DESIGN OF ACTIVE FILTERS USING A NEW ACTIVE DEVICE CDBA

by

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I certify that this thesis satisfies all the requirements as a thesis for the degree of Master of Science.

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This is to certify that I have read this thesis and that in my opinion it is fully adequate, in scope and quality, as a thesis for the degree of Master of Science.

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ABSTRACT

In the last decade, a new versatile active circuit building block called current differencing buffered amplifier (CDBA), has been proposed for the realization of continuous-time filters. Many applications of active filters based on CDBA have been reported in the literature.

This thesis presents three multi-function active filter configurations using CDBAs. Single current controlled CDBA (CC-CDBA) based three new electronically tunable voltage-mode (VM) second-order universal filters using minimum number of active and passive components, a new second-order multi-mode multi-function filter configuration using a single CDBA, and a three CDBA-based single input multiple-output (SIMO) multi-function filter configuration using minimum number of passive elements.

The proposed multifunction filter transfer functions are analyzed for both ideal and non-ideal cases. A good sensitivity performance for all configurations is obtained. Validity of presented topologies are verified through PSPICE simulations some of which are demonstrated experimentally.

Keywords: Current Differencing Buffered Amplifier (CDBA), Current Controlled CDBA, Active Filters, Multi-function Filters, Multi-mode Filters.

YENİ BİR AKTİF ELEMAN OLAN AFATK İLE AKTİF FİLTRELERİN TASARIMI

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ÖZ

Son on yılda sürekli zamanlı filtrelerin gerçekleştirilmesi için akım farkı alan tampon kuvvetlendirici (AFATK) isimli yeni bir çok yönlü aktif devre yapı bloğu önerilmiştir. AFATK tabanlı birçok aktif filtre uygulamaları literatürde bildirilmiştir.

Bu tez AFATK kullanan üç adet çok-işlevli aktif filtre yapısını sunmaktadır: Bir akım kontrollü AFATK tabanlı elektronik olarak ayarlanabilir ve en az sayıda aktif ve pasif eleman kullanan üç adet yeni gerilim modlu ikinci dereceden genel filtre yapısı, bir AFAKT kullanan ikinci dereceden çok modlu çok işlevli yeni bir filtre yapısı, en az sayıda pasif eleman kullanan üç AFAKT tabanlı tek girişli çok çıkışlı (TGÇÇ) çok işlevli filtre yapısı.

İdeal ve ideal olmayan durumlar için önerilen çok işlevli filtre geçiş fonksiyonları analiz edilmiştir. Bütün yapılar için iyi bir duyarlılık başarımı elde edilmiştir. Önerilen yapıların geçerliliği PSpice benzetimleri ile doğrulanmış bazıları ise deneysel olarak ispatlanmıştır.

Anahtar Kelimeler: Akım Farkı Alan Tampon Kuvvetlendirici (AFATK), Akım Kontrollü AFAKT, Aktif Filtreler, Çok İşlevli Filtreler, Çok Modlu Filtreler.

DEDICATION

To my parents

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SYMBOL/ABBREVIATION

b	Oscillation condition
k	Boltzman's constant
q	Electron charge
K	Kelvin in temperature
Q	Quality factor
R_n	Parasitic input resistance of *n*-terminal
R_p	Parasitic input resistance of *p*-terminal
T	Temperature in K
V_T	Thermal voltage
α_p	Current gain of *p*-terminal
α_n	Current gain of *n*-terminal
ε	Tracking error
ε_n	Tracking error of *n*-terminal
ε_p	Tracking error of *p*-terminal
β	Current gain of *w*-terminal
ω_0	Undamped natural frequency
ω_0 / Q	Bandwidth
AFAKT	Akım farkı alan tampon kuvvetlendirici
AP	All-pass
BJT	Bipolar junction transistor
BP	Band-pass
BR	Band-reject
CCII	Second generation current conveyor
CCII+	Non-inverting second generation current conveyor

CCII-	Inverting second generation current conveyor
CCIII	Third generation current conveyor
CC-CDBA	Current controlled current differencing buffered amplifier
CDBA	Current differencing buffered amplifier
CFA	Current feedback amplifier
CM	Current-mode
CMOS	Complementary metal-oxide-semiconductor
DCCCS	Differential current controlled current source
DCFA	Differential-input current feedback amplifier
DDCC	Differential difference current conveyor
DVCC	Differential voltage current conveyor
ECCII	Electronically controlled current conveyor
FTFN	Four terminal floating nullor
HP	High-pass
IBP	Inverting band-pass
KHN	Kerwin-Huelsman-Newcomb
LP	Low-pass
MCO	Minimum component oscillator
MISO	Multi-input single-output
MOS	Metal-oxide-semiconductor
MOSFET	Metal-oxide-semiconductor filed-effect transistor
NBP	Non-inverting band-pass
NF	Notch filter
NIC	Negative impedance converter
NMOS	n-channel MOSFET
OMA	Operational mirrored amplifier
OP-AMP	Operational amplifier
OTA	Operational transconductance amplifier
OTRA	Operational transresistance amplifier
PID	Proportional-integral-derivative
PMOS	p-channel MOSFET
R-L	Resistor-inductor
SFG	Signal flow graph
SIMO	Single-input multiple-output

TAM	Transadmittance-mode
TGÇÇ	Tek girişli çok çıkışlı
TIM	Transimpedance-mode
VCO	Voltage controlled oscillator
VIV	Voltage-current-voltage
VM	Voltage-mode
VMUF	Voltage-mode universal filter

CHAPTER 1

INTRODUCTION

A filter basically can be described as a device that passes electric signals at certain frequencies or frequency ranges while rejecting the passage of others. Filter circuits are used in a wide variety of applications.

Band-pass (BP) filters are used in electronic systems to separate a signal at one frequency or within a band of frequencies from signals at other frequencies. For instance, in the field of telecommunication, BP filters are used in the audio frequency range for modems and speech processing. High-frequency BP filters are used for channel selection in telephone central offices. Notch filters (NF) are used to remove an unwanted frequency from a signal, while affecting all other frequencies as little as possible. System power supplies often use NFs to suppress the line from high frequency transients. Low-pass (LP) filters are used whenever high frequency components must be removed from a signal. Data acquisition systems usually require anti-aliasing LP filters as well as LP noise filters which are placed at the output of an amplifier, and with its high cut-off frequency it allows the desired signal frequencies to pass so that the overall noise level is reduced. High-pass (HP) filters are used in applications requiring the rejection of low-frequency signals. One such application is in high-fidelity loudspeaker systems in which high frequency tweeters would be damaged by the power of low frequency components. All-pass (AP) filters are typically used to introduce phase shifts to each frequency component of a signal in order to cancel or partially cancel any unwanted phase shifts previously imposed upon the signals by other circuitry or transmission media (National Semiconductor, 2000).

At high frequencies, all of the filters usually consist of passive components such as resistors (R), inductors (L) and capacitors (C). They are then called as passive RLC filters. For the passive filters, inductor values become very large for low frequency operations and inductor itself gets quite bulky, which makes economical production difficult. In these cases, active filters play an important role. Active filters are circuits that use a voltage-mode (voltage input voltage output mode or VM) or current-mode (current input current output mode or CM) active device in combination with some resistors and capacitors to provide an RLC-like filter performance at low frequencies. These filters can be designed by resistors (Active R Filters), or by capacitors (Active C Filters) or by both of them (Active RC Filters). Active filters are superior over passive ones by exclusion of inductor element, providing signal gain, input and output impedances are more desirable than passive configurations and simplicity in design process (Kugelstadt, 2001). Besides, new integrated circuit technologies allow designers to construct resistors and capacitors in micro scales. Thus active elements are easily designed in very small dimensions.

An operational amplifier (OP-AMP), an integrated circuit that amplifies the difference between two input voltages and produces a single output, is the most commonly used active element and once used in analog computers to perform mathematical operations of solving differential and integral equations in the early 1960s. First commercially available OP-AMP, μA-709, was introduced by Fairchild Semiconductor in 1965 and classic μA-741 in late 1960s (Neamen, 2001). OP-AMPs have been used widely in circuit design with their high input and low output impedances. Although this wide-spread usage in circuit design, OP-AMP has the restriction of gain-bandwidth product (Wilson, 1990), (Kumar and Shukla, 1985). This means designer can produce a maximum 1 MHz distortion-free signal at the output with a gain of maximum one (National Semiconductor, 2000).

In recent years a new way of description in circuits has evolved. Some designers realized that choosing current as the active variable of a circuit and finding the transfer function for current variables have some advantages over choosing the active variable as voltage. This advance leads them to focus on CM circuits. Some CM active elements proposed instead of VM active elements (i.e. OP-AMPs) for their advantages on bandwidth, slew rate, linearity, high dynamic range, low-power consumption and simple circuitry (Toumazou et al., 1990). Also CM active elements need less power supply voltages than VM ones and the active variable current can be controlled easily by lower power supplies. This statement concludes that analog CM active elements can be used with traditional 5 V supplied digital circuits easily.

Several disadvantages mentioned at the previous paragraph steered designers towards the CM active elements. Now, CM active elements are designed by the dint of evolution in integrated circuit design technology. Some of them are operational transconductance amplifier (OTA), inverting and non-inverting second generation current conveyor (CCII-, CCII+), third generation current conveyor (CCIII), electronically controlled current conveyor (ECCII), differential voltage current conveyor (DVCC), differential difference current conveyor (DDCC), four terminal floating nullor (FTFN), operational mirrored amplifier (OMA), current feedback amplifier (CFA) and current differencing buffered amplifier (CDBA).

The subject of this thesis is the design of active filters using the new active element CDBA. The second chapter introduces CDBA with its block diagram, ideal and non-ideal characteristic equations with equivalent circuits. Subsequently, a comprehensive summary of proposals on CDBA from the first introduction to present are summarized with their advantages and disadvantages. Afterwards some of the implementations and modifications of CDBA are proposed to bring a conclusion to chapter.

The third chapter proposes electronically controllable biquads realized by using a single CDBA. Three new electronically tunable filter configurations which employ only one current controlled current differencing buffered amplifier (CC-CDBA) are proposed. The proposed filters operate in VM and use three passive elements. Each proposed filter can realize one of the BP, LP or HP filter responses. The parameters undamped natural frequency ω_0 and bandwidth ω_0 / Q enjoy independent electronic tunability. Effects of the non-idealities and the sensitivity analyses of the proposed filter are also examined. The validities of the proposed filters are verified through PSpice simulations.

Fourth chapter introduces a new multi-mode second order filter using only a single CDBA. This filter topology realizes LP, BP and HP filter responses in CM, VM, transimpedance-mode (TIM) and transadmittance-mode (TAM) by selection of required admittances. Multi-mode usability is the keynote of the proposal since there have not existed any multi-mode topologies using a single CDBA in the literature. Derivation and analyses of ideal and non-ideal filter characteristics are given in detail. Sensitivity investigations and simulations of all derived filter functions are also given subsequentially. Experimental verification is done through derived VM filter configurations and a new CDBA implementation due to insufficient CDBA response is designed.

Chapter five is devoted to a new VM single-input multiple-output (SIMO) multi-function filter using CDBAs. This multi-function filter is a further investigation on a presented work (Tangsrirat and Surakampontorn, 2005) in order to reduce number of used passive elements and realizes five conventional filter transfer functions, namely LP, BP, HP, AP and band-reject (BR) by using three CDBAs and eight passive elements. The given configuration is arranged in eight combinations of two admittances. For all of the combinations, ideal and non-ideal transfer functions are derived and resulting filter characteristics with requested conditions are given. Sensitivity analyses of filter characteristics are investigated with ideal and non-ideal conditions. Simulations of all obtained filters are demonstrated and a single topology that realizes all conventional filter characteristics is given.

Finally, conclusions for the thesis are presented in chapter six.

CHAPTER 2

CURRENT DIFFERENCING BUFFERED AMPLIFIER (CDBA)

In this chapter firstly, the new five terminal active component CDBA (Acar and Ozoguz, 1999) is presented with its block diagram and terminal definitions. Subsequently, mathematical model of CDBA is summarized by ideal and non-ideal characteristic equations and equivalent circuits. The literature proposals using CDBA from the first introduction to present are summarized at the next section. Following, some of the proposed implementations and modified CDBA-based new active elements in the literature are given for concluding the chapter.

2.1 MATHEMATICAL MODEL OF CDBA

CDBA is a five terminal (two inputs, two outputs and a ground) active element that can be used in CM and VM analog signal processing circuits and filters (Acar and Ozoguz, 1999). *p*-terminal of CDBA is named as positive (non-inverting) input and *n*-terminal is named as negative (inverting) input. Both of the input terminals are internally grounded which makes them preferable current inputs. *z*-terminal is a current output and *w*-terminal is voltage output that follows the potential at the *z*-terminal with respect to ground which is why called as voltage output. Low impedance inputs at *p* and *n*-terminals with low impedance output at *w*-terminal make CDBA a suitable cascadable active element. The block diagram of CDBA is shown in Fig. 2.1.

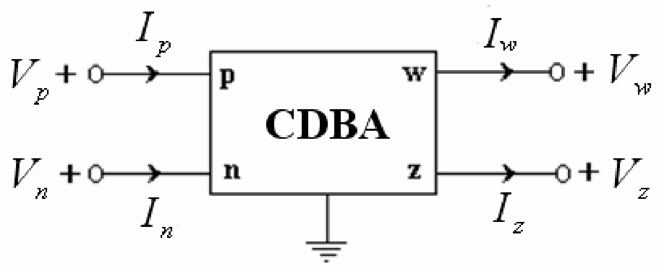

Figure 2.1 Block diagram of CDBA.

2.1.1 Ideal Case

Characteristic equation of CDBA for the ideal case is expressed with the following matrix equation,

$$\begin{bmatrix} I_z \\ V_w \\ V_p \\ V_n \end{bmatrix} = \begin{bmatrix} 0 & 0 & 1 & -1 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} V_z \\ I_w \\ I_p \\ I_n \end{bmatrix}. \tag{2.1.1}$$

From the Eq. (2.1.1), the following equalities can be obtained.

$$V_p = V_n = 0, \tag{2.1.2}$$

$$I_z = I_p - I_n, \tag{2.1.3}$$

$$V_w = V_z. \tag{2.1.4}$$

According to the above equations, this element converts difference of the input currents I_p and I_n into the output voltage V_w through the impedance which will be connected to *z*-terminal. Therefore CDBA can be considered as a transimpedance amplifier and from this viewpoint it is similar to CFA (Manetakis and Toumazou, 1996). Besides, the differential nature at the input makes this element especially suitable for fully integrated filter implementations.

An equivalent circuit of CDBA can be modeled as I_p and I_n current inputs and two outputs; one is a dependent voltage source output of V_w that follows the voltage source V_z and the other is a dependent current source output I_z that is equal to the difference of input currents I_p and I_n. Equivalent circuit is given in Fig. 2.2.

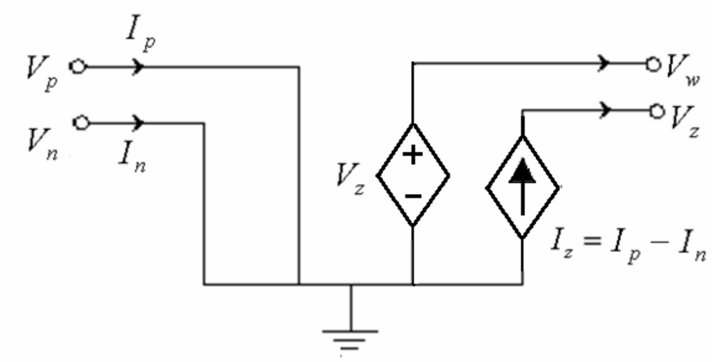

Figure 2.2 Equivalent circuit of CDBA.

