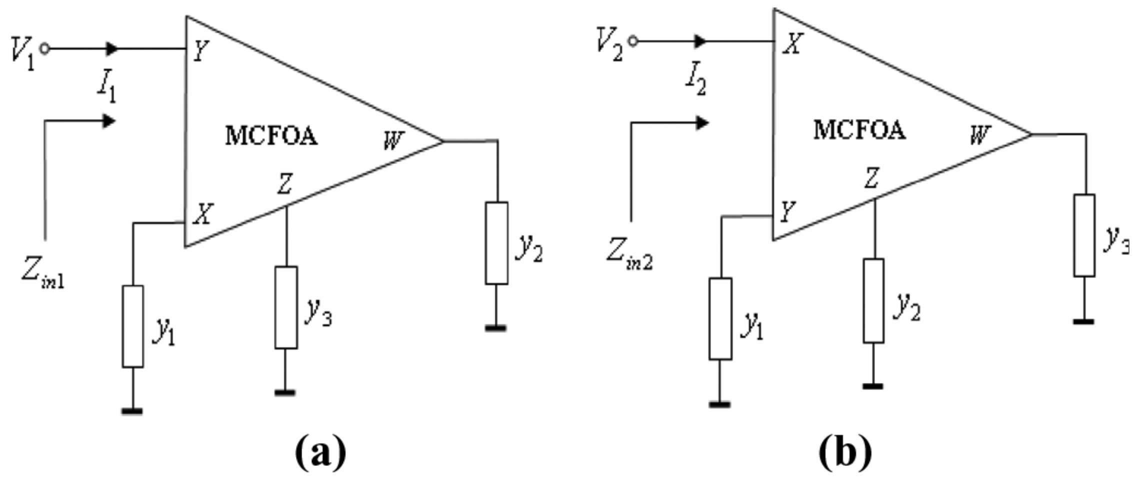


Fig. 4.1: (a) Proposed First Circuit (Y terminal input) and
(b) Second Circuit (X terminal input)

$$Y_{in1} = \frac{1}{Z_{in1}} = \frac{I_1}{V_1} = \frac{y_1 y_2}{y_3}\alpha_1\alpha_2\beta_1\beta_2 \tag{4.1}$$

$$Y_{in2} = \frac{1}{Z_{in2}} = \frac{I_2}{V_2} = \frac{y_1 y_2}{y_3}\frac{1}{\alpha_1\alpha_2\beta_1\beta_2} \tag{4.2}$$

It is seen from equation (4.1) and (4.2) that both circuits have the same input admittances in the ideal case ($\alpha_1 = \alpha_2 = \beta_1 = \beta_2 = 1$). By selecting various passive components for y_1, y_2 and y_3 in the proposed circuits, we can obtain an inductor simulator, a capacitor multiplier.

It is important to note that the impedances of the proposed inductor simulators are both short-circuit and open-circuit stable because all of the non ideal current and voltage gains of them are in the form of multipliers [2]. Apart from this, the first and second simulators have respectively high and low input impedances. Thus, it is good to drive the first and second ones by a voltage source and a current source, respectively. Note that each of the simulator circuits of figure 4.1 can be considered as an admittance element. Then, by using one of them instead of the admittances, or in simulator circuits of figure 4.1 with identical MCFOAs, six different inductor simulators with reduced nonideal gain effects can be obtained as follows.

1) The second simulator (Y_{in2}) is used instead of y_1 of the first simulator.
2) The second simulator (Y_{in2}) is connected instead of y_2 of the first simulator.
3) The first simulator (Y_{in1}) is connected instead of y_3 of the first simulator.
4) The first simulator (Y_{in1}) is connected instead of y_1 of the second simulator.
5) The first simulator (Y_{in1}) is connected instead of y_2 of the second simulator.
6) The second (Y_{in2}) is connected instead of y_3 of the second simulator.

The six cases mentioned above are useful only if the poles of the parasitic impedances are at higher frequencies than those of nonideal gains. In other words, if the values of the parasitic impedances of the MCFOA are reduced considerably using proper design, the nonideal gain effect reduction method will be helpful.

4.2 FLOATING INDUCTOR USING MCFOA

The MCFOA can also be used to realize floating inductor as shown in figure 4.2. The floating inductors like in the figure, which use a minimum number of passive elements (two resistors and a grounded capacitor), can be described by the following equations:

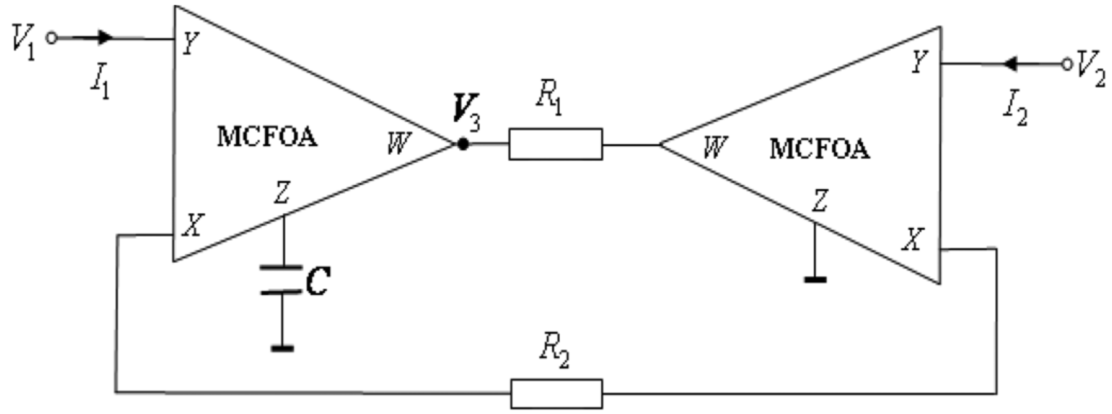


Fig. 4.2: Symbolic representation of the developed floating inductor simulator.

$$I_{in} = I_1 = I_2 \tag{4.3}$$

$$I_{in} = \frac{\pm V_3}{R_1} \tag{4.4}$$

$$\pm V_3 SC = \frac{V_1 - V_2}{R_2} \tag{4.5}$$

$$V_{in} = V_1 - V_2 = SCR_1R_2I_{in} = SL_{eq}I_{in} \tag{4.6}$$

$$\text{Where } L_{eq} = CR_1R_2 \tag{4.7}$$

Here, $V_{in} = V_1 - V_2$ and $I_{in}$are the input voltage and current of the inductor simulator, respectively. Moreover, V_3 is the internal node voltage of the simulated inductor.

To prove the performance of the developed floating inductor a RL series circuit has been designed using the developed floating inductor and this RL series circuit will results as a low pass filter when output is taken across the resistor and will show the Highpass filter response when the output is taken across the inductor. We can calculate the value of simulated floating inductor by measuring the value of cut off frequency of the Lowpass or the Highpass filter and can compare this value of inductor with the value of inductor obtained from equation 4.7.

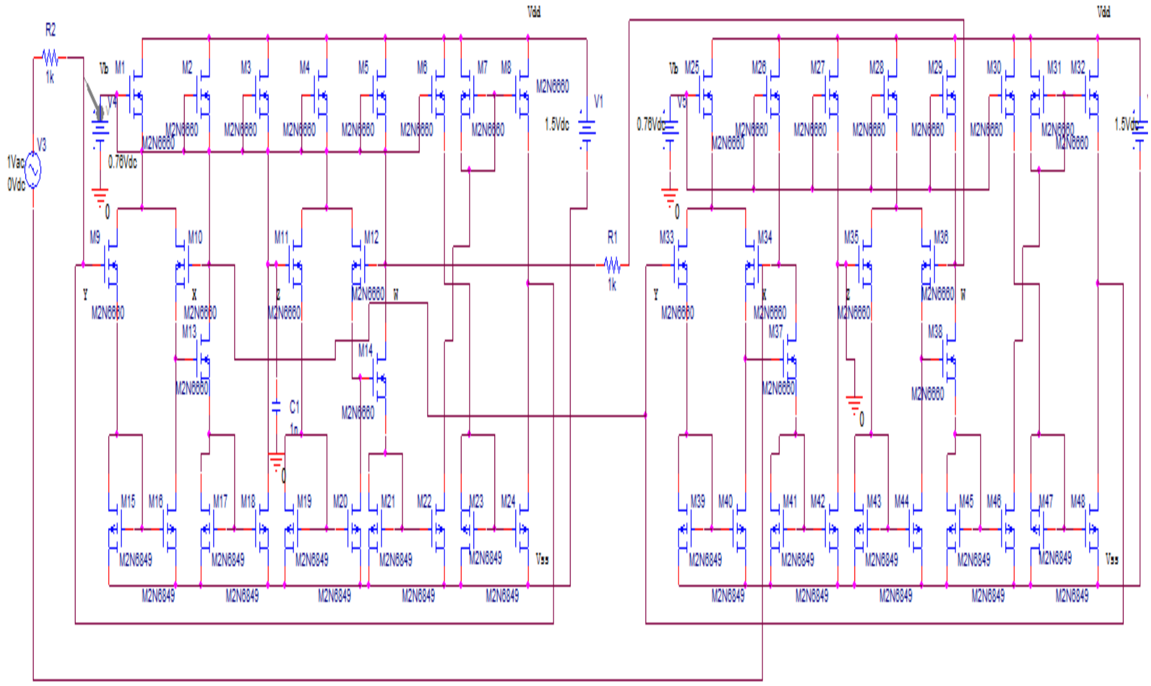


Fig. 4.3: Developed floating inductor simulator

4.3 PROPOSED VOLTAGE MODE FILTERS EMPLOYING A SINGLE MCFOA

The proposed MCFOA can be used easily to realize VM multifunction filters. Contrary to the previously published SITO VM universal filters which use two CCIIs [25], [26] or three CCIIs [27]–[29], the proposed ones employ only a single MCFOA. Both plus-type and minus-type CCII can be easily realized using a MCFOA. For example, if W terminal of the MCFOA is left open, it behaves like a CCII+. Similarly, if the X terminal of the MCFOA is left open, it behaves like a CCII- in which $I_z = 0$, $I_y = -I_w$, and $V_W = V_Z$.