2.1.2 Non-Ideal Case

Current input terminals and voltage output *w*-terminal of CDBA are affected by parasitic resistances. Also current output *z*-terminal has a parasitic resistance and capacitance due to the buffer that converts the current at *z*-terminal into the voltage seen at *w*-terminal (Maheshwari and Khan, 2005). The equivalent circuit for non-ideal CDBA is given below.

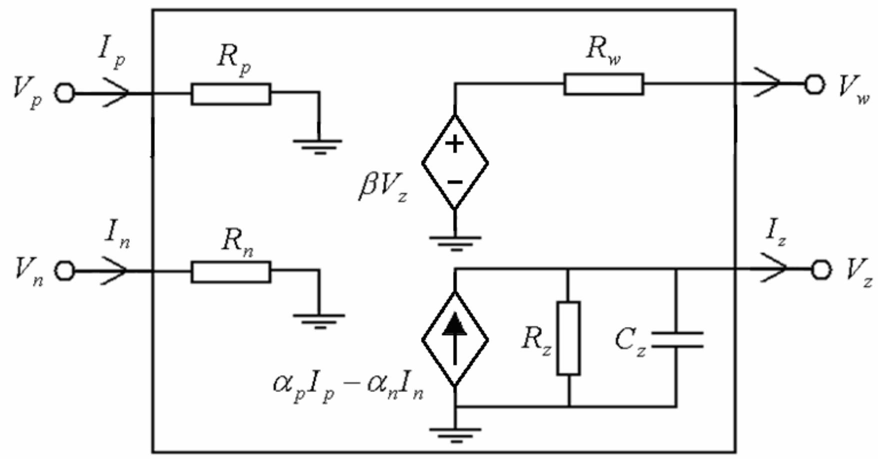

Figure 2.3 Equivalent circuit of non-ideal CDBA.

Taking the non-idealities of the CDBA into account by excluding the parasitic passive elements shown in Fig. 2.3 the relationship of the terminal voltages and currents can be written as:

$$\begin{bmatrix} I_z \\ V_w \\ V_p \\ V_n \end{bmatrix} = \begin{bmatrix} 0 & 0 & \alpha_p & -\alpha_n \\ \beta & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} V_z \\ I_w \\ I_p \\ I_n \end{bmatrix}. \tag{2.1.5}$$

The above characteristic equation includes current gains α_p and α_n from *p* and *n*-terminals and voltage transfer gain β from *z*-terminal to *w*-terminal. These gains are expressed as

$$\begin{aligned} \alpha_p &= 1-\varepsilon_p, \\ \alpha_n &= 1-\varepsilon_n, \\ \beta &= 1-\varepsilon_v. \end{aligned} \tag{2.1.6}$$

where $\varepsilon_p, \varepsilon_n$ and ε_v are their tracking errors with

$$\begin{aligned} &|\varepsilon_p| << 1, \\ &|\varepsilon_n| << 1, \\ &|\varepsilon_v| << 1. \end{aligned} \tag{2.1.7}$$

2.2 LITERATURE SURVEY ON CDBA

CM circuits are receiving much attention for their potential advantages such as inherent wider bandwidth, simple circuitry, lower power consumption and wider dynamic range (Roberts and Setra, 1989). In the last decade, new active building blocks like CCII+ and CCII-, CFA received considerable attention due to their larger dynamic range and wider bandwidth.

Acar and Ozoguz have realized their proposed active component, namely CDBA by using two commercially available AD844 (Analog Devices, 1990) CFAs. In their paper they also proposed how to realize an *n*-th order voltage transfer function by using signal flow graph. They had designed some of the signal flow graph components with CDBA and by using these realizations. They had also designed an AP filter in agreement with the theory of signal flow graph. This method is straightforward and simple in design that it gives not only the configuration but also the element values directly from the coefficients of *n*-th order transfer function. Nevertheless, they claimed to select the impedance-scaling factor properly by using statements of Svodoba (Svodoba et al., 1991) so that the effects of parasitic input impedances of CDBA would not reduce the performance of the filters.

Ozoguz et al. proposed CM continuous-time fully-integrated universal filter using CDBAs (Ozoguz et al., 1999). In this work a new complementary metal-oxide-semiconductor (CMOS) implementation of CDBA is given. A lossy-integrator based on CDBA is proposed and a fully integrated MOSFET-C filter is presented. A new CM universal filter illustrating the usefulness of this element is also provided. This universal filter uses a grounded capacitor and four lossy MOS (metal-oxide-semiconductor) devices in the realization which is less than negative impedance converter (NIC), CCII+ and CFA based realizations proposed earlier (Tsao et al., 1991), (Liu et al., 1990), (Meng and Yu, 1996).

Toker et al. proposed CM Kerwin-Huelsman-Newcomb (KHN) equivalent biquads using CDBAs (Toker et al., 1999). KHN biquads offers several advantages such as low passive and active sensitivity performance, low component spread and good stability (Kerwin et al., 1967). Proposed multi-input single-output (MISO) CM universal filter is derived from the adjoint signal-flow graph method applied on the classical KHN circuit and CDBA is used as active element. The topology has three inputs and a single output. Input selection decides which kind of biquad is realized by the topology. The proposed topology has the advantages of classical KHN circuit, grounded capacitors and easy cascadibility in CM.

Acar and Ozoguz proposed *n*-th order current transfer function synthesis using signal-flow graph approach (Acar and Ozoguz, 2000). Sub-graphs of signal flow graph model, a current distributor and a current integrator are designed by CDBAs using signal flow graph method. A third order AP filter topology using these sub-graphs is also realized in this work. Topology uses n capacitors, $3n+3$ resistors and $n+2$ CDBAs for the *n*-th order current transfer function. A suggestion for parasitic effects is given as selection of impedance scaling factor properly as stated by Svoboda (Svodoba et al., 1991).

Ozcan et al. proposed a single resistance controlled six oscillator topologies (Ozcan et al., 2000a). Design considerations of topologies achieve orthogonal control of oscillation condition *b* and ω_0. All six topologies employ two capacitors, three resistors (2C-3R) and a single CDBA. Oscillation frequency is controlled by a grounded resistance easily, which permits these topologies to be used as voltage controlled oscillator (VCO).

Toker et al. proposed two different types of first order canonical AP CM filters using CDBA and a new high-Q BP filter configuration (Toker et al., 2000). A simplified implementation of CMOS CDBA is also suggested in this work. The proposed AP filters use one resistance, one capacitor and one CDBA which uses less passive and active components than the other proposed active first-order AP filters. Voltage output terminal is not used and therefore in fully integrated CM realization it is not necessary to implement the buffer stage on chip. BP filter uses five resistors, three capacitors and three CDBAs and designed by using first-order AP filter network topology repeatedly.

In 2001, Salama et al. proposed a new circuit configuration for constructing a universal biquad filter by CDBAs (Salama et al., 2001). This configuration realizes six filter functions by using two CDBAs and seven passive components. Configuration offers independent control on quality factor (*Q*).

Tarim and Kuntman proposed a high performance CDBA implementation in 2001 (Tarim, N., and Kuntman, H., 2001). The new implementation employs two high performance CCIIs and a voltage buffer that offers a very low gain error with low output impedance (Palmisano et al., 2000). The suggested structure of CDBA can be easily implemented in CMOS technology. The performance of the new CDBA is demonstrated by an application example of first-order AP filter circuit topology and simulations.

Zeki et al. proposed a modified CDBA, differential-input current feedback amplifier (DCFA) by annulling the ground at *y*-terminal of conventional CDBA (Zeki et al., 2001). Afterwards they had revealed that with two additional MOSFETs a DCFA can be used as a linearly tunable transconductor. The configuration can be used as a variable-gain amplifier or an analog multiplier.

Ozcan et al. presented cascadable CM multipurpose filters in 2002 (Ozcan et al., 2002). Proposed filter basically realizes BP and LP filter current transfer functions by using two capacitors, three resistors and a single CDBA. By using an additional resistor and CDBA, two HP filter current transfer functions are also realized. The configuration is remarkable with its independent control of Q by a single grounded resistance.

Tangsrirat et al., proposed a possible realization of low-voltage CDBA in bipolar technology (Tangsrirat et al., 2004). A leapfrog simulation of the CM ladder network using CDBA is also introduced in this work. Two sub-circuits are designed by CDBAs using block diagram of the Leapfrog structure. A CM fifth order Butterworth LP filter is obtained by using these sub-circuits. In the same way, parallel and series LC circuits are realized by CDBAs and a CM sixth order Cheyshev BP filter design is suggested by using a sixth order RLC circuit.

Acar and Sedef proposed two possible circuit configurations to realize the most general *n*-th order CM transfer function by using $RC-RC$ decomposition technique (Acar and Sedef, 2003). One of the proposed circuits uses one CDBA with six passive elements and the other uses two CDBAs with four passive elements. As an example, a third-order normalized AP Butterworth filter function is obtained and simulated using $RC-RC$ decomposition technique (Lam, 1979). As a conclusion it is proved that any current transfer function can be realized by using one or two CDBAs in each section, in which less CDBAs are used in comparison to previously reported method (Acar and Ozoguz, 2000).

SFG simulation of general ladder filters using CDBA is presented in 2003 (Biolek and Biolkova, 2003). Voltage-Current-Voltage (VIV) flow graphs are also introduced as a supporting tool of synthesis of leapfrog active filter and for its subsequent optimization.

A technique of passive elliptic ladder RLC filter simulation by CDBAs is suggested (Biolek et al., 2003). An inductance simulation is obtained by using three CDBAs, four resistors and one capacitor. Afterwards, filter transfiguration of some circuits are given with the explanations. Then a fifth-order Cauer elliptic LP filter is realized by using inductance simulation and transfiguration perspective.

An integratable technique to realize a low-voltage wide-band n-channel MOSFET (NMOS) based CDBA is introduced in 2004 (Tangsrirat et al., 2004). The method to realize such a CDBA is based on the modification of CCII+ by using NMOS realization advantages over p-channel MOSFET (PMOS) one such as low-input resistance and wide bandwidth. Also a three-input and single-output CM universal biquadratic filter is presented in this paper. ω_0 and Q are independently controllable through the passive elements and have low passive and active sensitivities.

Cam proposed a novel CM second order NF configuration (Cam, 2004). Proposed circuit the configuration uses a single CDBA, four capacitors and four resistors. This letter shows the advantage of CDBA over the other introduced active elements with a NF configuration and uses minimum number of active and passive elements in comparison to previously reported ones in the literature.

In 2004, Kilinc and Cam suggested CDBA based CM filters (Kilinc and Cam, 2004). The circuit configuration uses total four admittances (two resistances and two capacitors) and one CDBA. Depending on the choice of admittances, configuration behaves in LP, HP, BP and NF characteristics.

Keskin proposed design of minimum component oscillators (MCO) using NICs approach (Keskin, 2004a). Using this approach, he had designed CCII, CDBA, operational transresistance amplifier (OTRA) and OPAMP based oscillators with two resistors and two capacitors.

Keskin proposed a new four quadrant analog multiplier configuration employing single CDBA (Keskin, 2004b). Proposed configuration uses one CDBA and four NMOS transistors and is in minimum of active and passive elements in comparison to previously reported ones in literature (Liu et al., 1996), (Babanezhad and Temes, 1985). Controllable multiplication factor is another important feature of the circuit suggested in this letter.

A new four terminal CDBA based current-controlled active element, CC-CDBA is introduced by Maheshwari and Khan (Maheshwari and Khan, 2004). A BJT implementation of the proposed new active element is also given. This active element accomplishes electronic adjustability in CDBA by using external current input through internal parasitic input resistances. Usefulness of CC-CDBA is shown by four BP configurations. One of these configurations uses an inductance simulation realization in which value of inductance can be controlled by external current source that changes value of internal resistances. Depending on inputs and outputs, HP and LP transfer function characteristics are easily obtained. Independent and orthogonal tuning of pole-ω_0 and pole-Q are other advantages of proposed filters.

Keskin and Hancioglu proposed a CM multifunction filter using CDBAs (Keskin and Hancioglu, 2005a). Suggested multifunction filter uses four capacitors, four resistors and two CDBAs. The topology uses grounded capacitors, which is an important aspect in IC implementation. Also, Q of the topology can be controlled independently.

A novel voltage-mode universal filter (VMUF) based on CDBA is proposed in 2005 (Maheshwari and Khan, 2005). The proposed VMUF uses two CDBAs to realize six filter responses; LP, HP, inverting band-pass (IBP), non-inverting band-pass (NBP), BR and AP. Moreover, the proposed configuration enjoys attractive features such as independent filter parameter control and low sensitivity measures.

A general synthesis procedure is given for a versatile signal flow graph realization of a general transfer function by using CDBA in 2007 (Koksal and Sagbas, 2007). Some of the signal flow sub-graphs are realized by CDBA-based sub-circuits, and interconnection of sub-circuits are simplified by a final configuration. The proposed configuration uses n+1 CDBAs, n capacitors and 2n+4 resistors for a general voltage ratio transfer function and n+3 less resistors for nth-order TIM multifunction filter realization. The configuration also promises conversion to different modes of operations, such as TIM, TAM, VM and CM, by the addition and/or removal of a single resistor at the terminations.

Keskin and Hancioglu proposed two CDBA-based floating inductor simulator circuits (Keskin and Hancioglu, 2005b). These circuits are fully integrable and have voltage tuning properties. Furthermore, they can be easily converted into fully integrable and linearly tunable resistance scaling circuits.

A realization of VM SIMO biquadratic filter configuration using CDBAs is presented in 2005 (Tangsrirat and Surakampontorn, 2005). Two different circuit topologies based on the use of the CDBA-based cross-coupled feedback configuration and the voltage substractor are possible from the same circuit configuration depending on the proper selection of the virtually grounded passive components. The circuit presented realizes LP, HP, BP, BR, AP filter voltage transfer functions simultaneously. The filters also provide independent control of ω_0 and Q. One of the circuit topologies proposed for the SIMO filter uses three CDBAs, seven resistors and two capacitors and the other uses three CDBAs, seven capacitors and two resistors.

A realization of OTA-based CDBA is proposed by Kaewpoonsunk et al. in ICCAS2005 (Kaewpoonsunk et al., 2005). This CDBA offers low-cost and electronically tunability advantages by using four OTAs in realization. In order to demonstrate the performance of the proposed realization, first order AP filters of Tarim and Kuntman (Tarim and Kuntman, 2001) and Ozcan et al. (Ozcan et al., 2000b) were implemented as application examples.

A single CDBA-based four VM high-Q BP filters, two having minimum number of components, are presented in 2005 (Keskin, 2005a). Two of these BP filters use three resistors and three capacitors and the other two use two resistors and two capacitors. Four CDBA-based oscillator circuits are also obtained by slight modification of proposed high-Q BP filter circuits in this paper. Two of these oscillators use two resistors and two capacitors which are in minimum. The configuration is based on the circuit presented by Sedef and Acar (Acar and Sedef, 2003) and offers a further investigation of this circuit.