4.3.1 Proposed SITO Voltage Mode Filter Using Single MCFOA

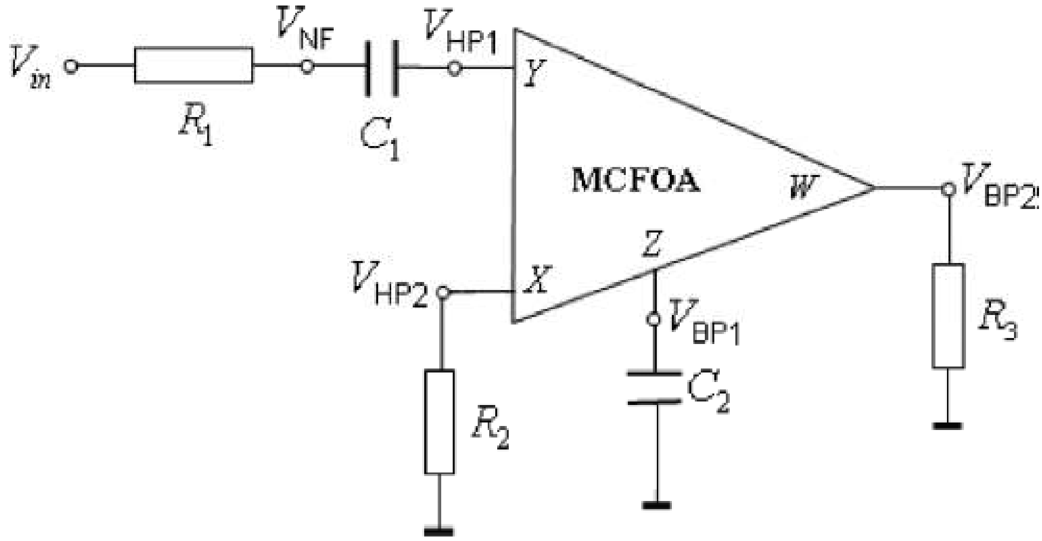


Fig. 4.4: Symbolic representation of proposed SITO VM filter

The proposed SITO VM filter using only one MCFOA is shown in figure 4.4. Routine analysis of the filter depicted in figure 4.4 yields the following filter TFs simultaneously:

$$V_{in} - V_{HP1} = (R_1 + \frac{1}{SC_1})\, I_Y$$

$$V_{in} - V_{HP1} = (R_1 + \frac{1}{SC_1})\, \alpha_2 I_w$$

$$V_{in} - V_{HP1} = \alpha_2\beta_2\, (R_1 + \frac{1}{SC_1}) \frac{V_z}{R_3}$$

$$V_{in} - V_{HP1} = \alpha_1\, \alpha_2\beta_2\, (R_1 + \frac{1}{SC_1}) \frac{I_x}{SR_3C_2}$$

$$V_{in} - V_{HP1} = \alpha_1\alpha_2\beta_2\, (R_1 + \frac{1}{SC_1}) \frac{V_{HP1}}{SR_3R_2C_2}$$

$$V_{in} = V_{HP1} + \alpha_1\, \alpha_2\beta_2\, (R_1 + \frac{1}{SC_1}) \frac{V_{HP1}}{SR_3R_2C_2}$$

$$\frac{V_{HP1}}{V_{in}} = \frac{S^2}{S^2 + \frac{S\alpha_1\, \alpha_2\beta_2 R_1}{R_3R_2C_2} + \alpha_1\, \alpha_2\beta_2} \qquad (4.8)$$

$$\frac{V_{HP1}}{V_{in}} = \frac{S^2}{D(S)} \qquad (4.9)$$

$$D(S) = S^2 + \frac{S\alpha_1\, \alpha_2\beta_2 R_1}{R_3 R_2 C_2} + \alpha_1\, \alpha_2\beta_2 \qquad (4.10)$$

$$V_{BP1} = \frac{I_Z}{SC_2} = \frac{\alpha_1 I_X}{SC_2}$$

$$V_{BP1} = \alpha_1 \frac{V_{HP1}}{SC_2 R_2}$$

$$V_{BP1} = \alpha_1 \frac{V_{in}}{SC_2 R_2} \frac{S^2}{D(S)}$$

$$\frac{V_{BP1}}{V_{in}} = \alpha_1 \frac{1}{SC_2 R_2} \frac{S^2}{D(S)}$$

$$\frac{V_{BP1}}{V_{in}} = \frac{\alpha_1}{C_2 R_2} \frac{S}{D(S)} \qquad (4.11)$$

The angular resonance frequency and quality factor of the filter are found as:

$$\omega_0 = \frac{1}{\sqrt{C_2 R_1 C_1 R_2}} \qquad (4.12)$$

$$Q = \frac{1}{R_1}\sqrt{\frac{C_2 R_3 R_2}{C_1}} \qquad (4.13)$$

It is observed from above equation that the parameter can be adjusted by changing R_2 and R_3. After adjusting, the parameter of the filter can be controlled via without disturbingω_0. In other words, this filter offers orthogonal control of ω_0 and Q.

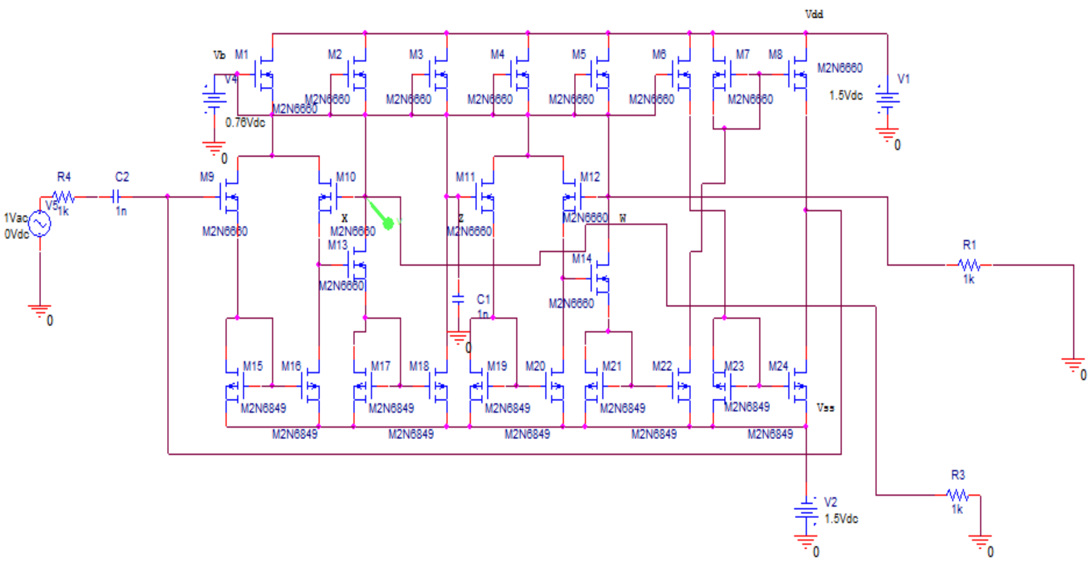


Fig 4.5: Proposed SITO VM filter using ORCAD capture

4.3.2 Proposed TISO Voltage Mode Filter Using Single MCFOA

TISO (Three input single output) voltage mode filter using a single MCFOA is shown in figure 4.6. Analysis of the filters in figure 4.6 gives the following output voltage:

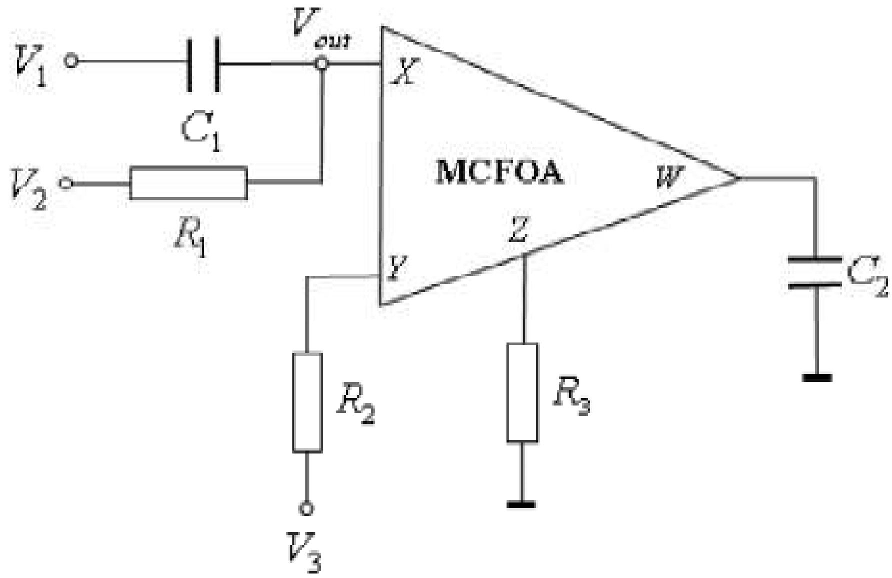


Fig.4.6: Symbolic representation of developed TISO VM filter

$$\mathrm{V_{out}} = \mathrm{V_x} = \mathrm{V_y}$$

$$\mathrm{V_{out}} = I_Y R_2 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -I_w R_2 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -V_w R_2 S C_2 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -V_w R_2 S C_2 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -V_z R_2 S C_2 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -I_z R_2 S C_2 R_3 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -I_X R_2 S C_2 R_3 + \mathrm{V_3}$$

$$\mathrm{V_{out}} = -R_2 S C_2 R_3 \left[(\mathrm{V_{out}} - \mathrm{V_1}) S C_1 + \frac{(\mathrm{V_{out}} - \mathrm{V_2})}{R_1}\right] + \mathrm{V_3}$$

$$\mathrm{V_{out}}\left(1 + S^2 C_1 C_2 R_2 R_3 + \frac{R_2 S C_2 R_3}{R_1}\right) = \mathrm{V_1} S^2 C_1 C_2 R_2 R_3 + \frac{R_2 S C_2 R_3}{R_1}\mathrm{V_2} + \mathrm{V_3}$$

$$\mathrm{V_{out}} = \frac{\mathrm{V_1} S^2 C_1 C_2 R_2 R_3 + \frac{R_2 S C_2 R_3}{R_1}\mathrm{V_2} + \mathrm{V_3}}{\left(1 + S^2 C_1 C_2 R_2 R_3 + \frac{R_2 S C_2 R_3}{R_1}\right)} \qquad (4.14)$$

Depending on the applied input voltages V_i, i = 1, 2, 3, non-inverting unity gain different filter responses can be easily obtained as:

1) Highpass response for $V_1 = V_{in}$ and $V_2 = V_3 = 0$.
2) Bandpass response for $V_2 = V_{in}$ and $V_1 = V_3 = 0$.
3) Lowpass response for $V_3 = V_{in}$ and $V_1 = V_2 = 0$.
4) All-pass response is realized for $V_1 = V_3 = -V_2 = V_{in}$.

The parameters ω_0 and Q of the proposed filters in figure 4.5 is given as

$$\omega_0 = \frac{1}{\sqrt{C_2 R_3 C_1 R_2}} \qquad (4.15)$$

$$Q = R_1 \sqrt{\frac{C_1}{C_2 R_3 R_2 \, C_1}} \qquad (4.16)$$

It is observed from (4.15) and (4.16) that the parameter Q can easily be tuned through R_1without disturbing ω_0 .

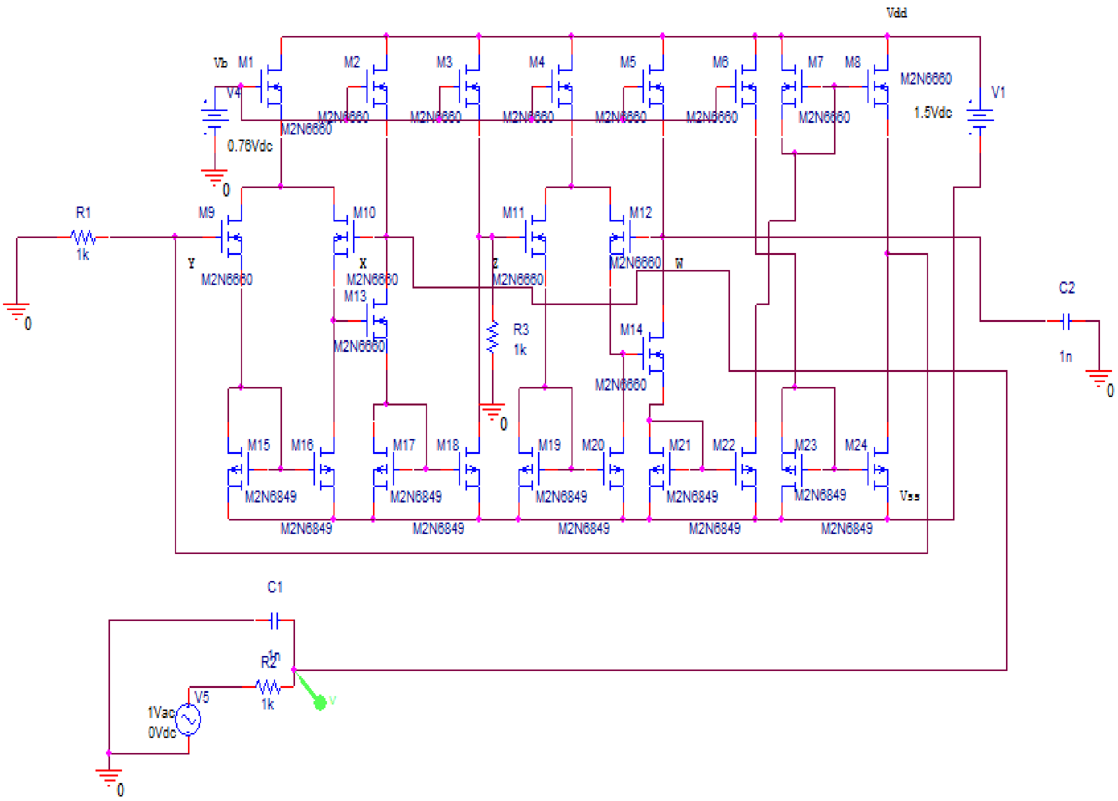


Fig. 4.7: A low pass filter realisation from the developed TISO VM filter

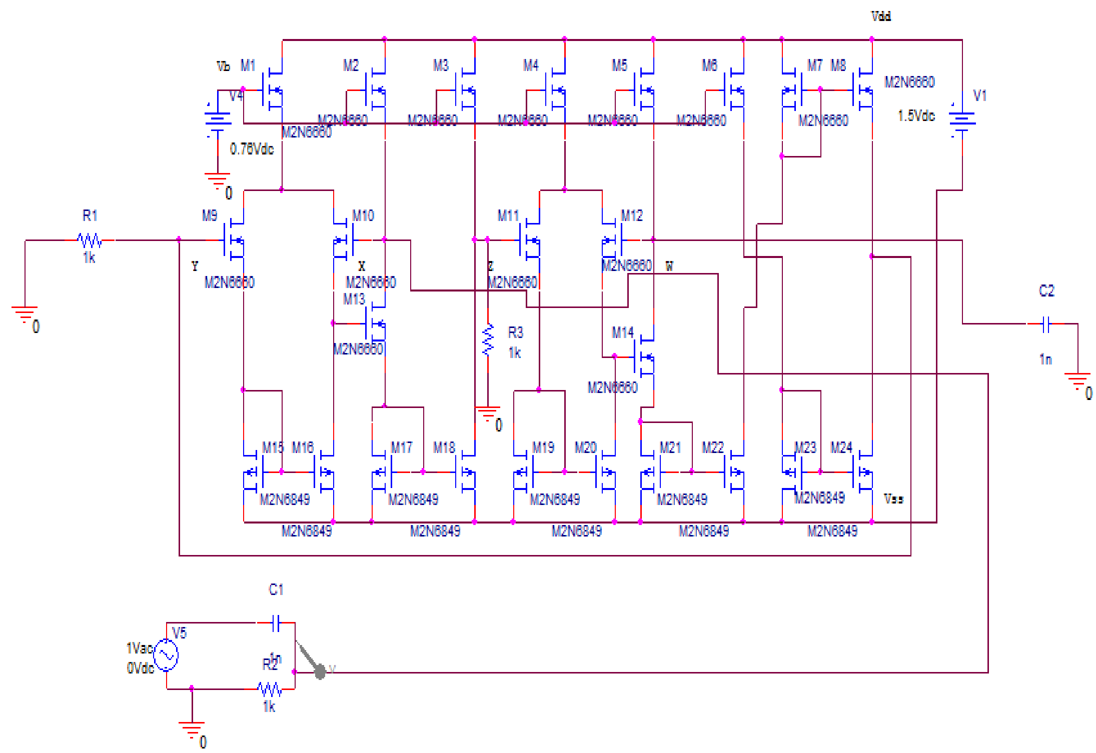


Fig. 4.8: A high pass filter using developed TISO VM filter.

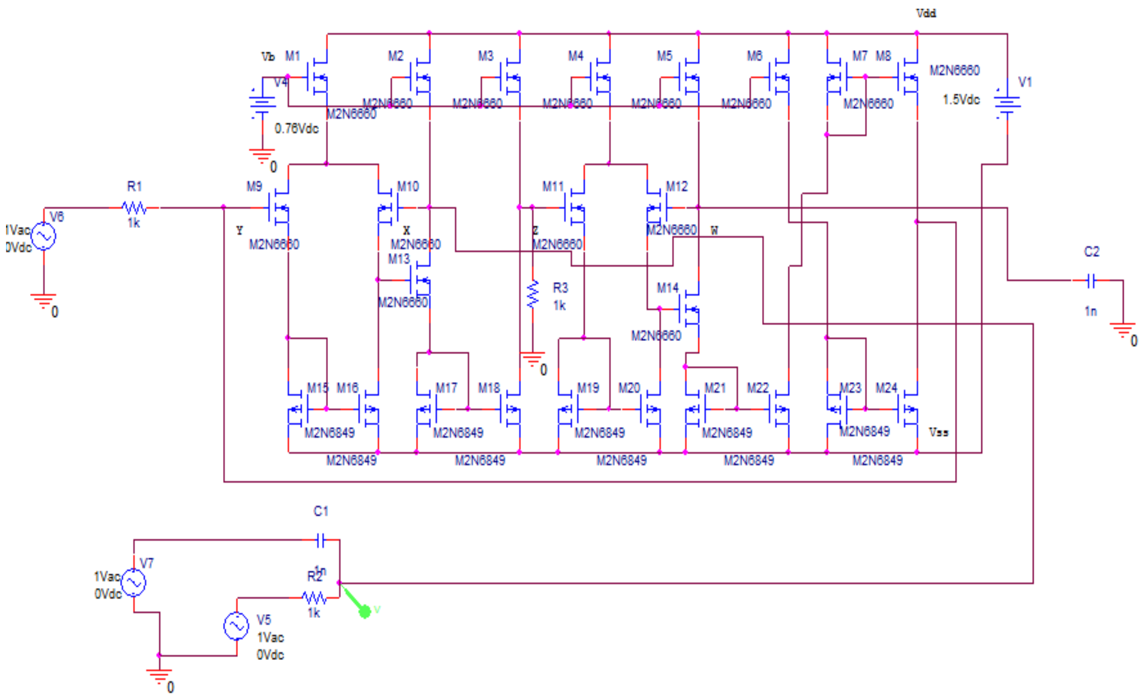


Fig. 4.9: An all pass filter using developed TISO VM filter.

CHAPTER 5

THE PROPOSED CIRCUITS FOR INVERSE FILTER

Various types of inverse filters have been devised, using CFOA based integrators and summers, in a variety of ways which exhibit a number of interesting features in respect of the total number of active and passive components involved, tunability properties of the resulting circuits and their input and output impedances.

The proposed new realisations of inverse low pass (ILP) filters with infinite input impedance are shown in figure 5.1. The proposed circuit for the realisation of inverse high pass (IHP) filter is shown in Figure 5.3, which also shows the transfer function realised by the circuit. It may be noted that the input impedance of the presented circuit is infinite.

The proposed new realisations of inverse bandpass (IBP) filters with infinite input impedance are shown in figure 5.2.

5.1 PROPOSED INVERSE LOWPASS FILTER

Considering CFOA to be characterized by $i_y = 0$, $v_x = v_y$, $i_z = i_x$, and $v_w = v_z$, the transfer function realized with these circuits are given by:

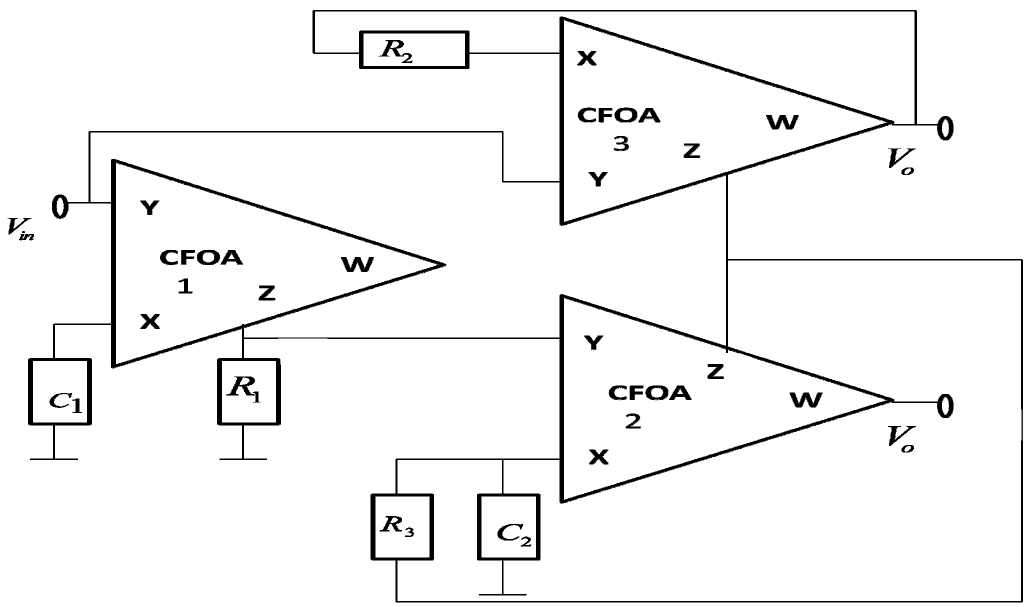


Fig. 5.1: New implementations of ILP filter

$$V_0 = V_{W_2} = V_{Z_3} = I_{Z_3}R_3 + I_{Z_2}R_3 + V_{Z_1}$$

$$V_0 = I_{X_3}R_3 + I_{X_2}R_3 + I_{Z_1}R_1$$

$$V_0 = V_{in}SC_1R_1 + I_{X_3}R_3 + \frac{R_3}{2}\left(V_{in}SC_1R_1SC_2 - I_{X_3}\right)$$

$$V_0 = V_{in}SC_1R_1\left[1 + \frac{R_3SC_2}{2}\right] + \frac{R_3}{2}\left(I_{X_3}\right)$$

$$V_0 = V_{in}SC_1R_1\left[1 + \frac{R_3SC_2}{2}\right] + \frac{R_3}{2R_2}\left(V_{in} - V_0\right)$$

$$V_0\left[1 + \frac{R_3}{2R_2}\right] = V_{in}\left(SC_1R_1 + \frac{R_3}{2R_2} + \frac{R_3}{2}S^2C_2SC_1R_1\right)$$

$$V_0\left[1 + \frac{R_3}{2R_2}\right] = V_{in}C_2C_1R_1\frac{R_3}{2}\left(\frac{2S}{C_2R_3} + \frac{1}{C_2C_1R_1R_2} + S^2\right)$$

$$\frac{V_0}{V_{in}} = \frac{\frac{1}{\left[1 + \frac{2R_2}{R_3}\right]\frac{1}{C_2C_1R_1R_2}}}{\left(\frac{2S}{C_2R_3} + \frac{1}{C_2C_1R_1R_2} + S^2\right)} \qquad (5.1)$$