A multiple-mode filter using CDBA is introduced by Sagbas and Koksal (Sagbas and Koksal, 2005). The filter is originally a current-input voltage-output TIM circuit and can be converted into current-input current-output, voltage-input voltage-output and voltage-input current-input by using one or two resistors. The conversion method is described in the paper. The topology uses three resistors, two capacitors and three

CDBAs in TIM. HP, LP, BP filter characteristics are easily obtained from output *w*-terminals.

Keskin presented four single CDBA-based cascadable VM biquadratic NFs in 2005 (Keskin, 2005b). Three of these configurations consist of three resistors and three capacitors. The remaining can be constituted by four resistors and four capacitors. Cascadibility and grounded capacitors in circuits (except one) claimed to be main advantages of these circuits.

Three single CDBA-based active inductor simulation circuits are presented by Gulsoy and Cicekoglu (Gulsoy and Cicekoglu, 2005). Two of these circuits realize series resistor-inductor (R-L) employing two resistors, one capacitor and one CDBA. The remaining one permits independent control of inductance value and realizes a parallel R-L circuit, which enables also lossless inductor realization with four resistors and one capacitor. Two Application examples using proposed circuits are constructed for demonstrating the performance of configurations.

A proportional-integral-derivative (PID) controller circuit employing CDBAs is introduced in 2006 (Keskin, 2006a). In order to realize PID controller circuit, a signal flow graph of general PID controller transfer function is drawn and sub-graphs are realized by CDBAs. Subsequently, sub-graphs are connected together and final configuration is obtained by using four CDBAs, eight resistors and two capacitors.

A novel CM universal biquadratic filter based on CC-CDBA is presented by Jaikla et al. (Jaikla et al., 2006). The configuration comprises three CC-CDBAs and two grounded capacitors. Proposed circuit offers the advantage of orthogonal control of Q and ω_0.

A VM, MISO type multi-function biquad is proposed by Keskin (Keskin, 2006b). The configuration realizes HP, LP, BP, BR and AP filter functions without changing the circuit topology and contains grounded and virtually grounded capacitances. Furthermore, Q-factor is independently adjustable if ω_0 is fixed. A single CDBA, four resistors and four capacitors are used to implement the proposed configuration.

A MISO CM universal filter based on CDBAs is introduced in 2006 (Tangsrirat and Surakampontorn, 2006). The proposed configuration consists of three CDBAs, two virtually grounded resistors, and two grounded capacitors. Depending on the selection of input signal five standard biquadratic filter functions namely LP, HP, BP, AP, BR are easily obtained from the same topology. Also, ω_0 and the parameter ω_0 / Q are independently controllable.

2.3 IMPLENMENTATIONS AND MODIFICATIONS OF CDBA

Due to the popularity in the last decade, several implementations and modifications of CDBA are presented in the literature. This section briefly summarizes given implementations and modifications of CDBA.

2.3.1 CFA Implementation

When Acar and Ozoguz first introduced CDBA (Acar and Ozoguz, 1999), they had also presented the first implementation of CDBA by composition of two CFAs which are commercially available as AD844 (Analog Devices, 1990). This implementation offers the well-known advantages of CFAs. Fig. 2.4 shows possible implementation of CDBA by using two AD844s.

For the given implementation, the terminal equation (2.1.2) of CDBA is obtained using AD844's voltage transfer characteristic at the input terminals, namely

$$V_+ = V_- . \tag{2.3.1}$$

Using above statement, due to the grounded non-inverting input terminals in Fig. 2.4, Eq. (2.1.2) is easily obtained.

The difference of currents in Eq. (2.1.3) is obtained by using conventional AD844 terminal equation that negates input current at the inverting input and transfers it into current output *z*-terminal. By using this property, I_p current is transferred as $-I_p$ at the *z* output of the first AD844 and summed with I_n. This summation of currents is again inverted by the second AD844 and the final output current is obtained as difference of input currents, i.e. $I_p - I_n$.

Again using another voltage transfer property of AD844 given as

$$V_w = V_z, \tag{2.3.2}$$

we obtain our third terminal equation of CDBA in Eq. (2.1.4).

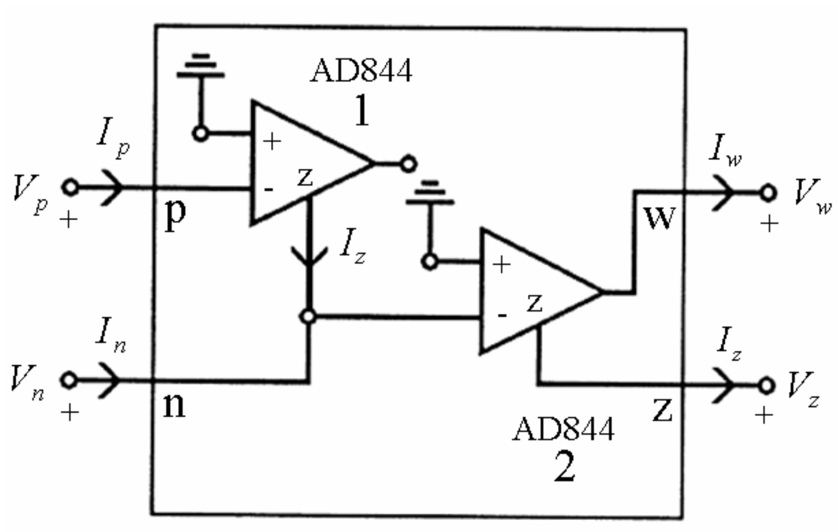

Figure 2.4 CFA (AD844) implementation of CDBA.

2.3.2 CMOS Implementation

Ozoguz et al. proposed a new implementation of CDBA by using MOSFETs which is a combination of a differential current controlled current source (DCCCS) followed by a voltage buffer (Ozoguz et al., 1999). This implementation is the first suggestion of CDBA as a fully integrated circuit. Fig. 2.5 is the proposed CDBA implementation.

As seen in this figure, virtual grounds by using symmetrical supply voltages, the difference of input currents by a DCCCS comprised of current mirrors and a voltage buffer transferring the voltage at the *z*-terminal into *w*-terminal realizes all terminal equations for CDBA.

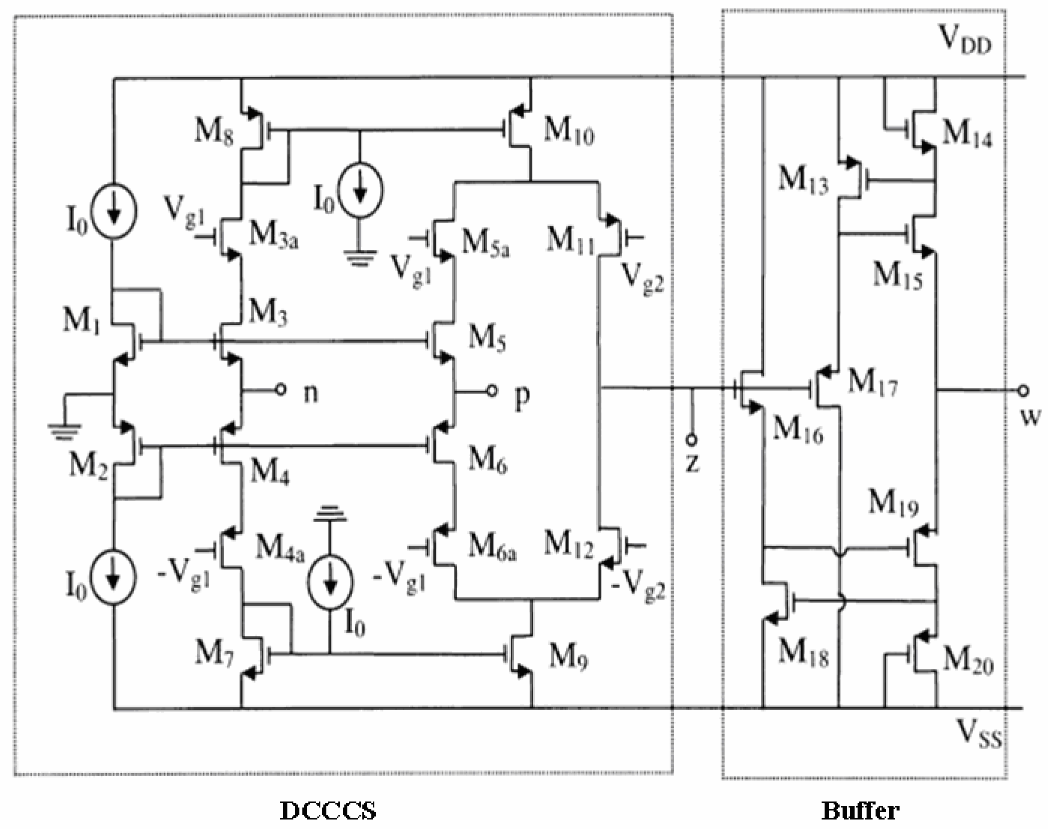

Figure 2.5 CMOS implementation of CDBA.

2.3.3 Simplified CMOS Implementation

A simplified fully integrated CMOS implementation of CDBA is designed by Toker et al. (Toker et al., 2000). Proposed circuit implementation is given in Fig. 2.6. In this circuit aspect ratio of transistor M_3 (M_4) should be twice as large as those of M_1 (M_2) and M_5 (M_6) and simplified current mirrors are used to realize virtual grounds, a DCCCS and a voltage buffer to obtain CDBA characteristics.

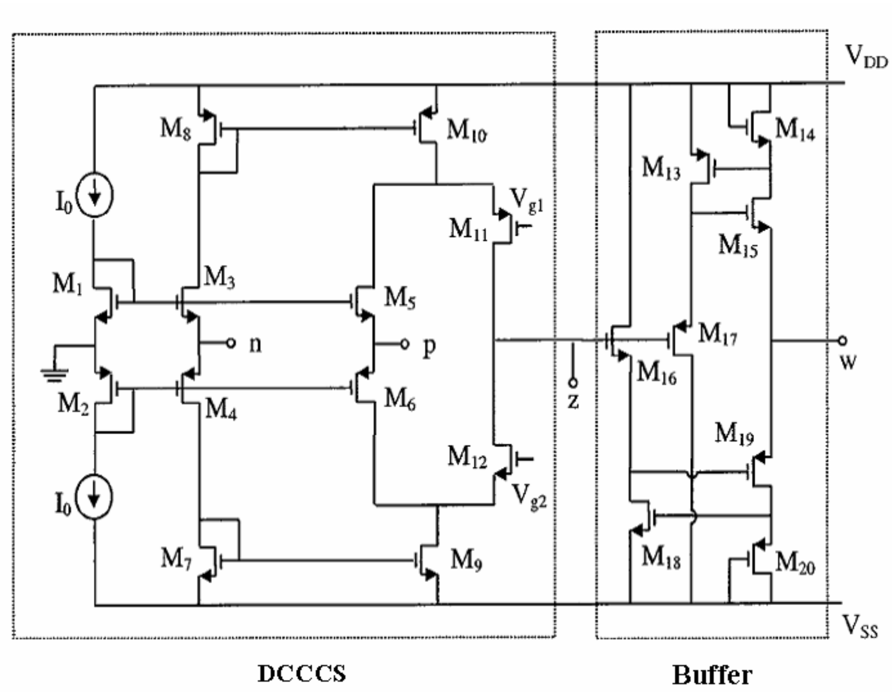

Figure 2.6 Simplified CMOS implementation.

2.3.4 A High Performance CDBA Using Current Conveyors

A high performance CDBA using CCIIs is proposed in 2001 (Tarim, N., and Kuntman, H., 2001). CFA implementation of CDBA (Acar and Ozoguz, 1999) implicitly employs two CCIIs and a voltage buffer. Using this relation, a new CDBA configuration using CCIIs and a voltage buffer is implemented. The overall performance of CDBA is determined by the performance of CCIIs and voltage buffer. Therefore high performance blocks are chosen to implement the circuit. The CDBA contains only MOS transistors and is designed to be implemented in CMOS technology. The CMOS realization of CCII chosen for the CDBA implementation is given in Fig. 2.7. Using improved active-feedback cascade current mirrors; this CCII has high impedances at the *y* and *z*-terminals and high accuracy of the current transfer ratio I_z / I_x. A high performance and simple CMOS unity-gain amplifier, which possesses very low gain error and low output impedance, is used for the voltage buffer block (Palmisano et al., 2000). The proposed voltage buffer is shown in Fig. 2.8.

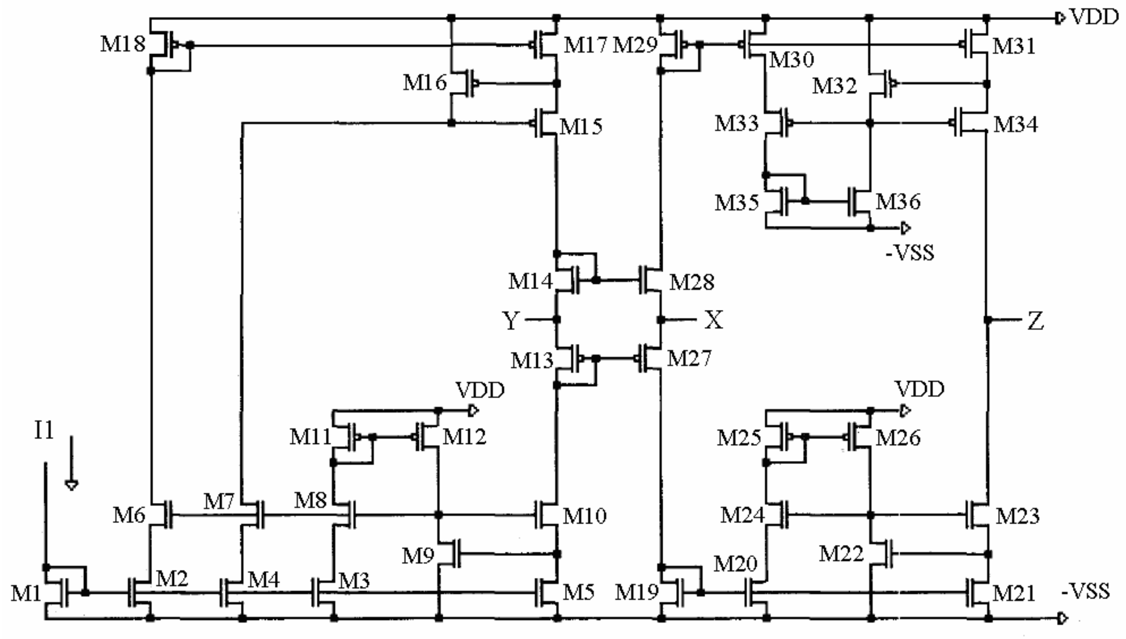

Figure 2.7 CCII used to realize CDBA.