5.2 PROPOSED INVERSE BANDPASS FILTER

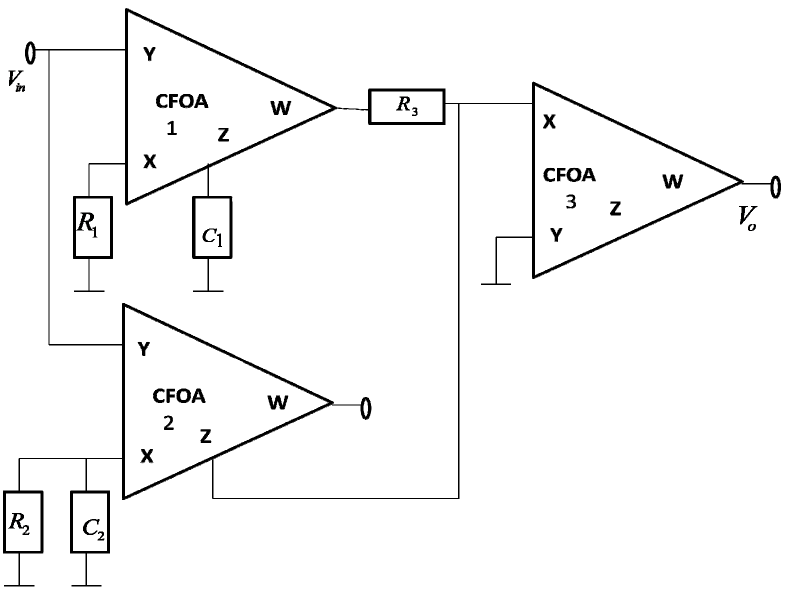


Fig. 5.2 New implementations of IBP filter

$$V_0 = V_{W_3} = V_{Z_3} = I_{Z_3}R_0 = I_{X_3}R_0$$

$$V_0 = R_0\left(\frac{V_{W_1}}{R_3} + I_{Z_2}\right)$$

$$V_0 = R_0\left(\frac{V_{W_1}}{R_3} + I_{X_2}\right)$$

$$V_0 = R_0\left(\frac{V_{in}}{R_2} + V_{in}SC_2\right) + \frac{R_0 I_{Z_1}}{R_3 SC_1}$$

$$V_0 = R_0\left(\frac{V_{in}}{R_2} + V_{in}SC_2\right) + \frac{R_0 I_{X_1}}{R_3 SC_1}$$

$$V_0 = R_0 V_{in}\left(\frac{1}{R_2} + SC_2\right) + \frac{R_0}{R_3 SC_1}\frac{V_{in}}{R_1}$$

$$\frac{V_0}{V_{in}} = R_0\left(\frac{1}{R_2} + SC_2 + \frac{1}{SC_1 R_1 R_3}\right)$$

$$\frac{V_0}{V_{in}} = \frac{R_0 C_2 R_2}{SR_2}\left(S^2 + \frac{S}{C_2 R_2} + \frac{1}{C_1 R_1 C_2 R_3}\right)$$

$$\frac{V_O}{V_{in}} = \frac{\frac{1}{\left(\frac{R_2}{R_0}\right)\frac{S}{C_2 R_2}}}{S^2 + \frac{S}{C_2 R_2} + \frac{1}{C_1 C_2 R_1 R_3}} \tag{5.2}$$

5.3 PROPOSED INVERSE HIGH PASS FILTER

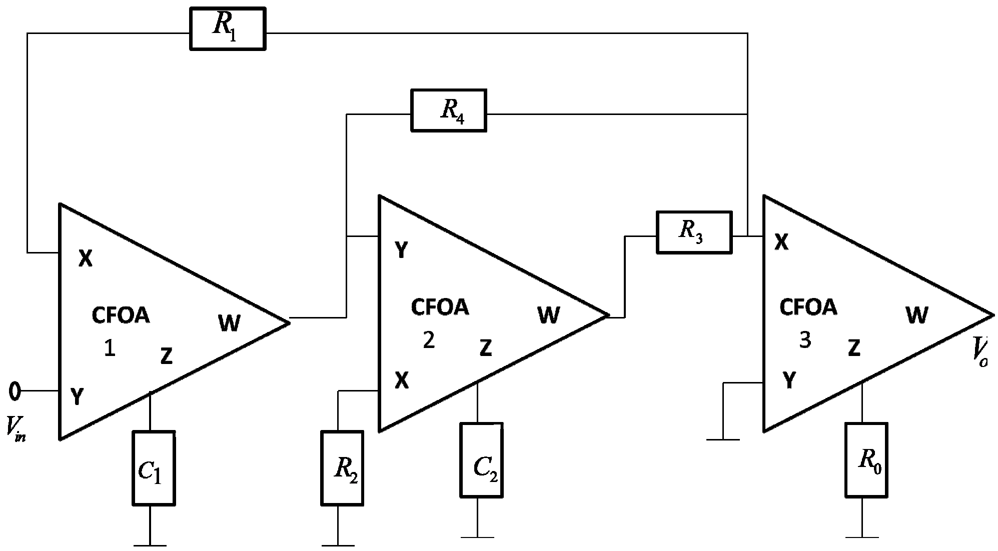


Fig. 5.3: New implementation of IHP filter

$$V_0 = V_{W_3} = V_{Z_3} = I_{Z_3}R_0 = I_{X_3}R_0$$

$$V_0 = \left(\frac{V_{W_2}}{R_3} + \frac{V_{W_1}}{R_4} + \frac{V_{in}}{R_1}\right)R_0$$

$$V_0 = \left(\frac{V_{z_2}}{R_3} + \frac{V_{z_1}}{R_4} + \frac{V_{in}}{R_1}\right)R_0$$

$$V_0 = \left(\frac{I_{z_2}}{SC_2R_3} + \frac{I_{z_1}}{SC_1R_4} + \frac{V_{in}}{R_1}\right)R_0$$

$$V_0 = \left(\frac{I_{X_2}}{SC_2R_3} + \frac{I_{X_1}}{SC_1R_4} + \frac{V_{in}}{R_1}\right)R_0$$

$$V_0 = \left(\frac{V_{z_1}}{R_2SC_2R_3} + \frac{V_{in}}{R_1SC_1R_4} + \frac{V_{in}}{R_1}\right)R_0$$

$$V_0 = \left(\frac{V_{in}}{S^2C_2R_1R_2C_1R_3} + \frac{V_{in}}{R_1SC_1R_4} + \frac{V_{in}}{R_1}\right)R_0$$

$$V_0 = \left(\frac{1}{S^2C_2R_2C_1R_3} + \frac{1}{SC_1R_4} + 1\right)\frac{V_{in}}{R_1}R_0$$

$$V_0 = \left(\frac{1}{S^2C_2R_2C_1R_3} + \frac{1}{SC_1R_4} + 1\right)\frac{V_{in}}{R_1}R_0$$

$$\frac{V_0}{V_{in}} = \left(\frac{1}{C_2R_2C_1R_3} + \frac{S}{C_1R_4} + S^2\right)\frac{1}{S^2R_1}R_0$$

$$\frac{V_O}{V_{in}} = \frac{\frac{1}{\left(\frac{R_1}{R_0}\right)S^2}}{S^2 + \frac{S}{C_1R_4} + \frac{1}{C_1C_2R_2R_3}} \tag{5.3}$$

From equation (5.1)-(5.3) we can realize inverse Lowpass, inverse Bandpass and inverse Highpass filters. Since a number of new circuits have been proposed for realising standard inverse Lowpass, Bandpass and Highpass responses it is useful to compare the various structures in each class on the basis of some features of practical interest. In this context, the following may be observed:

(1) Input and output impedances:

It is desirable that a Voltage Mode filter should have ideally infinite input impedance and ideally zero output impedance. The latter is available in all the circuits as the output is taken from the w terminal of the CFOA. All the circuits offer ideally infinite input impedance and hence, are superior to those described earlier.

(2) Controllability of coefficients of transfer functions:

Various circuits offer different types of control of coefficients of transfer functions realised.

- From the given transfer functions it can be readily inferred that Inverse Lowpass and Inverse Bandpass can provide control of coefficients in a sequential way through various resistors.
- Inverse Lowpass can provide independent control of bandwidth and cut off frequency. In case of Inverse Highpass it is interesting to note that all the three coefficients can be adjusted with complete decoupling of resistors with no resistor being common in the coefficients.

(3) Total component count:

Since all the circuits employ three CFOAs and two grounded capacitors, total component count is governed by the number of resistors employed. It is seen that the total number of resistors is as small as three as in inverse Lowpass and as large as five as in inverse Highpass.

(4) Suitability for IC implementation:

Since all the proposed structures employ both Grounded Capacitors, they are suitable for IC implementation.

(5) PID Controller

It must be noted that in an Inverse Bandpass filter, the output V_0 of the circuit is related to the input Vin by the following equation in the time domain:

$$V_0(t) = -\left[\frac{R_0}{R_4}V_{in} + \frac{R_0}{R_1R_5C_1}\int V_{in}\,dt + \frac{R_2R_0C_2}{R_3}\frac{dV_{in}}{dt}\right]$$

And, hence the inverse Bandpass filter proposed here can be used as a proportional integral differential (PID) controller. Conversely, a PID controller such as that in Yuce, Tokat, Minaei and Cicekoglu (2006) is in fact also an Inverse Bandpass filter.

(6) Consideration of CFOA parasitic:

The prominent non idealities of the CFOAs include a finite non-zero input resistance r_x at port x (typically around 50 Ω), the z port parasitic impedance consisting of a parasitic resistance Rp (typically 3 MΩ) in parallel with a parasitic capacitance Cp (typically 5 pF) and the y port parasitic impedance consisting of a parasitic resistance Ry (typically 2 MΩ) in parallel with a parasitic capacitance Cy (typically, 2 pF).

The errors caused by the influence of CFOA can be kept small by choosing all external resistors to be much larger than r_x but much smaller than Rp and Ry and choosing external capacitors to be much larger than Cp and Cy.

It may be noted that in circuits Inverse Lowpass filter in Figure 5.1, there is a capacitor from the x-terminal of the CFOA to ground. It may appear that this capacitor in association with the parasitic x-resistance will form an additional pole which may not only disturb the function of the inverse filter but may also lead to instability.

The study of the effect of this parasitic pole for one of the present circuits (Inverse Lowpass filter of Figure 5.1) shows the non-ideal transfer function from which it is found that the parasitic poles are in the left half of the s plane and the non-ideal frequency response is in good correspondence with the practical results. Furthermore, the experimental results of Inverse Lowpass filter in Figure 5.1 have confirmed that in spite of this capacitor being at x-terminal, none of the mentioned circuits have exhibited any problem of instability.

5.4 A COMPARATIVE STUDY OF VARIOUS INVERSE FILTER CIRCUITS

As already pointed out, although several earlier works have dealt with the general procedures for dealing with inverse transfer functions or specific inverse filter structures, only in Abuelmaatti (2000) the various circuits for realising four standard inverse functions (namely, Lowpass, Bandpass, Highpass and Band reject) using a single FTFN (Four terminal floating nullor) have been presented. Therefore, a comparison of the proposed configurations with those presented in Abuelmaatti (2000) reveals the following:

- The inverse Lowpass circuit of Abuelmaatti (2000) provides Current mode operation, employs two grounded capacitors/one floating capacitor and has non-ideal (non-zero) input impedance whereas by contrast the proposed Inverse Lowpass of Figure 5.1 provide Voltage mode operation, employ two grounded capacitors and offer ideal (infinite) input impedance.

- The Inverse Highpass circuit of Abuelmaatti (2000) provides current mode operation, employs three/two capacitors with one of them floating and has non-ideal (non-zero) input impedance whereas by contrast the proposed Inverse Highpass of Figure 5.3 provides voltage mode operation, two grounded capacitors and has ideal (infinite) input impedance.
- The inverse band pass circuit of Abuelmaatti (2000) provides current mode operation, employs three capacitors and has non-ideal (non-zero) input impedance whereas by contrast the proposed Inverse Bandpass of Figure 5.2 provide voltage mode operation, employ two grounded capacitors and offer ideal (infinite) input impedance.
- The inverse bandreject circuit of Abuelmaatti (2000) provides current mode operation, employs two capacitors with one or both floating, has ideal (zero) input impedance and the condition needs to be fulfilled.

It may be seen that the circuits of Abuelmaatti (2000) have the advantage of employing only one FTFN; however, they realise various inverse filters in current mode . By contrast, the circuits presented here require three CFOAs (one more than the circuits of Abuelmaatti (2000) if FTFN is assumed to be realised by two CFOAs) and realise various inverse functions in voltage mode and offer several advantages.

When compared with the earlier proposed circuits of Gupta et al. (2009), it is observed that the Inverse Bandpass filters of Figure 5.2 have properties similar to the circuit of Inverse Bandpass filter given in of Gupta et al. (2009). However on the other hand, the new Inverse Lowpass proposed in Figure here have the advantage of employing one resistor less and still provide sequential/independent control of filter parameters; and lastly, the new inverse highpass proposed here in Figure 5.3 is capable of providing independent controllability of all the three parameters of the filter.

It is thus, observed that various circuits proposed here have different kinds of advantages/features over those of Leuciuc (1997), Chipipop and Surakampontorn (1999), Wang and Lee (1999), Abuelmaatti (2000) and Gupta et al. (2009).

CHAPTER 6

SIMULATION RESULTS

Simulations for the MCFOA of figure 3.5 based on 0.25 um TSMC CMOS technology with power supply voltages $V_{DD} = -V_{SS} = 1.25V$ and $V_B = .8V$ are performed. The dimensions of the MOS transistors used in the proposed MCFOA are given in Table I and the voltage gain versus frequency and current gain versus frequency are shown in figure 3.6 and 3.7 respectively.

6.1 SIMULATION RESULT OF INDUCTOR

The floating inductor given in figure 4.2 is simulated using Modified CFOA. The MCFOA is realized using CMOS TSMC structure as shown figure 3.5.Then to confirm the workability of floating inductor an RL series circuit has been designed which will give high pass filter response when we take the output across the inductor. The simulation result for the Highpass filter using developed floating inductor is shown in figure 6.1 The passive elements are selected as $R_1 = R_2 = 1K\Omega$ and $C = 1nF$. Then we have measured the 3 db cut off frequency of Highpass filter and from that we have obtained the valve of the floating inductor.

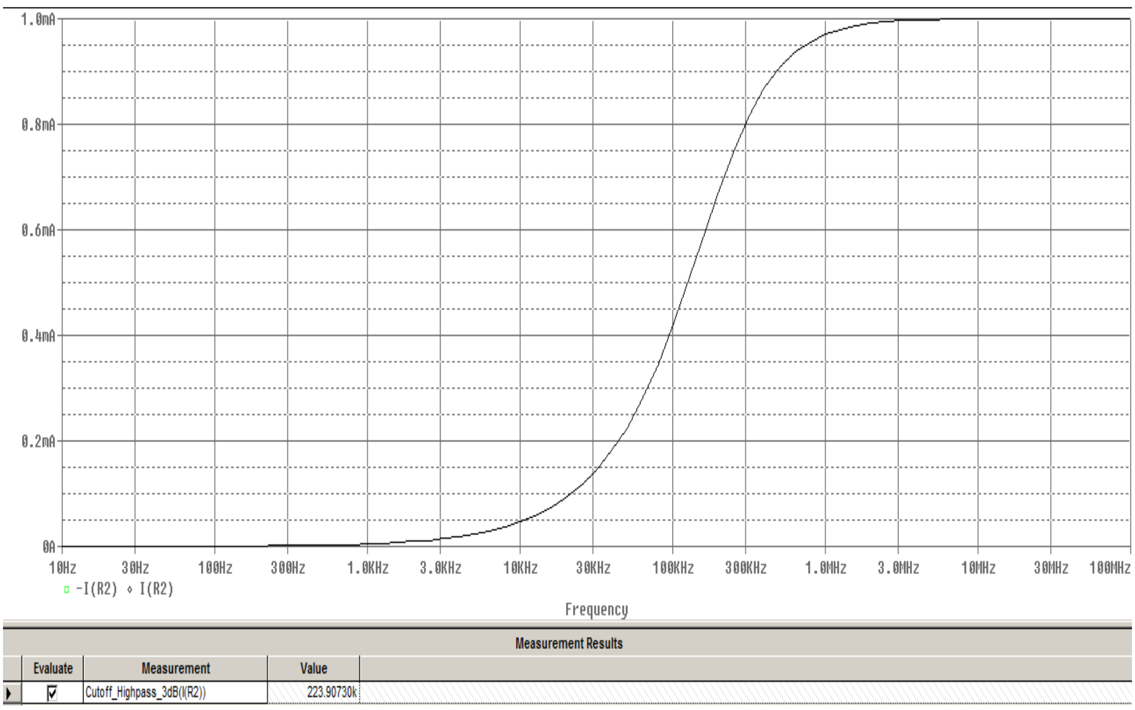


Fig.6.1: Output of the Highpass filter constructed using simulated floating inductor

From the equation (4.7) we can write

$$L_{eq} = CR_1R_2$$

Thus on selecting $R_1 = R_2 = 1$ kΩ and C = 1 nF, we obtain the value of simulated inductor = 1000 μH.

From the simulation result of floating inductor we obtained the cut off frequency of low pass filter = 223.907 KHz

3 db Cut off frequency (f_0) of high pass RL filter is given by

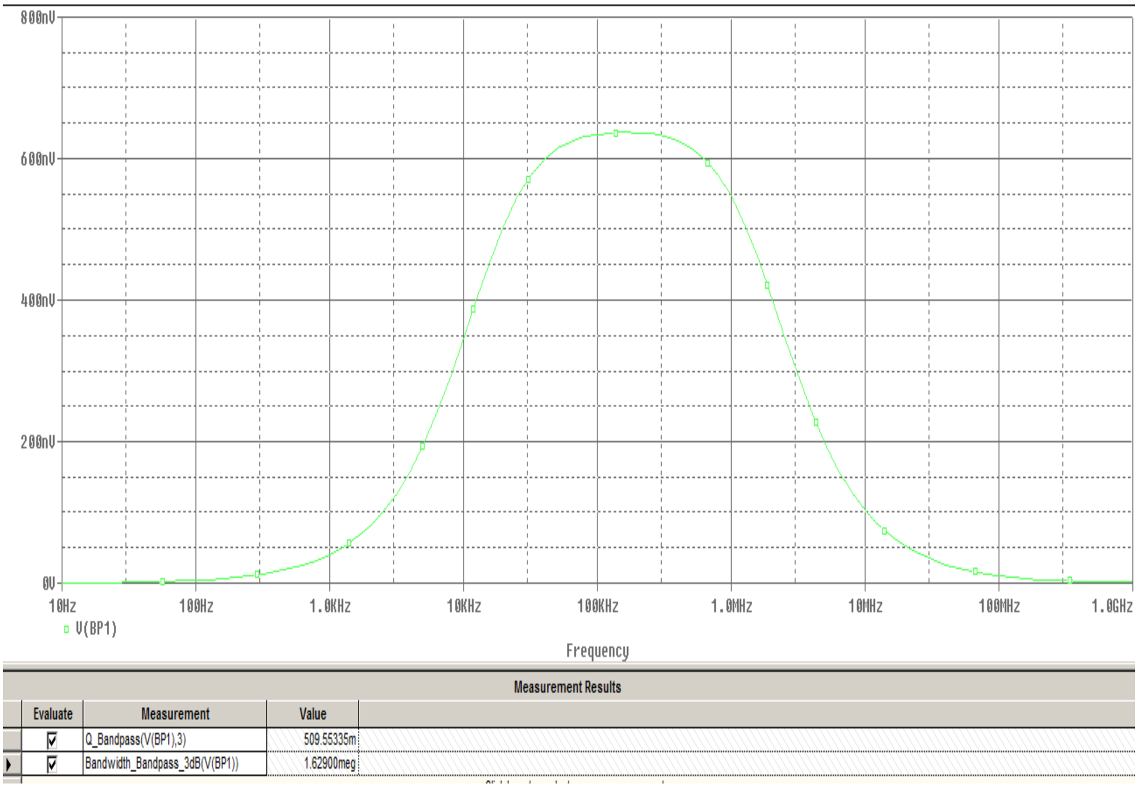
$$f_0 = \frac{R}{2\pi L}$$

Thus on selecting the value of R = 1 k Ω, we get the value of simulated inductor = 811 μH.

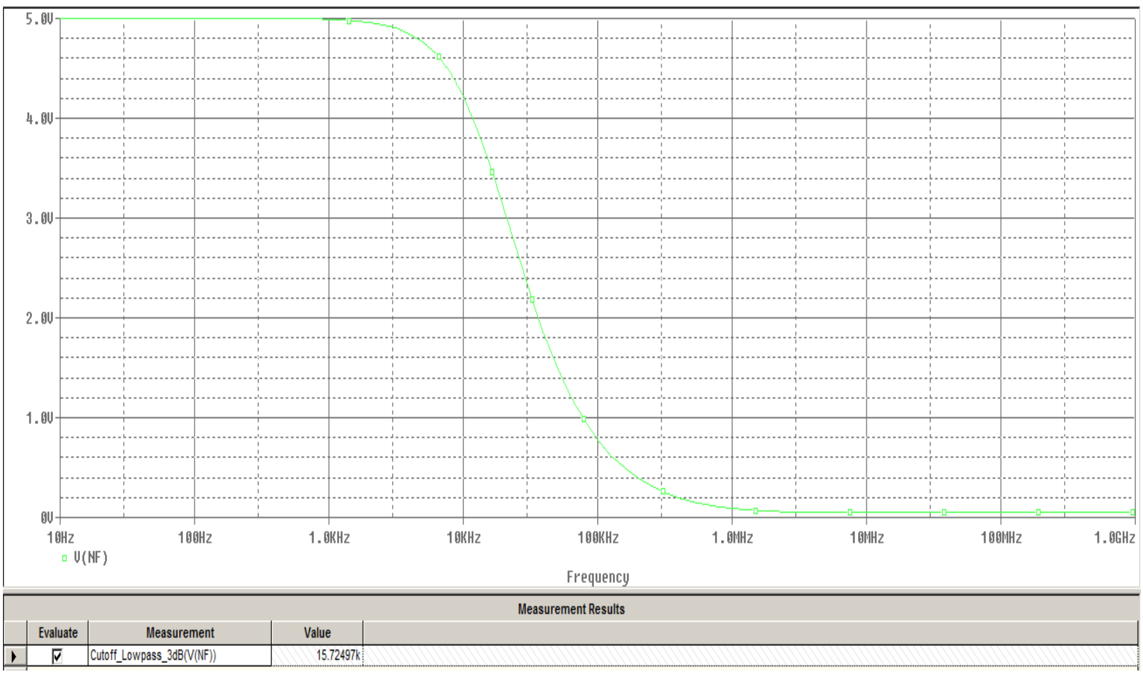
6.2 SIMULATION RESULTS OF SITO FILTER

The SITO (single input three outputs) Voltage mode filters given in figure 4.4 are simulated using Modified CFOA to confirm the workability of the filters. The MCFOA is realized using CMOS TSMC structure as shown Figure 3.5. The simulation results for the filters in figure 4.4 are obtained using the MCFOA in Fig. 3.5. The passive elements are selected as $R_1 = R_2 = R_3 = 10$KΩ and $C_1 = C_2 = 1$nF.