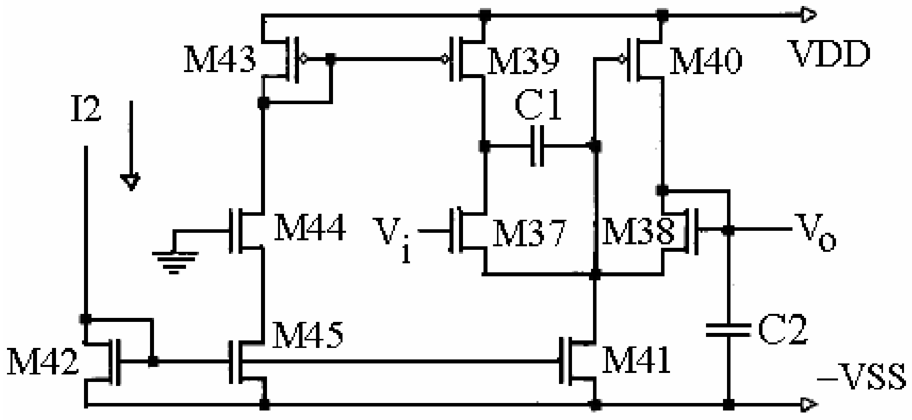

Figure 2.8 The voltage buffer used to realize CDBA.

2.3.5 A Modified CDBA: Differential-input Current Feedback Amplifier (DCFA)

Zeki et al. proposed a new active element called DCFA in 2001 (Zeki et al., 2001). This new active element is actually a modified CDBA. If *y*-terminal of the CDBA is not grounded and used as a high-impedance input terminal, DCFA is obtained. Since DCFA is a modified CDBA, it has all the advantages of CDBA over the other CM and VM active elements. Terminal equations of the proposed new active element are as follows;

$$\begin{aligned} V_p &= V_y, \\ V_n &= V_y, \\ I_y &= 0, \\ I_z &= I_p - I_n, \\ V_w &= V_z. \end{aligned} \tag{2.3.3}$$

Block diagram of DCFA is given in Fig. 2.9, and its CMOS implementation is given in Fig. 2.10.

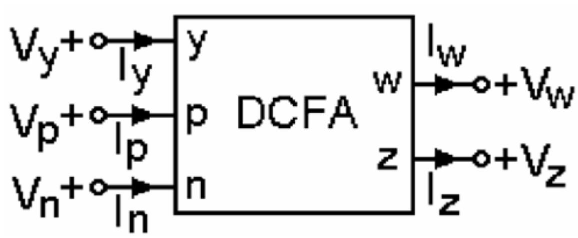

Figure 2.9 Block diagram of modified CDBA or DCFA.

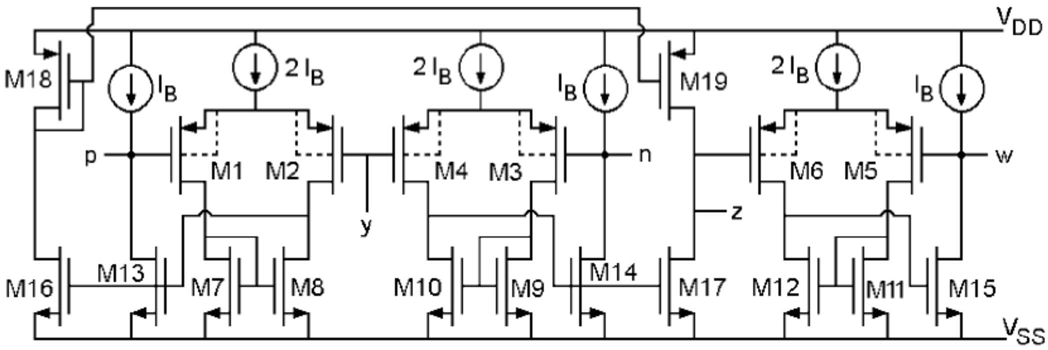

Figure 2.10 CMOS DCFA structure proposed.

2.3.6 A Low-voltage CDBA Implementation in Bipolar Technology

Tangsrirat et al. proposed the first low-voltage bipolar implementation in 2002 (Tangsrirat et al., 2002). The CDBA implementation is given in Fig. 2.11. This implementation is designed by considering two blocks of CDBA, namely current substractor and buffer amplifier. Using the synthesis of a novel CM OP-AMP (Nagasaku et al., 1996), group transistors $Q_1 - Q_4$ and $Q_7 - Q_{10}$ function as low-input resistance input stages. Also complementary NPN and PNP transistors $Q_{14} - Q_{17}$ form the buffer stage that forces the *w*-terminal to the potential of the *z*-terminal.

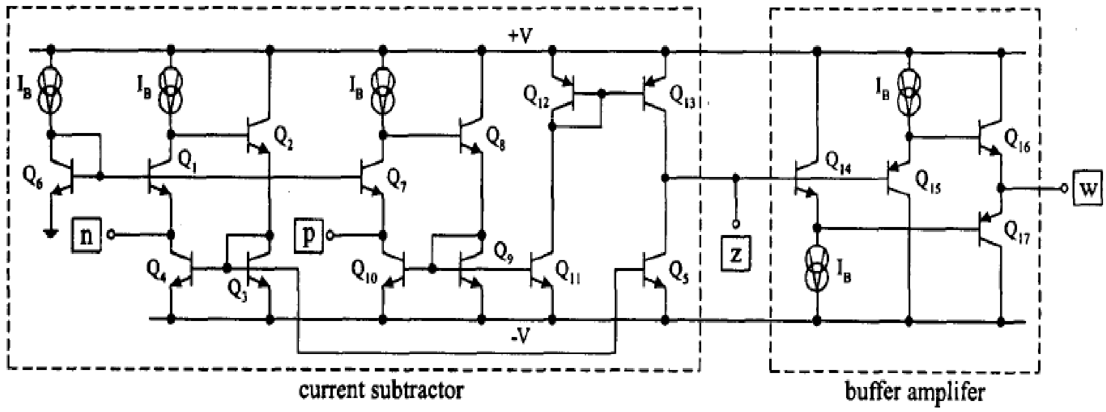

Figure 2.11 Realization of low-voltage CDBA.

2.3.7 A Low-Voltage Wide-Band NMOS-Based CDBA

A fully integratable low-voltage wide-band NMOS-based CDBA designed with a current differentiator and a buffer is presented by Tangsrirat et al. in 2004 (Tangsrirat et al., 2004).

The main purpose of Tangsrirat et al. was to design a CDBA with wide-band operation and low-resistance inputs. Conventionally, a current differencing function is achieved through negative current mirrors using PMOS transistors. Limitation of the high frequency operation effected from PMOS transistors is an important drawback. This shortcoming is overcome by using NMOS transistors in realization of CDBA. Furthermore, the terminal resistances of the PMOS CDBA are quite high. Tangsrirat et al. proposed an NMOS low-input resistance input stage to overcome the mentioned problem. The input resistance of this configuration is calculated as

$$r_{in} = \left(\frac{1}{g_{m1}}\right)\left(\frac{1}{1+F}\right), \qquad (2.3.4)$$

where,

$$F = \left(\frac{g_{m2} g_{m4} r_{0B}}{g_{m2} + g_{m3}} \right), \tag{2.3.5}$$

and g_{mi} represents the transconductance of the transistors $M_i \left(i = 1,2,3,4 \right)$ and r_{0B} denotes the output resistance of the current source I_B. Usually $r_{0B} >> 1/g_{mi}$, then $F >> 1$. Therefore, a very low resistance input is obtained.

Afterwards, they have designed an NMOS-based unity gain current amplifier by using low-input resistance stage and an NMOS-based negative current mirror. The output current i_{out} of the amplifier is expressed as

$$i_{out} = -\left(\frac{F}{1+F} \right) i_{in} \,. \tag{2.3.6}$$

Since $F >> 1$ usually, the output current i_{out} can be approximated to

$$i_{out} \cong -i_{in} \,. \tag{2.3.7}$$

The combination of these three circuits (low-input resistance stage, current amplifier and negative current mirror) forms an NMOS current differencing circuit shown in Fig. 2.12.

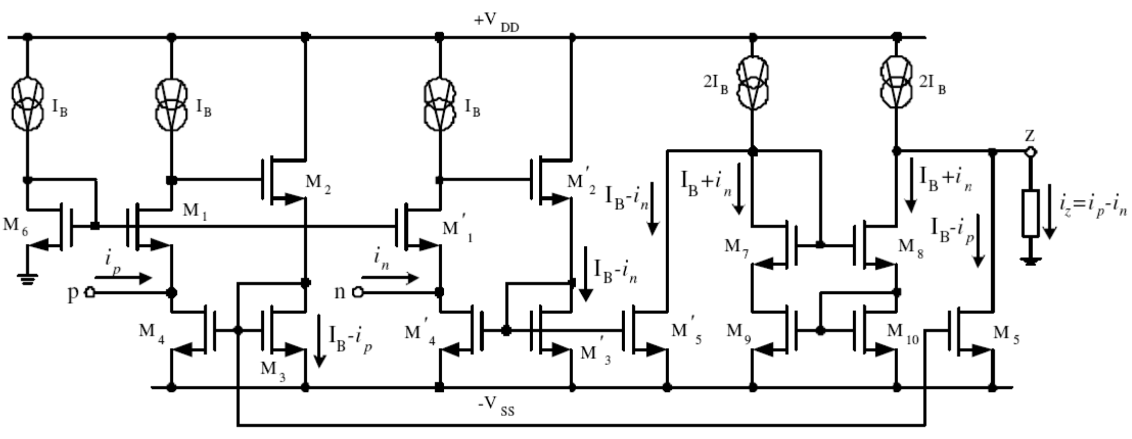

Figure 2.12 NMOS current differencing circuit.

In order to account for the non-ideal performance, let α_p and α_n are the current gains for the inputs from the terminals *p* and *n*, respectively. From the routine analysis, the output current i_z can be given by

$$i_z = \alpha_p i_p - \alpha_n i_n , \tag{2.3.8}$$

where,

$$\alpha_p = \left(\frac{F_p}{1 + F_p} \right). \tag{2.3.9}$$

$$\alpha_n = \left(\frac{g_{m7} g_{m8}}{g_{m9} g_{m10}} \right) \left(\frac{F_n}{1 + F_n} \right), \tag{2.3.10}$$

$$F_p = \left(\frac{g_{m2} g_{m4} r_{0B}}{g_{m2} + g_{m3}} \right), \tag{2.3.11}$$

and,

$$F_n = \left(\frac{g'_{m2} g'_{m4} r_{0B}}{g'_{m2} + g'_{m3}} \right). \tag{2.3.12}$$

Then as long as $F_p >> 1$, $F_n >> 1$ and $g_{m7} \cong g_{m8} \cong g_{m9} \cong g_{m10}$, the current gains $\alpha_p \cong \alpha_n \cong 1$. The input resistances of the terminals *p* and *n* can also be expressed as

$$r_p = \left(\frac{1}{g_{m1}} \right) \left(\frac{1}{1 + F_p} \right), \tag{2.3.13}$$

$$r_n = \left(\frac{1}{g'_{m1}} \right) \left(\frac{1}{1 + F_n} \right). \tag{2.3.14}$$

We can notice that, the input resistance r_p and r_n are very low due to the factors from the feedback $(1 + F_p)$ and $(1 + F_n)$.

Finally, a buffered voltage amplifier with NMOS transistors is used as buffer stage as shown in Fig. 2.13.

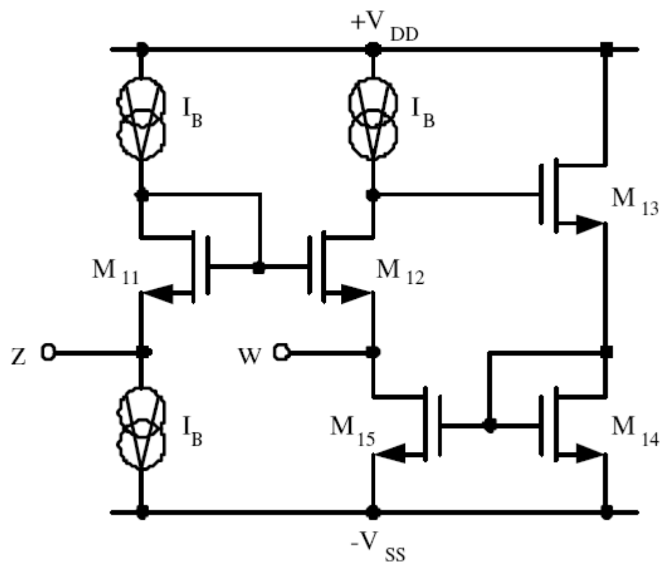

Figure 2.13 NMOS voltage amplifier.

Relationship between input *z* and output *w* of the buffer stage is

$$v_w = \beta_v v_z \,. \tag{2.3.15}$$

where,

$$\beta_v = \left(\frac{g_{m11} r_{0B}}{1 + g_{m11} r_{0B}} \right) \left[\frac{g_{m12}\left(1 + \frac{g_{m15} r_{0B}}{2}\right)}{g_w + g_{m12}\left(1 + \frac{g_{m15} r_{0B}}{2}\right)} \right], \tag{2.3.16}$$

$g_w = 1/R_w$ and R_w is the resistor connected at the terminal *w*. If $g_{m11} r_{0B} >> 1$ and $g_{m12}\left(1 + \frac{g_{w15} r_{0B}}{2}\right)$, then $v_w \cong v_z$. The output resistance of the *w*-terminal is low and equal to;

$$r_w = \left(\frac{1}{g_{m12}} \right) \left(\frac{1}{1 + F_w} \right), \tag{2.3.17}$$

where,

$$F_w = \left(\frac{g_{m13} g_{m15} r_{0B}}{g_{m13} + g_{m14}} \right) \quad . \tag{2.3.18}$$

If $r_{0B} >> 1/g_{m11}$, the input resistance looking into the z-terminal becomes a high value and is approximated as

$$r_z = \frac{r_{0B}}{2} . \tag{2.3.19}$$

The combined integrated figure of the stages is shown below.

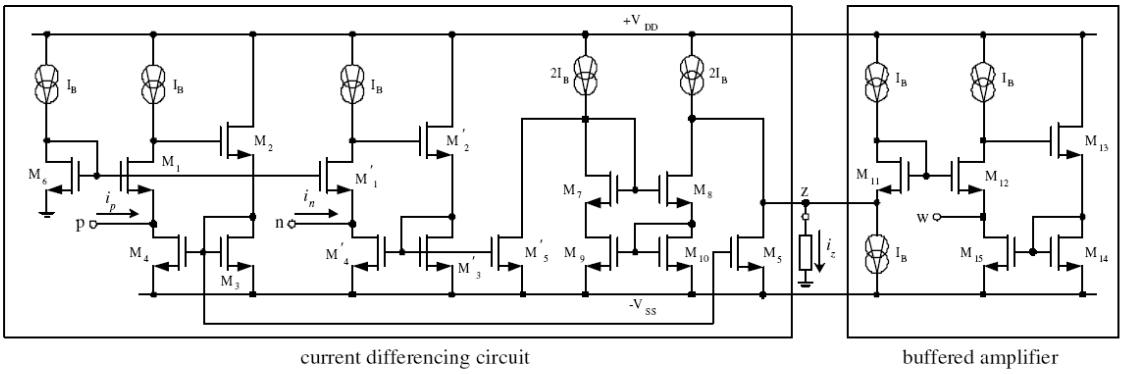

Figure 2.14 Proposed low-voltage NMOS-based CDBA.