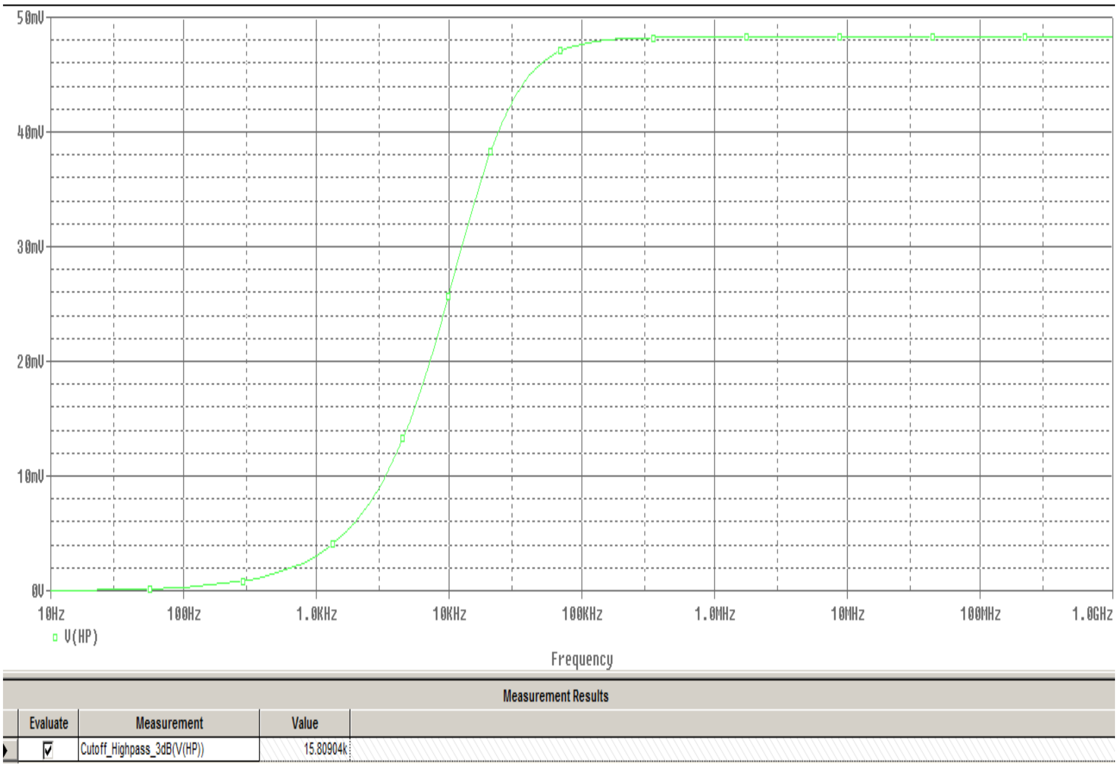
All the simulated filter responses of the circuits in figure 4.4 are given in figure. 6.2. It is important to note that the discrepancy among ideal, simulation responses mainly arises from the parasitic impedance and nonideal gain effects of the active devices and tolerances of the passive elements.



(a) Bandpass filter



(b) Low pass filter



(c) High pass filter

Fig.6.2: Outputs of the Proposed SITO VM filter

Thus on selecting $R_1 = R_2 = R_3 = 10K\Omega$ and $C_1 = C_2 = 1nF$ in equation 4.12, we would obtain 3 db cut off frequency of Highpass filter from the equation 4.12

$$f_0 = 15.92 \text{ KHz}$$

And from the equation 4.13, we obtained Q = 1

From the simulation result, we obtained the 3 db cut off frequency of high pass filter

$$f_0 = 15.72 \text{ KHz}$$

Band width of the Bandpass filter is 1.62 MHz.

6.3 SIMULATION RESULTS OF TISO FILTER

The TISO (three input single output) Voltage mode filters given in figure 4.6 are simulated using Modified CFOA to confirm the workability of the filters. The MCFOA is realized using CMOS TSMC structure as shown Figure 3.5. The simulation results for the filters in figure 4.6 are obtained using the MCFOA in Fig. 3.5. The passive elements are selected as $R_1 = R_2 = R_3 = 10K\Omega$ and $C_1 = C_2 = 1nF$. All the simulated Lowpass, highpasss and all pass filter responses of the circuits in figure 4.6 are given in figure 6.3, 6.4 and 6.5 respectively.

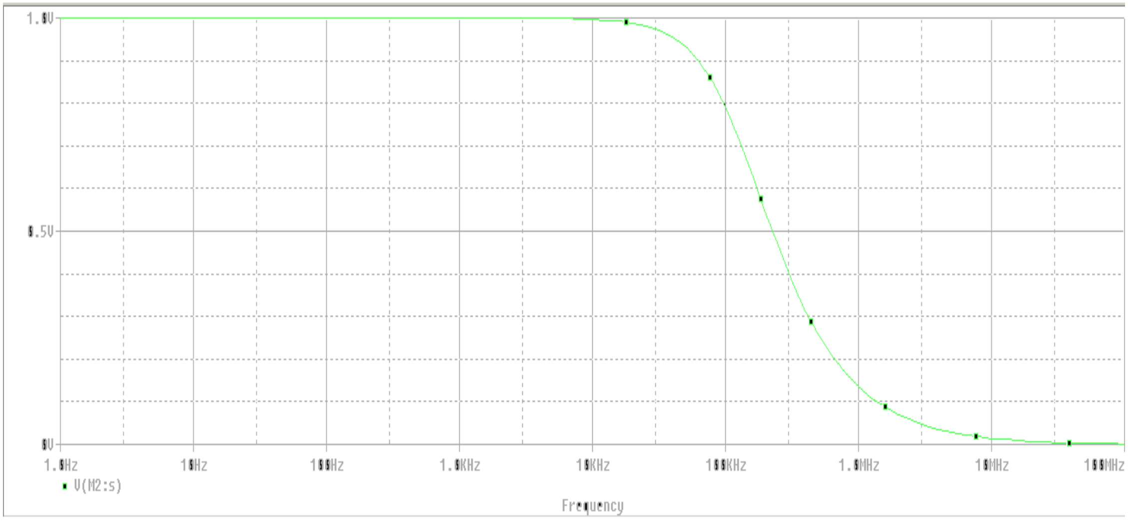


Fig.6.3: Output of Lowpass filter developed TISO VM filter

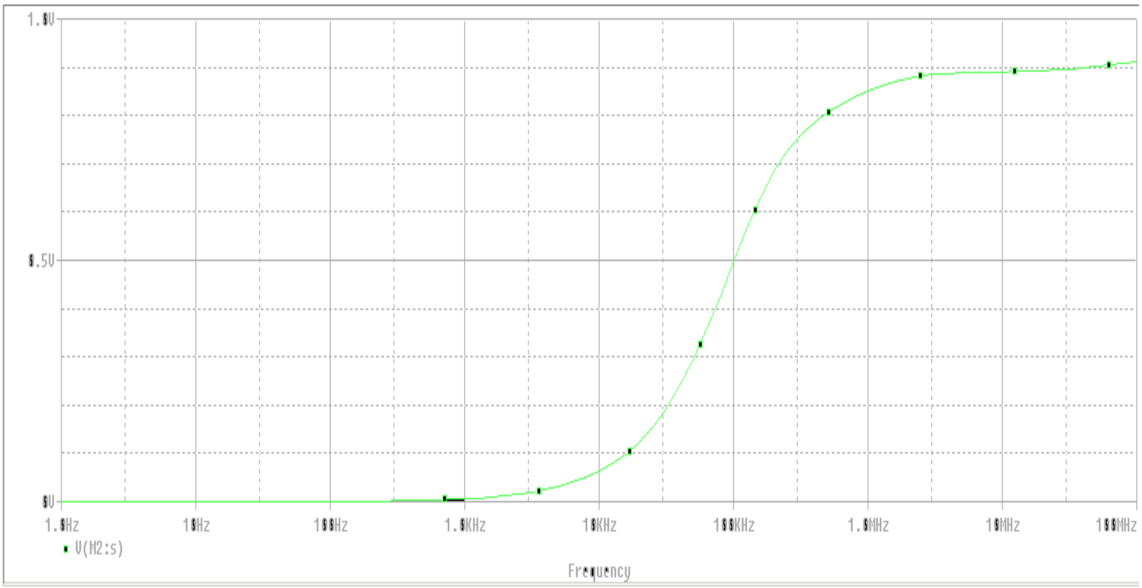


Fig. 6.4: Output of a high pass filter using developed TISO VM filter.

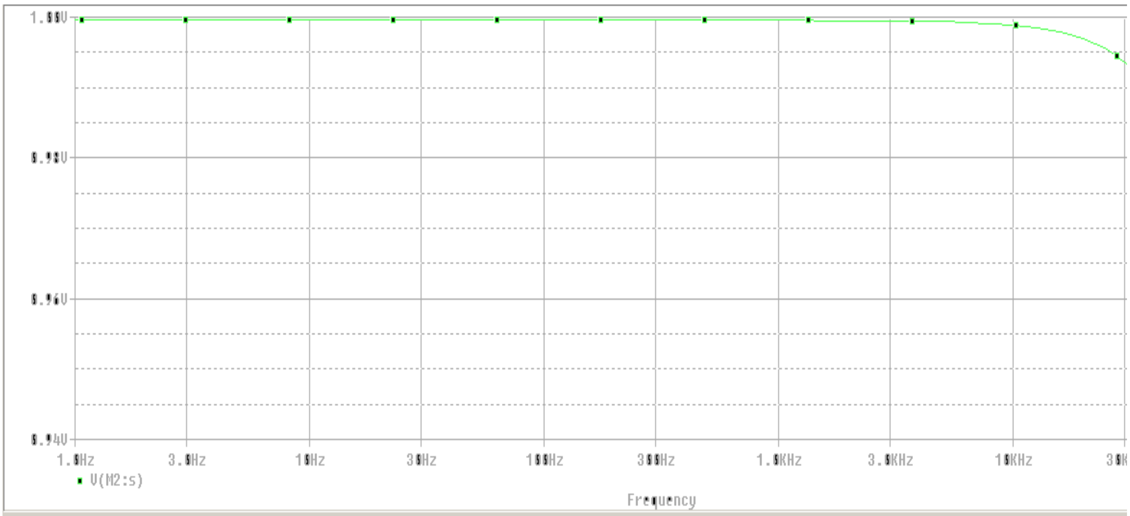


Fig. 6.5: Output of all pass filter using developed TISO

6.4 SIMULATION RESULTS OF INVERSE FILTERS

Inverse filters have been designed using both the Modified Current Feedback Operational Amplifier (MCFOA) and standard AD844 CFOA. In the simulation, the values of all resistors and all capacitors are 40 kΩ and 1 nF, respectively. The commercial current feedback amplifiers AD844 macro model with ± 12 V voltage supply is used to realize the CFOA in figure 3.1**.**

It is found that the workability of all the inverse biquids implemented using modified CFOA is in good agreement with inverse filter designed using commercial AD844 CFOA. The typical frequency responses of inverse lowpass, inverse bandpass and inverse highpass using both commercial AD844 CFOA and Modified CFOA are shown in the following figures. The deviation of frequency response is due to the parasitic impedance of nonideal standard AD844 CFOA and difference in the parameters with the Modified CFOA. On comparison from the figures, we can conclude that the cut off frequency of the inverse filters designed using modified CFOA has increased than the inverse filter designed using AD844.