2.3.8 Another Modification: Current Controlled CDBA

In order to accomplish electronic adjustability in CDBA, Maheshwari and Khan have introduced current controlled differencing buffered amplifier (CC-CDBA) (Maheshwari and Khan, 2004). Block diagram of CC-CDBA is shown in Fig. 2.15. An equivalent circuit of CC-CDBA is given in Fig. 2.16 (Jaikla et al., 2006) and its terminal equations can be written as follows

$$V_p = R_p I_p , \tag{2.3.20}$$

$$V_n = R_n I_n , \tag{2.3.21}$$

$$I_z = I_p - I_n , \tag{2.3.22}$$

$$V_w = V_z. \tag{2.3.23}$$

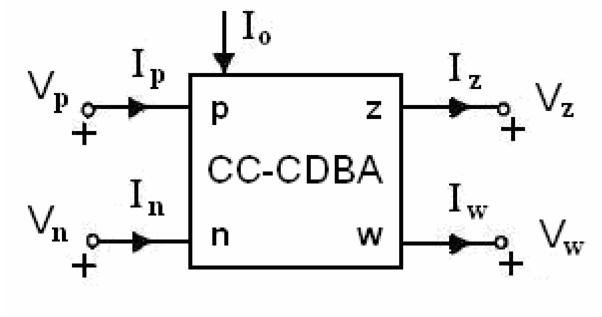

Figure 2.15 Block diagram of CC-CDBA.

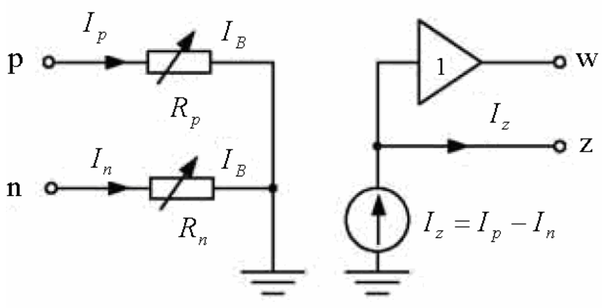

Figure 2.16 Equivalent circuit of CC-CDBA.

Current controlled CDBA can easily be implemented by using bipolar junction transistor (BJT) technologies as shown in Fig. 2.17 (Frey, 1993).

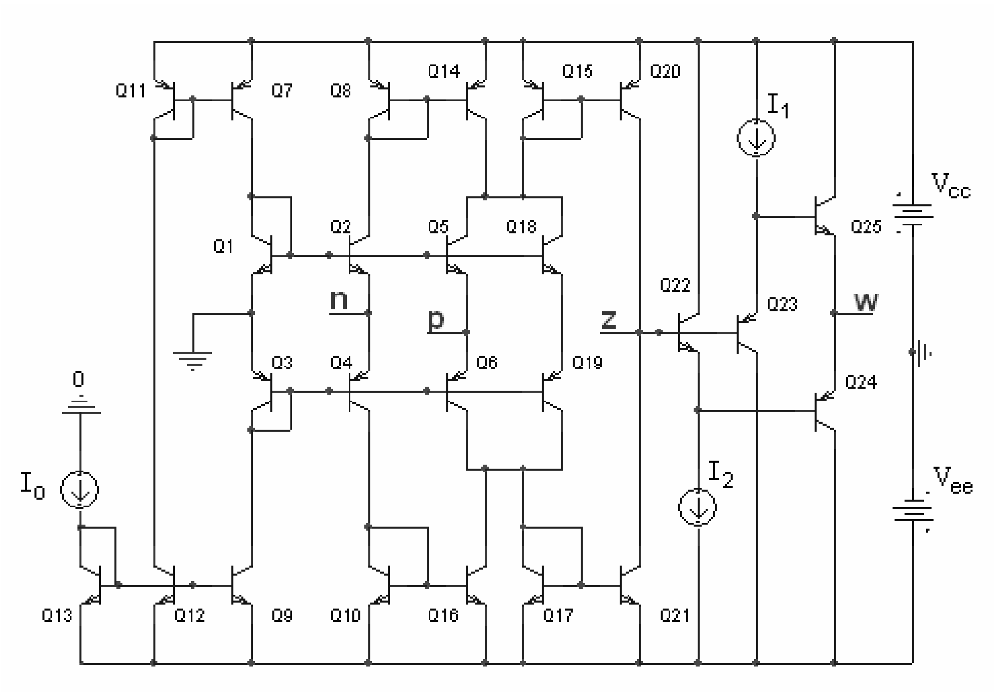

Figure 2.17 Schematic implementation for CC-CDBA using BJT technology.

The parasitic input resistances R_p and R_n using BJT implementation for $I_x(t) << 2I_O$ can be obtained as

$$R_p = R_n = \frac{kT}{2I_0} = \frac{V_T}{2I_0} \ . \tag{2.3.24}$$

where, k is the Boltzman's constant, T is the temperature in K and q is the electron-charge; $V_T = kT/q$ is the thermal voltage which is 25.6 mV at room temperature. Hence, R_p and R_n can be controlled by varying the bias current I_o. In addition to this, Q and ω_0 depend on R_p and R_n, which makes them electronically adjustable.

Taking the non-idealities of CC-CDBA into account, the terminal equations can be rewritten as

$$\begin{aligned} V_p &= R_p I_p, \\ V_n &= R_n I_n, \\ I_z &= \alpha_p I_p - \alpha_n I_n, \\ V_w &= \beta V_z. \end{aligned} \tag{2.3.25}$$

where α_p, α_n and β are the current and voltage gains, respectively, and can be expressed as $\alpha_p = 1-\varepsilon_p$, $\alpha_n = 1-\varepsilon_n$, $\beta = 1-\varepsilon_v$, with $|\varepsilon_p| << 1$, $|\varepsilon_n| << 1$, $|\varepsilon_v| << 1$. ε_p and ε_n denote the current-tracking errors and ε_v denotes voltage tracking error.

2.3.9 An Implementation of OTA-based CDBA

Realization of OTA-based CDBA is proposed by Kaewpoonsunk et al. in 2005 (Kaewpoonsunk et al., 2005). This implementation is proposed due to the fact that OTAs are the low-cost and electronically tunable characteristic devices. The realization of mentioned CDBA comprises four OTAs is shown in Fig. 2.18. The relations between the input voltages and currents at *p*-terminal and *n*-terminal can be stated as

$$
\begin{aligned}
V_p &= \frac{I_p}{g_{m1}}, \\
V_n &= \frac{I_n}{g_{m2}}.
\end{aligned}
\tag{2.3.26}
$$

The transconductance gain g_{mi} of the OTA A_i is equal to $I_{Bi}/2V_T$. Where I_{Bi} and V_T are the external bias current of the OTA A_i and the thermal voltage, respectively.

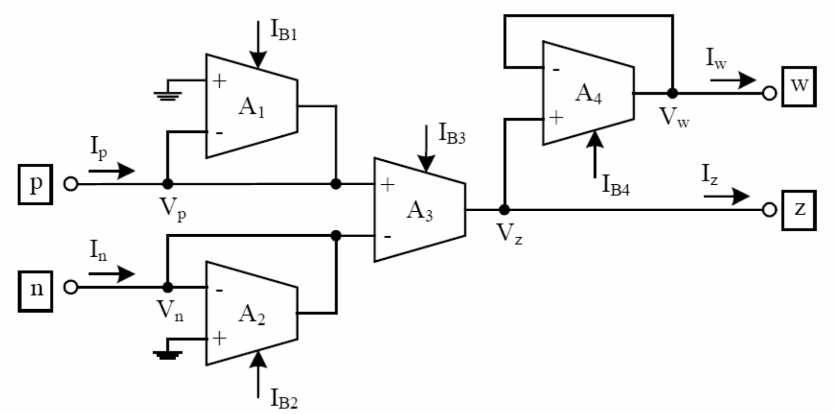

Figure 2.18 OTA-based CDBA realization.

Considering at the unity-gain feedback OTAs A_1, A_2, and A_4, if their transconductance gains are set to high values, then these OTAs will act as the impedance voltage follower (Surakampontorn et al., 1991) for the input voltages as

$$V_p = V_n = 0, \\ V_z = V_w. \tag{2.3.27}$$

Considering *z*-terminal, the current I_z can be given by

$$I_z = g_{m3}\left(V_p - V_n\right). \tag{2.3.28}$$

Substituting Eq. (2.3.26) in to (2.3.28) the output I_z can be written as

$$I_z = g_{m3}\left(\frac{I_p}{g_{m1}} - \frac{I_n}{g_{m2}}\right). \tag{2.3.29}$$

From Eq. (2.3.29), if $g_{m1} = g_{m2} = g_{m3}$ is chosen, the output current becomes as in Eq. (2.1.3).

Thus all terminal equations of CDBA are realized and an OTA-based CDBA is implemented.

2.3.10 CDBA Based Current Differencing Transconductance Amplifier (CDTA)

A new active element with two current outputs, namely current differencing transconductance amplifier (CDTA) is proposed in 2003 (Biolek, 2003). This element is a synthesis of known CCII+ (commercially available AD844) and OTA. The block diagram of CDTA element is shown in Fig. 2.19.

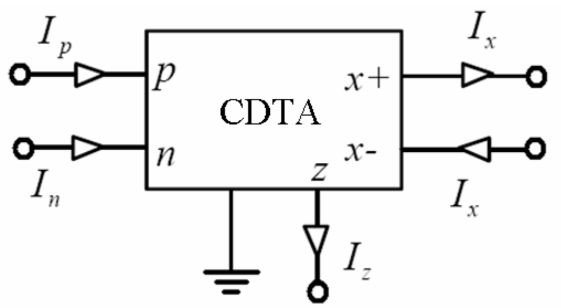

Figure 2.19 Block diagram of CDTA.

In CDTA, the difference of input currents at *p* and *n*-terminal is transferred to *z*-terminal and the voltage across the *z*-terminal is transferred by a transconductance *g* to a current that is taken out as a current pair to the *x*-terminals. The mentioned pairs are obtained by using familiar OTA. An equivalent circuit of CDTA is given in Fig. 2.20 and the pair of output currents from the *x*-terminals may have three combinations of direction; currents flow out, the currents have different directions, currents flow inside. Depending on the direction of these currents we get the CDTA++, CDTA+-, and CDTA-- elements. Current directions are signed with + and – mean outside and inside, respectively.

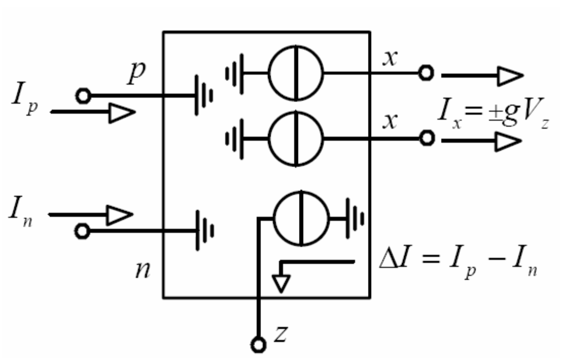

Figure 2.20 Equivalent circuit of CDTA.

A possible implementation of CDTA by using the familiar CCII+ and OTA elements is given in Fig. 2.21. Actually in this figure, a CDBA without buffer stage is realized by using CCII+ pair and an OTA is used to obtain current outputs.

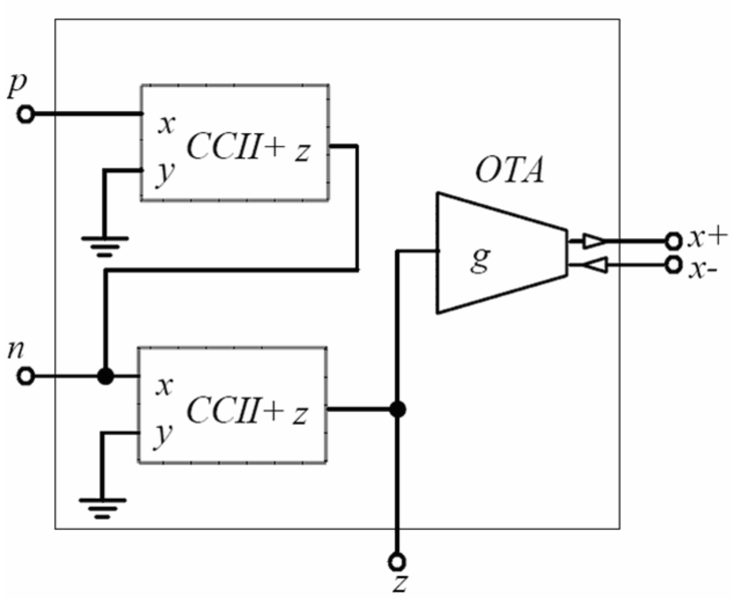

Figure 2.21 Implementation of CDTA using two CCIIs (CDBA) and one OTA.

If the voltages of *p*, *n*, *x*, and *z*-terminals are marked as V_p, V_n, V_x and V_z, respectively, we will obtain the following characteristic equation for CDTA

$$\begin{bmatrix} I_z \\ I_{x+} \\ I_{x-} \\ V_p \\ V_n \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 1 & -1 \\ g & 0 & 0 & 0 & 0 \\ -g & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} V_z \\ V_{x+} \\ V_{x-} \\ I_p \\ I_n \end{bmatrix}. \qquad (2.3.30)$$

CHAPTER 3

ELECTRONICALLY CONTROLLABLE BIQUADS USING SINGLE CDBA

In this chapter, three new electronically tunable VM second-order universal filters using single active component are proposed. Circuit configurations, analysis and sensitivity calculations are given in the second section of the chapter. PSpice simulations in agreement with the theoretical analysis are presented in the last section.

3.1 INTRODUCTION

The proposed configurations in this chapter use two capacitors, one resistor and single CC-CDBA. Q and ω_0 can be adjusted electronically without changing the values of the passive components and ω_0 and bandwidth ω_0 / Q enjoy independent electronic tunability. Each of the proposed filters realizes three basic second-order filter functions simultaneously: LP, BP, and HP. The analysis of non-ideal effects and sensitivity analyses are also investigated. The magnitude characteristics are obtained using PSpice.

The filters proposed in this chapter offer the following advantages in comparison to previously reported ones (Abuelma'atti, 2000), (Salama et al., 2001), (Minae et al., 2001), (Ozcan et al., 2002), (Acar and Sedef, 2003), (Keskin, 2005), (Kilinc and Cam, 2004):

- They use single active element.
- They use minimum number of active and passive elements.

- The Q and ω_0 are electronically controllable.
- The sensitivities are smaller than 0.5.
- The parameters ω_o and ω_o/Q are orthogonally controllable.

3.2 CIRCUIT CONFIGURATIONS AND THEIR ANALYSIS

3.2.1 LP Filter Case

The proposed circuit for the LP case is given below.

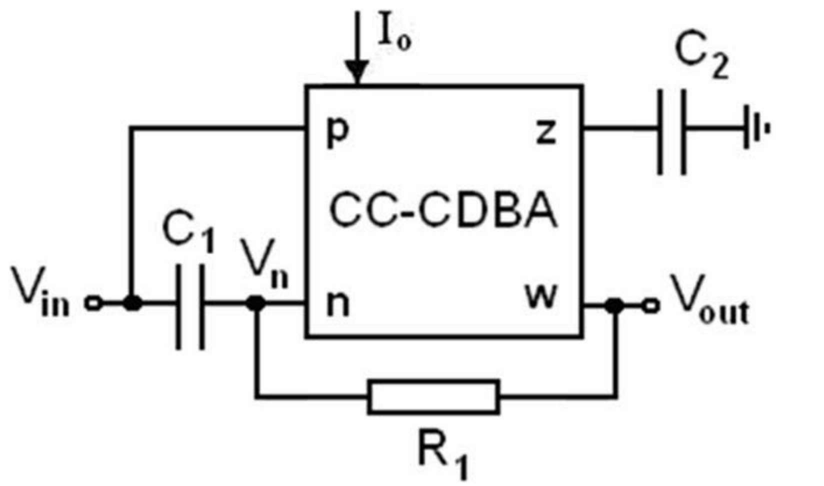

Figure 3.1 Electronically tunable VM LP filter.

For CC-CDBA given in the above figure we choose

$$R_p = R_n = R_2, \tag{3.2.1}$$

$$I_p = \frac{V_p}{R_p} = \frac{V_{in}}{R_2}, \tag{3.2.2}$$

$$I_n = \frac{V_n}{R_n} = \frac{V_n}{R_2}. \tag{3.2.3}$$

Using Eq. (3.2.3) and KCL at the *n*-terminal we obtain

$$\frac{V_n}{R_2} = sC_1\left(V_{in} - V_n\right) + \frac{V_{out} - V_n}{R_1}. \tag{3.2.4}$$

V_n is found by simplifying Eq. (3.2.4) as

$$V_n = \frac{V_{in}\,(sC_1R_1) + V_{out}}{R_1 + R_2 + sC_1R_1R_2}R_2. \tag{3.2.5}$$

By using Eq. (3.2.3) we can write I_n as

$$I_n = \frac{V_n}{R_2} = \frac{V_{in}\,(sC_1R_1) + V_{out}}{R_1 + R_2 + sC_1R_1R_2}. \tag{3.2.6}$$

Substitution I_p and I_n in Eq. (2.3.22) gives I_z as

$$I_z = \frac{V_{in}\left(R_1 + R_2 + sC_1R_1R_2 - sC_1R_1R_2\right) - V_{out}R_2}{R_2\left(R_1 + R_2 + sC_1R_1R_2\right)}. \tag{3.2.7}$$

By using Eq. (2.3.20), V_{out} is found to be

$$V_{out} = \frac{I_z}{sC_2}. \tag{3.2.8}$$

Substitution Eq. (3.2.7) into Eq. (3.2.8) gives

$$V_{out} = \frac{I_z}{sC_2} = \frac{V_{in}\left(R_1 + R_2 + sC_1R_1R_2 - sC_1R_1R_2\right) - V_{out}R_2}{sC_2R_2\left(R_1 + R_2 + sC_1R_1R_2\right)}. \tag{3.2.9}$$

Eq. (3.2.9) can be simplified as

$$\frac{V_{out}}{V_{in}} = \frac{R_1 + R_2}{s^2C_1C_2R_1R_2^2 + sC_2R_1R_2 + sC_2R_2^2 + R_2}. \tag{3.2.10}$$

This transfer function can be rewritten as

$$\frac{V_{out}}{V_{in}} = \frac{R_1 + R_2}{R_2} \frac{\frac{1}{C_1 C_2 R_1 R_2}}{s^2 + s\frac{R_1 + R_2}{C_1 R_1 R_2} + \frac{1}{C_1 C_2 R_1 R_2}}. \tag{3.2.11}$$

The parameters ω_o and ω_o/Q are obviously

$$\omega_0 = \frac{1}{\sqrt{C_1 C_2 R_1 R_2}}, \tag{3.2.12}$$

$$\frac{\omega_0}{Q} = \frac{R_1 + R_2}{R_1 C_1 R_2}. \tag{3.2.13}$$

From Eq. (3.2.12) and (3.2.13), it is seen that the parameters ω_0 and ω_0/Q are orthogonally controllable by adjusting the values of C_2.

The ideal sensitivities of the ω_0 and Q with respect to passive components are readily concluded using sensitivity analysis as in Appendix A (Sagbas, 2004) from Eqs. (3.2.12) and (3.2.13). The results are

$$\begin{aligned}
S_{R_1}^{w_0} &= S_{R_2}^{w_0} = -\frac{1}{2}, \\
S_{C_1}^{w_0} &= S_{C_2}^{w_0} = -\frac{1}{2}, \\
S_{R_1}^{Q} &= -S_{R_2}^{Q} = \frac{1}{2} - \frac{R_1}{R_1 + R_2}, \\
S_{R_1}^{Q} &= -S_{R_2}^{Q} = \frac{1}{2}. \\
S_{C_1}^{Q} &= -S_{C_2}^{Q} = \frac{1}{2}.
\end{aligned} \tag{3.2.14}$$

Obviously all sensitivities are restricted by 0.5.

3.2.2 BP Filter Case

The second proposed filter shown in Fig. 3.2 uses one CC-CDBA, two capacitors and one resistor for realizing BP filter.

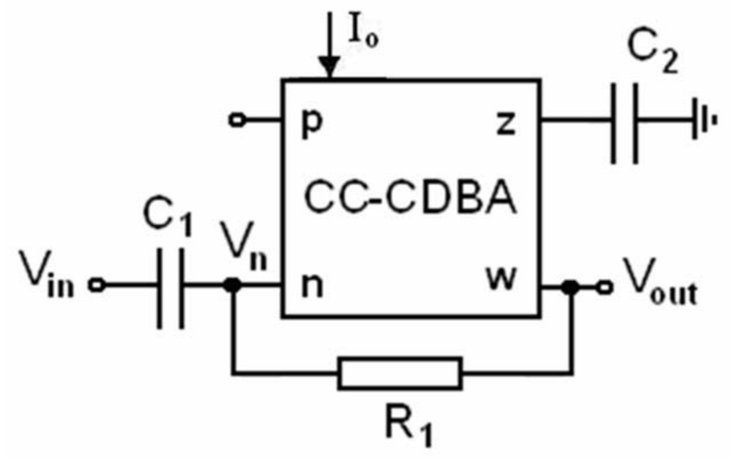

Figure 3.2 Electronically tunable VM BP filter.

For the proposed circuit input currents are

$$I_p = 0, \tag{3.2.15}$$

$$I_n = \frac{V_n}{R_n} = \frac{V_n}{R_2}. \tag{3.2.16}$$

Using Eq. (3.2.16) and KCL at the terminal *n*, we obtain

$$\frac{V_n}{R_2} = sC_1(V_{in} - V_n) + \frac{V_{out} - V_n}{R_1}. \tag{3.2.17}$$

V_n is found by simplifying Eq. (3.2.17) as

$$V_n = R_2 \frac{V_{in}(sC_1R_1) + V_{out}}{R_1 + R_2 + sC_1R_1R_2}. \tag{3.2.18}$$

From the equation Eq. (3.2.16) we can write I_n as

$$I_n = \frac{V_n}{R_2} = \frac{V_{in}\,(sC_1R_1) + V_{out}}{R_1 + R_2 + sC_1R_1R_2}. \tag{3.2.19}$$

Substitution I_p and I_n in Eq. (2.3.22) gives I_z as

$$I_z = -\frac{V_{in}\,(sC_1R_1) + V_{out}}{R_1 + R_2 + sC_1R_1R_2}. \tag{3.2.20}$$

By using Eq. (2.3.20), V_{out} is found as

$$V_{out} = \frac{I_z}{sC_2}. \tag{3.2.21}$$

Substitution Eq. (3.2.20) into Eq. (3.2.21) yields

$$V_{out} = \frac{I_z}{sC_2} = -\frac{V_{in}\,(sC_1R_1) + V_{out}}{sC_2\,(R_1 + R_2 + sC_1R_1R_2\,)}. \tag{3.2.22}$$

Eq. (3.2.22) can be simplified as

$$\frac{V_{out}}{V_{in}} = -\frac{sC_1R_1}{s^2C_1C_2R_1R_2 + sC_2R_1 + sC_2R_2 + 1}. \tag{3.2.23}$$

This voltage transfer function can be simplified as

$$\frac{V_{out}}{V_{in}} = -\frac{s/C_2R_2}{s^2 + s\dfrac{R_1 + R_2}{C_1R_1R_2} + \dfrac{1}{C_1C_2R_1R_2}}. \tag{3.2.24}$$

The parameters ω_o and ω_o/Q are obtained the same as in Eqs. (3.2.12) and (3.2.13). Therefore, the parameters ω_o and ω_o/Q are orthogonally controllable by adjusting the values of C_2, and the ideal sensitivities of ω_0 and Q with respect to passive components are the same as those given in Eq. (3.2.14) for the LP filter configuration.

3.2.3 HP Filter Case

The third proposed circuit shown in Fig. 3.3 employs one CC-CDBA, two capacitors and one resistor and produces HP response.

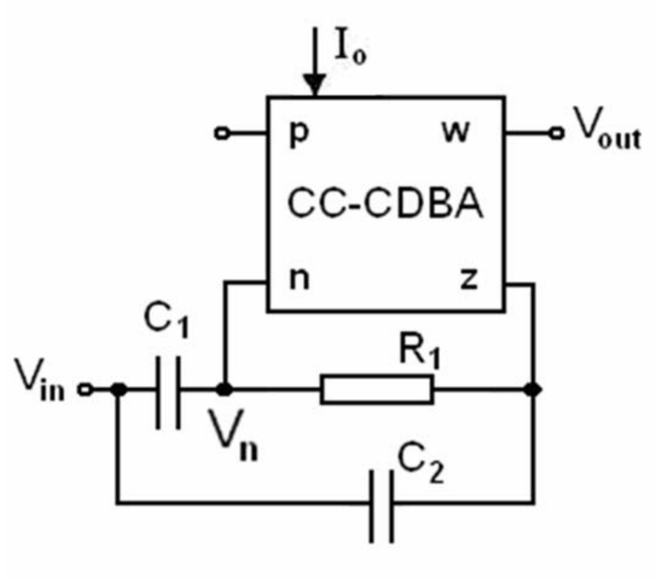

Figure 3.3 Electronically tunable VM HP filter.

Terminal input currents of the proposed circuit are

$$I_p = 0, \tag{3.2.25}$$

$$I_n = \frac{V_n}{R_n} = \frac{V_n}{R_2}. \tag{3.2.26}$$

Using Eq. (3.2.26) and KCL at the terminal n, we obtain

$$\frac{V_n}{R_2} + (V_n - V_{in})sC_1 + (V_n - V_{out})\frac{1}{R_1} = 0. \tag{3.2.27}$$

The following equality is easily found by simplifying Eq. (3.2.27) as

$$V_n(sC_1R_1R_2 + R_1 + R_2) + V_{out}(-R_2) = V_{in}(sC_1). \tag{3.2.28}$$

By using Eq. (2.3.22), we can write the current equation at the output node z as

$$-\frac{V_n}{R_2}=\left(V_{out}-V_n\right)\frac{1}{R_1}+\left(V_{out}-V_{in}\right)sC_2. \tag{3.2.29}$$

And after arrangements we obtain

$$V_n\left(R_2-R_1\right)+V_{out}\left(-sC_2R_1R_2-R_2\right)=-V_{in}\left(sC_2R_1R_2\right). \tag{3.2.30}$$

Eqs. (3.2.28) and (3.2.30) can be written in matrix form as

$$\begin{bmatrix} sC_1R_1R_2+R_1+R_2 & -R_2 \\ R_2-R_1 & -sC_2R_1R_2-R_2 \end{bmatrix}\begin{bmatrix} V_n \\ V_{out} \end{bmatrix}=\begin{bmatrix} sC_1 \\ sC_2R_1R_2 \end{bmatrix}V_{in}. \tag{3.2.31}$$

Then, the transfer function matrix is obtained as

$$\begin{bmatrix} V_n \\ V_{out} \end{bmatrix}=\frac{1}{\Delta}\begin{bmatrix} sC_1R_1R_2+R_1+R_2 & -R_2 \\ R_2-R_1 & -sC_2R_1R_2-R_2 \end{bmatrix}\begin{bmatrix} sC_1 \\ sC_2R_1R_2 \end{bmatrix}V_{in}, \tag{3.2.32}$$

where,

$$\begin{aligned}\Delta&=\left(sC_1R_1R_2+R_1+R_2\right)\left(-sC_2R_1R_2-R_2\right)+R_2\left(R_2-R_1\right)\\&=s^2R_1^2R_2^2C_1C_2+sR_1R_2\left[C_2R_1+R_2\left(C_1+C_2\right)\right]+2R_1R_2\end{aligned}. \tag{3.2.33}$$

Using Eq. (3.2.32) and Eq. (3.2.33) we obtain high-pass filter function as

$$\frac{V_{out}}{V_{in}}=\frac{s^2+s\dfrac{R_1\left(C_2-C_1\right)+R_2\left(C_1+C_2\right)}{C_1C_2R_1R_2}}{s^2+s\dfrac{R_1C_2+R_2\left(C_1+C_2\right)}{C_1C_2R_1R_2}+\dfrac{2}{C_1C_2R_1R_2}}. \tag{3.2.34}$$

Choosing the passive component value as $R_1\left(C_1-C_2\right)=R_2\left(C_1+C_2\right)$, HP filter response can be obtained. The parameters ω_o and ω_o/Q can be given as

$$\omega_0=\sqrt{\frac{2}{C_1C_2R_1R_2}}, \tag{3.2.35}$$

$$\frac{\omega_0}{Q}=\frac{R_1C_2+R_2\left(C_1+C_2\right)}{C_1C_2R_1R_2}. \tag{3.2.36}$$

The ideal sensitivities of ω_0 are the same as in the previous two cases and given in Eq. (3.2.14) by using sensitivity calculation in Appendix A. The sensitivity of the quality factor with respect to passive components are found using Eqs. (3.2.28) and (3.2.29) as

$$
\begin{aligned}
S_{R_1}^{Q} &= \frac{1}{2} - \frac{R_1 C_1}{T}, \\
S_{R_2}^{Q} &= \frac{1}{2} - \frac{R_2 (C_1 + C_2)}{T}, \\
S_{C_1}^{Q} &= \frac{1}{2} - \frac{R_2 C_1}{T}, \\
S_{C_2}^{Q} &= \frac{1}{2} - \frac{C_2 (R_1 + R_2)}{T},
\end{aligned}
\tag{3.2.37}
$$

where, $\mathrm{T} = R_1 C_2 + R_2 (C_1 + C_2)$. Again, all the sensitivities are restricted by 0.5.

3.3 SIMULATION RESULTS AND ANALYSIS

The performances of the filter topologies given in Figures 3.4-7 are verified using PSpice. Each CC-CDBA is realized by its BJT implementation shown in Fig. 2.17 with the transistor model of PR100N (PNP) and NR100N (NPN) of the bipolar arrays ALA400 from AT&T. In all of the simulations, the voltage supplies of CC-CDBA are taken as V_{cc}=2.5 V and V_{ee}=-2.5 V.

The passive components are taken as $C_1 = 0.5$ nF, $C_2 = 1$ nF, $R_1 = 2$ kΩ, and two different bias currents of the CC-CDBA are tested as 10 µA and 20 µA for LP case. The PSpice simulations with the above parameters are shown in Fig. 3.4. It can be easily seen that, the dc gain and ω_0 can be adjusted by the bias current I_0 of CC-CDBA through Eq. (2.3.24) which yields 640 (1280) Ω for $I_0 = 20(10)\mu A$ in accordance to Eqs. (3.2.11), (3.2.12), and (3.2.13), respectively. It is obvious that the simulation results deviate a lot from the theoretical ones at high frequencies after $10\,Mr/s$, which due to the parasitic capacitances of the active device.