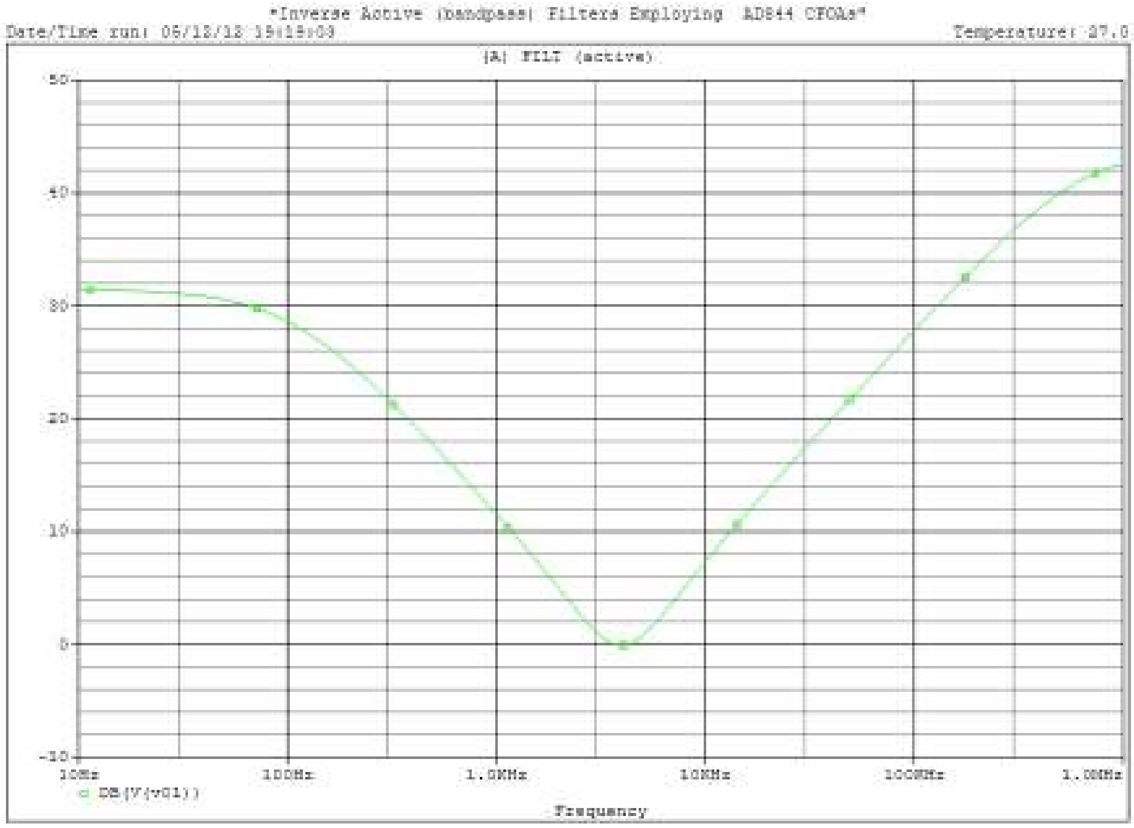


Fig.6.6: Frequency response of inverse band pass filter using AD844 CFOA

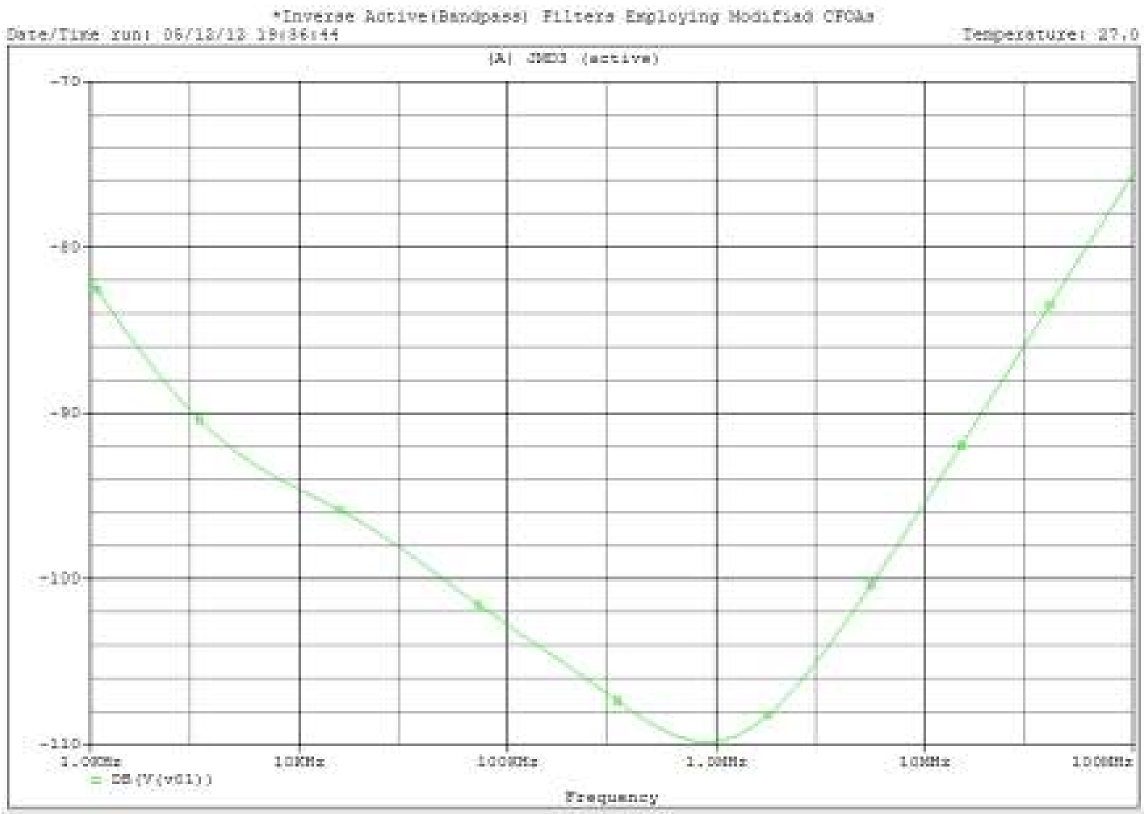


Fig. 6.7: Frequency response of inverse band pass filter using Modified CFOA.