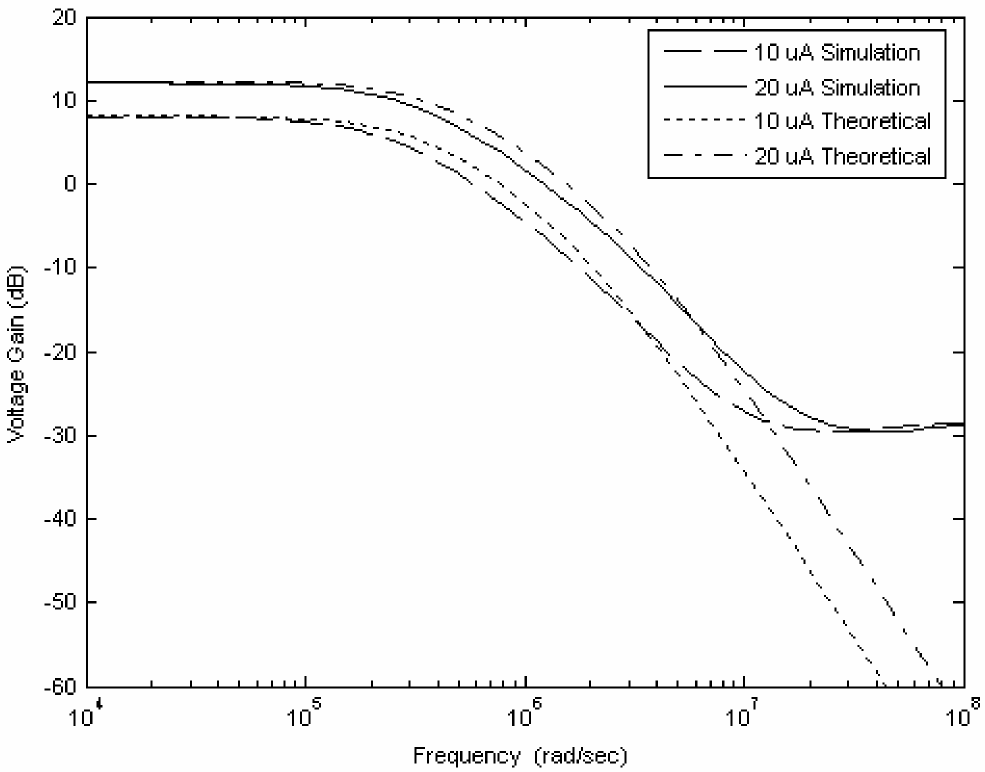

Figure 3.4 PSpice simulation results for LP case.

The performance of the BP filter topology given in Fig. 3.2 is verified using PSpice. For this simulation, the passive components are taken as $C_1 = 1$ nF, $C_2 = 0.5$ nF and $R_1 = 2\,\text{k}\Omega$. The bias current of CC-CDBA is 20 μA which yields $R_n = R_2 = 640\,\Omega$ due to Eq. (2.3.24). The PSpice simulation with the above parameters is shown in Fig. 3.5. In this simulation, the differences between the theoretical and simulated results, which are increasing with frequency and have drastic values after $100\,Mr/s$ originate from the tracking errors and parasitic capacitances of the active device (Frey, 1993).

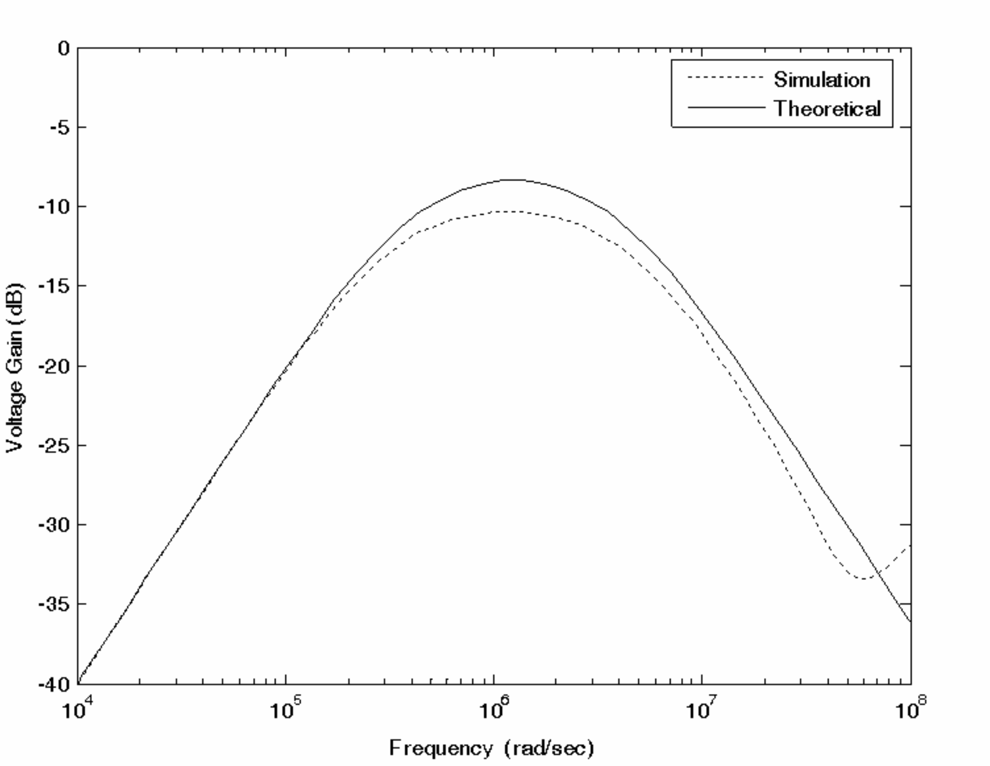

Figure 3.5 PSpice simulation results for BP case.

For the simulation of HP structure in Fig. 3.3, the bias current of CC-CDBA is chosen 20 μA and the passive components are taken as $C_1 = 1.5$ nF, $C_2 = 0.5$ nF, $R_1 = 2R_2 = 1.280$ kΩ. The PSpice simulation with the above parameters is shown in Fig. 3.6. Obviously the cancellation of the constant term in the numerator of Eq. (3.2.28) is achieved and the expected HP characteristics are obtained with 0 dB high frequency gain. Discrepancies at low frequencies are mainly due to the incomplete cancellation in the coefficient of the s-term at the numerator of Eq. (3.2.24) due to the non-ideal effects of CC-CDBA.

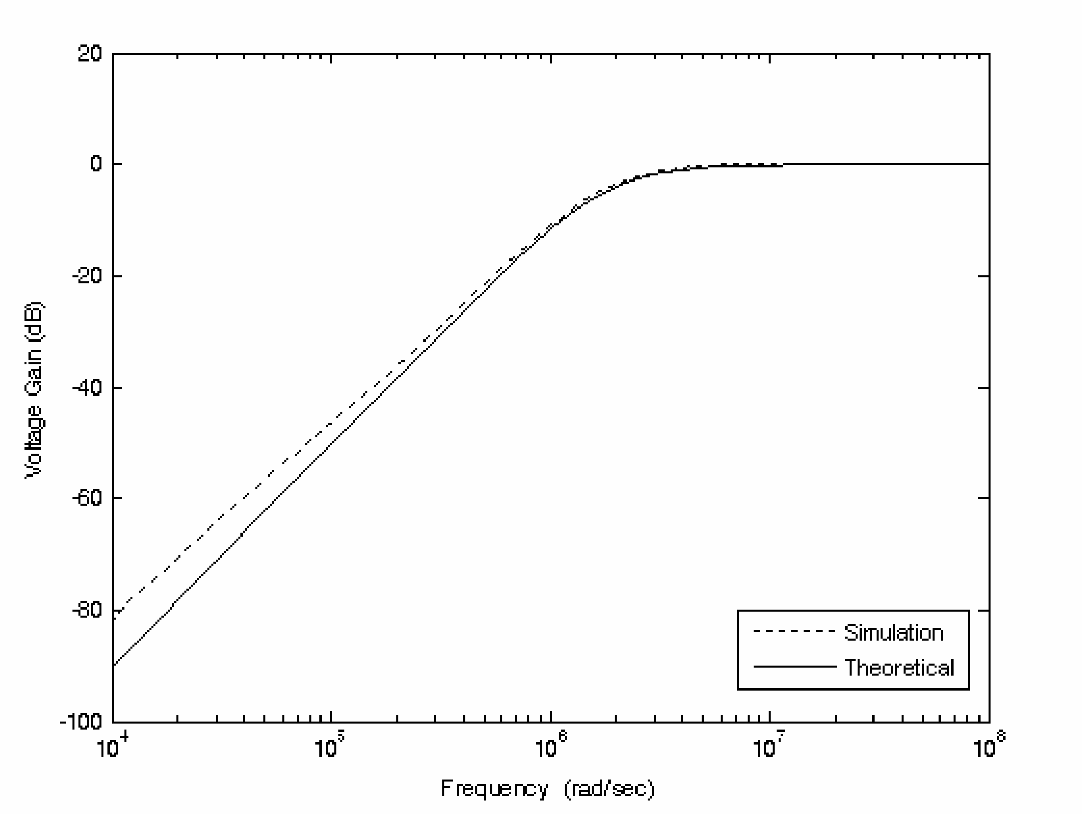

Figure 3.6 PSpice simulation results for HP case.

CHAPTER 4

A NEW SECOND-ORDER MULTI-MODE MULTI-FUNCTION FILTER USING A SINGLE CDBA

In this chapter, a new second-order multi-mode multifunction filter configuration is introduced. This configuration uses a single CDBA and four capacitors and four/five resistors depending on topology. In the sub-sections, CM, VM, TIM and TAM topologies with their non-linear analyses are presented for realizing BP, LP and HP filter responses. Third section demonstrates the sensitivity analysis of presented filters. The validity of the proposed filters is verified through PSpice simulations and experiments as a conclusion to the chapter.

4.1 INTRODUCTION

Several implementations of active filters in VM or CM using single CDBA and in multi-mode using three CDBAs have been reported in the literature (Ozoguz et al., 1999), (Kilinc and Cam, 2004), (Keskin, 2005a), (Sagbas and Koksal, 2005), (Keskin, 2006b), (Koksal and Sagbas, 2007). Many of them require many active and/or passive elements. The new circuit configuration presented here offers a new multi-mode topology using minimum number of active and passive elements in comparison to previous proposals for realizing BP, LP and HP filter responses.

4.2 CIRCUIT CONFIGURATIONS AND THEIR ANALYSES

General multi-mode filter topology is given in Fig. 4.1. This topology uses a single CDBA and five passive elements. Depending on the choices of passive elements three filter transfer functions can be obtained in voltage, current, TIM and TAM. Due to the different admittance selections for the mentioned modes, this section is divided into four sub-sections according to the usage mode to demonstrate the employability of the general topology.

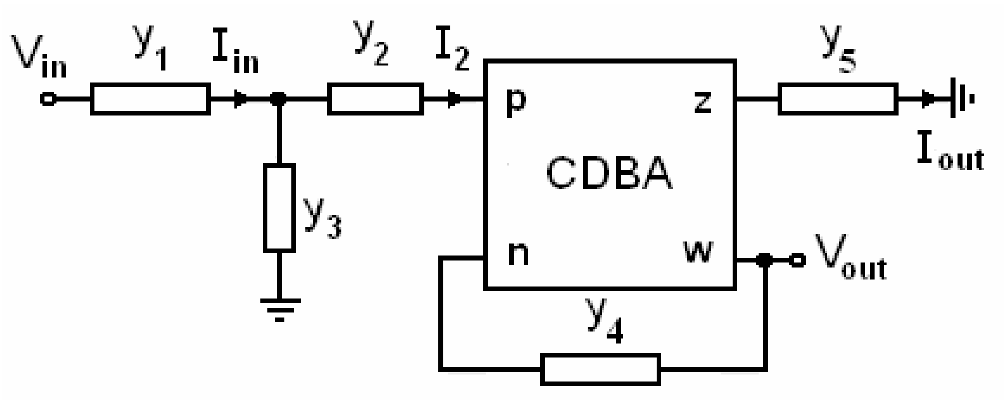

Figure 4.1 General filter topology.

4.2.1 CM Configuration

First, we analyze the configuration for the CM filter. This analysis requires a current source input and removal of admittance y_1 in Fig. 4.1.

Using ideal terminal equation in Eq. (2.1.3) for our configuration we get

$$I_z = I_{out} = I_2 - V_{out} y_4 . \tag{4.2.1}$$

Here by using current divider rule, I_2 can be found as

$$I_2 = I_{in} \frac{y_2}{y_2 + y_3} . \tag{4.2.2}$$

Also, V_{out} potential is expressed as

$$V_{out} = \frac{I_{out}}{y_5}. \tag{4.2.3}$$

By substituting Eq. (4.2.2) and Eq. (4.2.3) into Eq. (4.2.1) we obtain

$$I_{out} = I_{in} \frac{y_2}{y_2 + y_3} - \frac{I_{out}}{y_5} y_4. \tag{4.2.4}$$

After making necessary calculations, we get the following transfer function for our general filter topology;

$$\frac{I_{out}}{I_{in}} = \frac{y_2 y_5}{(y_4 + y_5)(y_2 + y_3)}. \tag{4.2.5}$$

This general filter function allows us to realize three basic filter transfer functions easily. For example, if we choose

$$y_2 = G_2,\ y_3 = sC_3,\ y_4 = sC_4,\ y_5 = G_5,$$

we will obtain the transfer function

$$\frac{I_{out}}{I_{in}} = \frac{G_2 G_5}{(sC_4 + G_5)(G_2 + sC_3)}. \tag{4.2.6}$$

By necessary arrangements, we obtain a conventional LP filter transfer function

$$H_{LP}(s) = \frac{I_{out}}{I_{in}} = \frac{\dfrac{G_2 G_5}{C_3 C_4}}{s^2 + s\dfrac{C_3 G_5 + C_4 G_2}{C_3 C_4} + \dfrac{G_2 G_5}{C_3 C_4}}. \tag{4.2.7}$$

Taking the non-ideal effects into account, the transfer function of LP filter becomes

$$H_{NI-LP}(s) = \frac{I_{out}}{I_{in}} = \frac{\dfrac{\alpha_p}{\alpha_n \beta} \dfrac{G_2 G_5}{C_3 C_4}}{s^2 + s\dfrac{\alpha_n C_4 G_2 + C_3 G_5}{\alpha_n C_3 C_4} + \dfrac{G_2 G_5}{\alpha_n C_3 C_4}}. \tag{4.2.8}$$

Ideal and non-ideal ω_0 and Q are given as

$$\omega_0 = \sqrt{\frac{G_2 G_5}{C_3 C_4}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{G_2 G_5}{\alpha_n C_3 C_4}}, \tag{4.2.9}$$

$$Q = \frac{\sqrt{C_3 C_4 G_2 G_5}}{C_4 G_2 + C_3 G_5}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_3 C_4 G_2 G_5}}{\alpha_n C_4 G_2 + C_3 G_5}. \tag{4.2.10}$$

Another topology is given with the admittance selections

$$y_2 = G_2,\ y_3 = sC_3,\ y_4 = G_4,\ y_5 = sC_5.$$

This topology results with the transfer function

$$\frac{I_{out}}{I_{in}} = \frac{sC_5 G_2}{(G_4 + sC_5)(G_2 + sC_3)}. \tag{4.2.11}$$