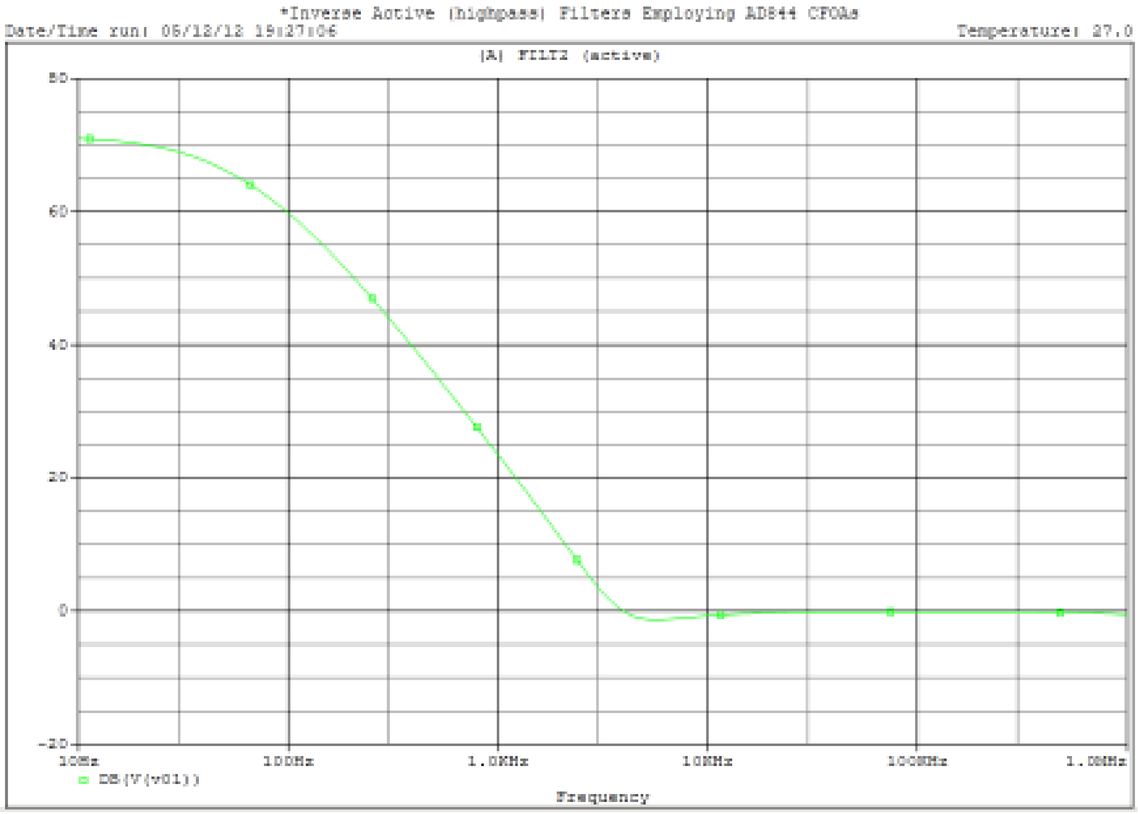


Fig.6.8: Frequency response of inverse high pass filter using AD844 CFOA

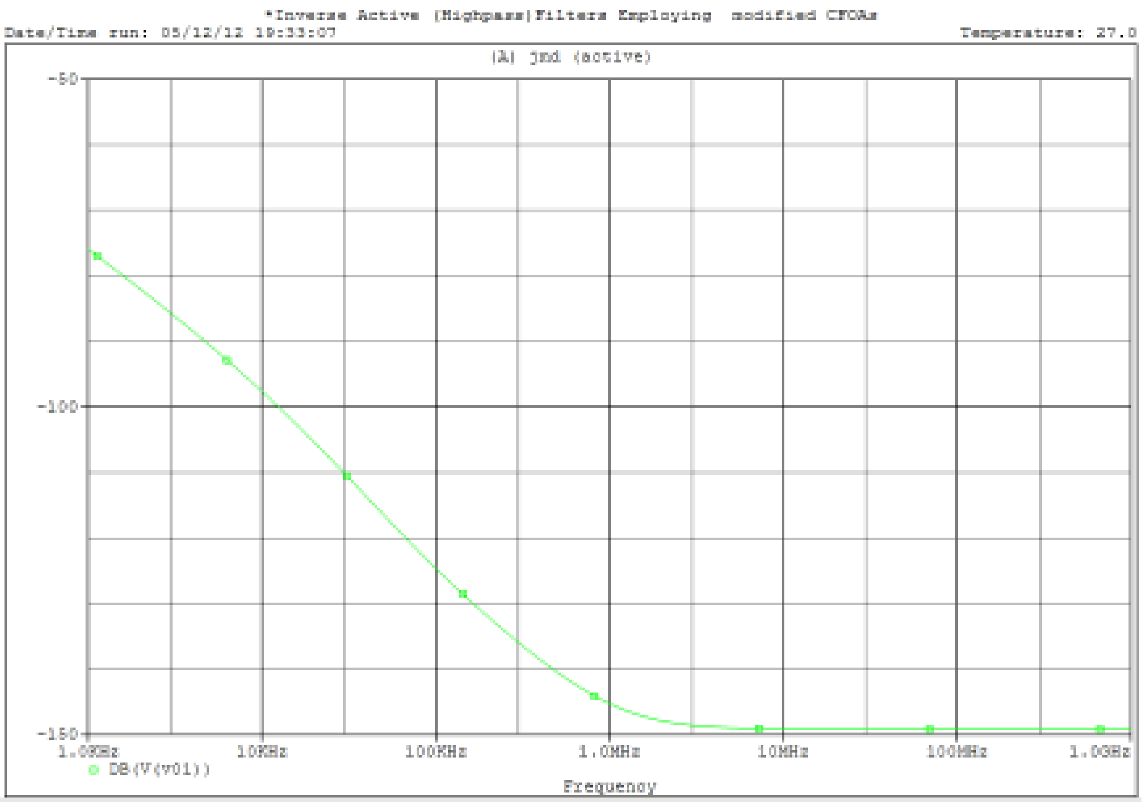


Fig.6.9: Frequency response of inverse high pass filter using Modified CFOA

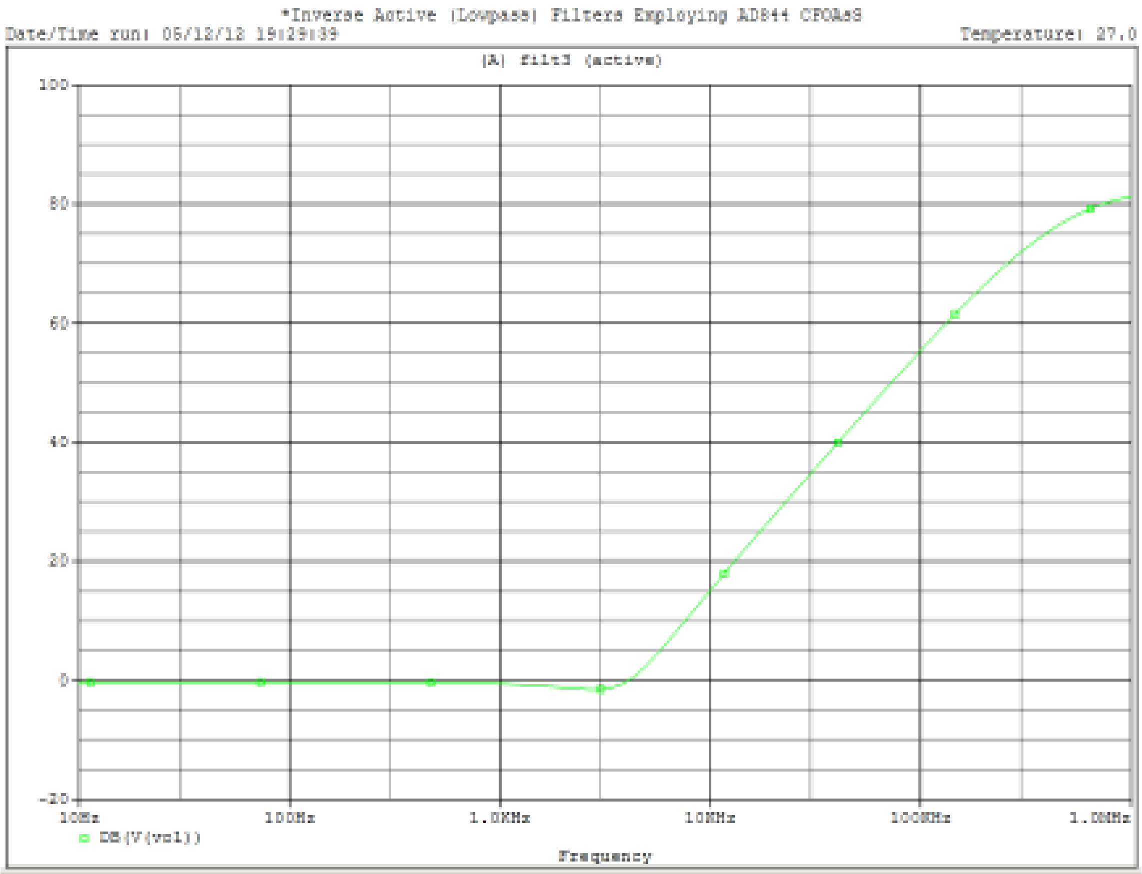


Fig.6.10: Frequency response of inverse low pass filter using AD844 CFOA

Fig.6.11: Frequency response of inverse low pass filter using Modified CFOA

As we can measure from the given circuits for inverse Lowpass filter that in the cut off frequency of inverse Lowpass filter designed with CFOA and MCFOA are respectively 554.71 kHz and 531.025 M Hz and it has been increased 1000 times. In case of inverse Highpass filter the cut of frequencies are 1.405 KHz and 37.07 MHz respectively, with CFOA and MCFOA.

CHAPTER 7

CONCLUSION AND FUTURE WORK

6.1 CONCLUSIONS

The main contributions of our research are summarized as follows:

1. First, we proposed a versatile CMOS active device, namely, a modified current feedback operational amplifier (MCFOA), and then we analyzed the different ways for simulating an inductor to remove the disadvantage of using an inductor in analog filter synthesis and IC fabrication

2. Then a modified CFOA (MCFOA) based grounded inductor, floating inductor, single input three outputs (SITO) filter, three input single output (TISO) filters and inverse filters has been realized. All of the presented circuits have the advantage that they don't need component matching conditions and cancellation constraints so it is easy to implement them in IC design. Proposed inductors possess the feature of reduction in size, weight and area. It can also be used at low frequency with better quality factor. Simulation results of these circuits are in close agreement with theoretical values.

3. Then we discussed the need of inverse filter in communication and control and after that we have presented three inverse filter configurations (inverse lowpass, inverse highpass and inverse Bandpass) based on the modified CFOA and AD844 CFOA.

4. The proposed circuits, thus, add new configurations to the existing repertoire of inverse analog filters known earlier and should be useful in offering new alternatives to the analog designers.

6.2 FUTURE WORK

We plan to carry our research for all the levels of analog filters and to propose some more methods to improve the effectiveness and performance of analog filters using different simulated inductor.

AC analysis of modified CFOA based filters and inverse filters is presented in this thesis, but other analysis like time domain analysis, noise analysis, effect of parameter variation, effect of power supply variation, dc analysis etc. are yet to be explored to make these circuits a market product.

In this work, inductor, filters and inverse filters based on the modified CFOA has been realized. This realization can be carried out for other circuit configurations like oscillators, amplifiers, integrators and differentiators for improving the circuit performance. These MCFOA based designs are best suited for the Low Power and High Speed applications so these can be used in MEMS (Micro-electromechanical systems) and Low Power Medical Devices.

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APPENDIX

Simulations have been performed using SPICE at 0.25 um TSMC CMOS technology and the model files provided by MOSIS (AGILENT) used have the following parameters

.MODEL NMOS NMOS (LEVEL = 3

+ TOX = 5.7E-9 NSUB = 1E17 GAMMA = 0.4317311
+ PHI = 0.7 VTO = 0.4238252 DELTA = 0
+ UO = 425.6466519 ETA = 0 THETA = 0.1754054

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

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

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

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

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

+ CJSW = 5.341337E-10 MJSW = 0.5)

.MODEL PMOS PMOS (LEVEL = 3

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

+ PHI = 0.7 VTO = -0.5536085 DELTA = 0

+ UO = 250 ETA = 0 THETA = 0.1573195

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

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

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

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

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

+ CJSW = 3.625544E-10 MJSW = 0.5)

*** AD844 SPICE Macro-model**

.SUBCKT AD844 1 2 99 50 28 12

*

* INPUT STAGE

*

```
R1   99  8     1E3
R2   10 50     1E3
V1   99  9     11
D1   9   8     DX
V2   11 50     11
D2   10 11     DX
I1   99  5     258E-6
I2   4  50     258E-6
Q1   50  3  5  QP
Q2   99  3  4  QN
Q3   8   6 30  QN
Q4   10  7 30  QP
R3   5   6     300E3
R4   4   7     300E3
*C1   99  6     8.8E-15
*C2   50  7     8.8E-15
*
* INPUT ERROR SOURCES
*
GB1  99  1     POLY(1)  1  22  150E-9  90E-9
GB2  99 30     POLY(1)  1  22  200E-9  90E-9
VOS  3   1     50E-6
LS1  30  2     1E-8
CS1  99  2     1E-12
CS2  50  2     1E-12
*
EREF 97  0     22  0  1
```

```
*
* GAIN STAGE & DOMINANT POLE
*
R5   12 97     3E6
C3   12 97     5.5E-12
G1   97 12     99  8  1E-3
G2   12 97     10 50  1E-3
V3   99 13     4.3
V4   14 50     4.3
D3   12 13     DX
D4   14 12     DX
*
* POLE AT 70 MHZ
*
R8  17 97     1E6
C4  17 97     3.18E-15
G4  97 17     12 22  1E-6
*
* POLE AT 300 MHZ
*
R12 21 97     1E6
C8  21 97     0.318E-15
G8  97 21     17 22  1E-6
*
* OUTPUT STAGE
*
ISY 99 50     5.1E-3
```

```
R13 22 99     16.7E3

R14 22 50     16.7E3

R15 27 99     30

R16 27 50     30

L2  27 28    6E-8

G9  25 50     21 27  33.33E-3

G10 26 50     27 21  33.33E-3

G11 27 99     99 21  33.33E-3

G12 50 27     21 50  33.33E-3

V5  23 27     0.5

V6  27 24     0.5

D5  21 23     DX

D6  24 21     DX

D7  99 25     DX

D8  99 26     DX

D9  50 25     DY

D10 50 26     DY

* MODELS USED

.MODEL QN   NPN(BF=1E9 IS=1E-15)

.MODEL QP   PNP(BF=1E9 IS=1E-15)

.MODEL DX   D(IS=1E-15)

.MODEL DY   D(IS=1E-15 BV=50)

.ENDS AD844
```