By making necessary calculations, we come into possession of a suitable BP filter transfer function;

$$H_{BP}(s) = \frac{I_{out}}{I_{in}} = \frac{s\dfrac{G_2}{C_3}}{s^2 + s\dfrac{C_3 G_4 + C_5 G_2}{C_3 C_5} + \dfrac{G_2 G_4}{C_3 C_5}}. \tag{4.2.12}$$

The transfer function of the proposed filter with the non-ideal effects becomes

$$H_{NI-BP}(s) = \frac{I_{out}}{I_{in}} = \frac{s\dfrac{\alpha_p G_2}{\beta C_3}}{s^2 + s\dfrac{\alpha_n C_3 G_4 + C_5 G_2}{C_3 C_5} + \dfrac{\alpha_n G_2 G_4}{C_3 C_5}}. \tag{4.2.13}$$

Ideal and non-ideal ω_0 and Q for the above BP filter functions are given as

$$\omega_0 = \sqrt{\frac{G_2 G_4}{C_3 C_5}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_2 G_4}{C_3 C_5}}, \tag{4.2.14}$$

$$Q = \frac{\sqrt{C_3 C_5 G_2 G_4}}{C_3 G_4 + C_5 G_2}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_3 C_5 G_2 G_4}}{\alpha_n C_3 G_4 + C_5 G_2}. \tag{4.2.15}$$

The last configuration is obtained by choosing the following admittances;

$$y_2 = sC_2,\; y_3 = G_3,\; y_4 = G_4,\; y_5 = sC_5.$$

If we insert these admittances into Eq. (4.2.5), we obtain

$$\frac{I_{out}}{I_{in}} = \frac{s^2 C_2 C_5}{(G_4 + sC_5)(sC_2 + G_3)}. \tag{4.2.16}$$

This transfer function can easily be simplified and the following HP filter transfer function equation is obtained;

$$H_{HP}(s) = \frac{I_{out}}{I_{in}} = \frac{s^2}{s^2 + s\dfrac{C_2 G_4 + C_5 G_3}{C_2 C_5} + \dfrac{G_3 G_4}{C_2 C_5}}. \tag{4.2.17}$$

The non-ideal HP transfer function is expressed as

$$H_{NI-HP}(s) = \frac{I_{out}}{I_{in}} = \frac{s^2 \dfrac{\alpha_p}{\beta}}{s^2 + s\dfrac{\alpha_n C_2 G_4 + C_5 G_3}{C_2 C_5} + \dfrac{\alpha_n G_3 G_4}{C_2 C_5}}. \tag{4.2.18}$$

Ideal and non-ideal ω_0 and Q are given as

$$\omega_0 = \sqrt{\frac{G_3 G_4}{C_2 C_5}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_3 G_4}{C_2 C_5}}, \tag{4.2.19}$$

$$Q = \frac{\sqrt{C_2 C_5 G_3 G_4}}{C_2 G_4 + C_5 G_3}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_2 C_5 G_3 G_4}}{\alpha_n C_2 G_4 + C_5 G_3}. \tag{4.2.20}$$

4.2.2 VM Configuration

For a voltage input case, in which admittance y_1 is included in Fig. 4.1, the new general transfer function can be obtained by the following steps:

Output current is expressed by using Eq. (2.1.3) as

$$I_z = I_{out} = I_2 - V_{out} y_4 . \tag{4.2.21}$$

The *p*-terminal input I_2 current can be written as

$$I_2 = I_{in} \frac{y_2}{y_2 + y_3} . \tag{4.2.22}$$

In this equality input current I_{in} can be expressed as

$$I_{in} = V_{in} \frac{y_1 (y_2 + y_3)}{y_1 + y_2 + y_3} , \tag{4.2.23}$$

and substituting Eq. (4.2.23) into Eq. (4.2.22) we obtain

$$I_2 = V_{in} \frac{y_1 y_2}{y_1 + y_2 + y_3} . \tag{4.2.24}$$

The *z*-terminal output voltage and current relation is given as follows;

$$I_z = V_{out} y_5 . \tag{4.2.25}$$

By using Eq. (4.2.24) and (4.2.25) we obtain

$$V_{out} y_5 = V_{in} \frac{y_1 y_2}{y_1 + y_2 + y_3} - V_{out} y_4 . \tag{4.2.26}$$

After re-arrangement in Eq. (4.2.26), we are able to obtain our general voltage transfer function

$$\frac{V_{out}}{V_{in}}=\frac{y_1 y_2}{(y_1+y_2+y_3)(y_4+y_5)}. \tag{4.2.27}$$

All three basic voltage mode filter functions, namely LP, BP and high-pass can be obtained by selecting suitable admittances.

The first choice is

$$y_1=G_1,\ y_2=G_2,\ y_3=sC_3,\ y_4=G_4,\ y_5=sC_5.$$

Then by using Eq. (4.2.27) we obtain

$$\frac{V_{out}}{V_{in}}=\frac{G_1 G_2}{(G_1+G_2+sC_3)(G_4+sC_5)}. \tag{4.2.28}$$

And re-organizing Eq. (4.2.28) we arrive a LP filter characteristic equation

$$H_{LP}(s)=\frac{V_{out}}{V_{in}}=\frac{\dfrac{G_1G_2}{C_3C_5}}{s^2+s\dfrac{C_3G_4+C_5G_1+C_5G_2}{C_3C_5}+\dfrac{G_4(G_1+G_2)}{C_3C_5}}. \tag{4.2.29}$$

The transfer function of the proposed filter taking the non-ideal effects into account becomes

$$H_{NI-LP}(s)=\frac{V_{out}}{V_{in}}=\frac{\dfrac{\alpha_p G_1G_2}{\beta C_3C_5}}{s^2+s\dfrac{\alpha_n C_3G_4+C_5G_1+C_5G_2}{C_3C_5}+\dfrac{\alpha_n G_4(G_1+G_2)}{C_3C_5}}. \tag{4.2.30}$$

Ideal and non-ideal ω_0 and Q for LP transfer functions are obtained as

$$\omega_0=\sqrt{\frac{G_4(G_1+G_2)}{C_3C_5}},\qquad \omega_{0_{NI}}=\sqrt{\frac{\alpha_n G_4(G_1+G_2)}{C_3C_5}}, \tag{4.2.31}$$

$$Q=\frac{\sqrt{C_3C_5G_4(G_1+G_2)}}{C_3G_4+C_5G_1+C_5G_2},\qquad Q_{NI}=\frac{\sqrt{\alpha_n C_3C_5G_4(G_1+G_2)}}{\alpha_n C_3G_4+C_5G_1+C_5G_2}. \tag{4.2.32}$$

Another topology can be presented by the selection

$$y_1 = sC_1,\ y_2 = G_2,\ y_3 = sC_3,\ y_4 = G_4,\ y_5 = sC_5.$$

Again using Eq. (4.2.27) we get

$$\frac{V_{out}}{V_{in}} = \frac{sC_1G_2}{\left(sC_1 + G_2 + sC_3\right)\left(G_4 + sC_5\right)}. \tag{4.2.33}$$

And with the final arrangement we obtain a BP filter transfer function

$$H_{BP}(s) = \frac{V_{out}}{V_{in}} = \frac{\dfrac{sC_1G_2}{C_5\left(C_1 + C_3\right)}}{s^2 + s\dfrac{C_1G_4 + C_3G_4 + C_5G_2}{C_5\left(C_1 + C_3\right)} + \dfrac{G_2G_4}{C_5\left(C_1 + C_3\right)}}. \tag{4.2.34}$$

The transfer function of the proposed filter taking the non-ideal effects into account becomes

$$H_{NI-BP}(s) = \frac{V_{out}}{V_{in}} = \frac{s\dfrac{\alpha_p C_1G_2}{\beta C_5\left(C_1 + C_3\right)}}{s^2 + s\dfrac{\alpha_n C_1G_4 + \alpha_n C_3G_4 + C_5G_2}{C_5\left(C_1 + C_3\right)} + \dfrac{\alpha_n G_2G_4}{C_5\left(C_1 + C_3\right)}}. \tag{4.2.35}$$

Ideal and non-ideal ω_0 and Q are given as

$$\omega_0 = \sqrt{\frac{G_2G_4}{C_5\left(C_1 + C_3\right)}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_2G_4}{C_5\left(C_1 + C_3\right)}}, \tag{4.2.36}$$

$$Q = \frac{\sqrt{C_5G_2G_4\left(C_1 + C_3\right)}}{C_1G_4 + C_3G_4 + C_5G_2}, \quad Q_{NI} = \frac{\sqrt{\alpha_n C_5G_2G_4\left(C_1 + C_3\right)}}{\alpha_n C_1G_4 + \alpha_n C_3G_4 + C_5G_2}. \tag{4.2.37}$$

The final topology for VM is realized by selecting

$$y_1 = sC_1,\ y_2 = sC_2,\ y_3 = G_3,\ y_4 = G_4,\ y_5 = sC_5.$$

General transfer function in Eq. (4.2.27) becomes

$$\frac{V_{out}}{V_{in}} = \frac{s^2 C_1 C_2}{(sC_1 + sC_2 + G_3)(G_4 + sC_5)}. \tag{4.2.38}$$

Arrangement on Eq. (4.2.38) gives HP filter VM transfer function in the form

$$H_{HP}(s) = \frac{V_{out}}{V_{in}} = \frac{\dfrac{s^2 C_1 C_2}{C_5(C_1 + C_2)}}{s^2 + s\dfrac{C_1 G_4 + C_2 G_4 + C_5 G_3}{C_5(C_1 + C_2)} + \dfrac{G_3 G_4}{C_5(C_1 + C_2)}}. \tag{4.2.39}$$

The transfer function of HP filter with non-ideal effects becomes

$$H_{NI-HP}(s) = \frac{V_{out}}{V_{in}} = \frac{s^2 \dfrac{\alpha_p C_1 C_2}{\beta C_5(C_1 + C_2)}}{s^2 + s\dfrac{\alpha_n C_1 G_4 + \alpha_n C_2 G_4 + C_5 G_3}{C_5(C_1 + C_2)} + \dfrac{\alpha_n G_3 G_4}{C_5(C_1 + C_2)}}. \tag{4.2.40}$$

Ideal and non-ideal ω_0 and Q are given as

$$\omega_0 = \sqrt{\frac{G_3 G_4}{C_5(C_1 + C_2)}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_3 G_4}{C_5(C_1 + C_2)}}, \tag{4.2.41}$$

$$Q = \frac{\sqrt{C_5 G_3 G_4 (C_1 + C_2)}}{C_1 G_4 + C_2 G_4 + C_5 G_3}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_5 G_3 G_4 (C_1 + C_2)}}{\alpha_n C_1 G_4 + \alpha_n C_2 G_4 + C_5 G_3}. \tag{4.2.42}$$

4.2.3 TIM Configuration

The derivation of TIM filter transfer function requires removal of admittance y_1 in Fig. 4.1. By using Eq. (4.2.3) and Eq. (4.2.5), one can derive the following TIM transfer function

$$\frac{V_{out}}{I_{in}} = \frac{y_2}{(y_4 + y_5)(y_2 + y_3)}. \tag{4.2.43}$$

Eq. (4.2.43) can be arranged by the following admittances.

$$y_2 = G_2,\ y_3 = sC_3,\ y_4 = G_4,\ y_5 = sC_5.$$

The resulting transfer function is given as

$$\frac{V_{out}}{I_{in}} = \frac{G_2}{(G_4 + sC_5)(G_2 + sC_3)}. \tag{4.2.44}$$

This equation can be simplified and a LP filter characteristic is derived as

$$H_{LP}(s) = \frac{V_{out}}{I_{in}} = \frac{\frac{G_2}{C_3C_5}}{s^2 + s\frac{C_3G_4 + C_5G_2}{C_3C_5} + \frac{G_2G_4}{C_3C_5}}. \tag{4.2.45}$$

The transfer function of the proposed filter taking the non-ideal effects into account becomes

$$H_{NI-LP}(s) = \frac{V_{out}}{I_{in}} = \frac{\frac{\alpha_p G_2}{\beta C_3C_5}}{s^2 + s\frac{\alpha_n C_3G_4 + C_5G_2}{C_3C_5} + \frac{\alpha_n G_2G_4}{C_3C_5}}. \tag{4.2.46}$$

Ideal and non-ideal ω_0 and Q are given by

$$\omega_0 = \sqrt{\frac{G_2G_4}{C_3C_5}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_2G_4}{C_3C_5}}, \tag{4.2.47}$$

$$Q = \frac{\sqrt{C_3C_5G_2G_4}}{C_3G_4 + C_5G_2}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_3C_5G_2G_4}}{\alpha_n C_3G_4 + C_5G_2}. \tag{4.2.48}$$

Another TIM topology is obtained by the following admittance choices;

$$y_2 = sC_2,\ y_3 = G_3,\ y_4 = G_4,\ y_5 = sC_5.$$

We obtain the following filter transfer function

$$\frac{V_{out}}{I_{in}} = \frac{sC_2}{(G_4 + sC_5)(sC_2 + G_3)}. \tag{4.2.49}$$

If we multiply the factors in denominator and make necessary arrangements, we have a BP filter transfer function in the form

$$H_{BP}(s) = \frac{V_{out}}{I_{in}} = \frac{\dfrac{s}{C_5}}{s^2 + s\dfrac{C_2G_4 + C_5G_3}{C_2C_5} + \dfrac{G_3G_4}{C_2C_5}}. \tag{4.2.50}$$

The transfer function of the proposed filter with the non-ideal effects becomes;

$$H_{BP}(s) = \frac{V_{out}}{I_{in}} = \frac{s\dfrac{\alpha_p}{\beta C_5}}{s^2 + s\dfrac{\alpha_n C_2G_4 + C_5G_3}{C_2C_5} + \dfrac{\alpha_n G_3G_4}{C_2C_5}}. \tag{4.2.51}$$

Ideal and non-ideal ω_0 and Q are obtained as

$$\omega_0 = \sqrt{\frac{G_3G_4}{C_2C_5}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_3G_4}{C_2C_5}}, \tag{4.2.52}$$

$$Q = \frac{\sqrt{C_2C_5G_3G_4}}{C_2G_4 + C_5G_3}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_2C_5G_3G_4}}{\alpha_n C_2G_4 + C_5G_3}. \tag{4.2.53}$$

4.2.4 TAM Configuration

General transfer function for a TAM filter can be obtained by using Eq. (4.2.3) and Eq. (4.2.27) in which admittance y_1 is comprised in Fig. 4.1. Then,

$$\frac{I_{out}}{V_{in}} = \frac{y_1y_2y_5}{(y_1 + y_2 + y_3)(y_4 + y_5)}. \tag{4.2.54}$$

The first configuration for this transfer function is obtained by choosing

$$y_1 = G_1,\ y_2 = G_2,\ y_3 = sC_3,\ y_4 = sC_4,\ y_5 = G_5.$$

If we locate these admittances into Eq. (4.2.54) we obtain

$$\frac{I_{out}}{V_{in}} = \frac{G_1 G_2 G_5}{(G_1 + G_2 + sC_3)(sC_4 + G_5)}. \tag{4.2.55}$$

After making necessary calculations, we come into possession of a suitable BP filter transfer function in the form

$$H_{LP}(s) = \frac{I_{out}}{V_{in}} = \frac{\dfrac{G_1 G_2 G_5}{C_3 C_4}}{s^2 + s\dfrac{C_4 G_1 + C_4 G_2 + C_3 G_5}{C_3 C_4} + \dfrac{G_5(G_1 + G_2)}{C_3 C_4}}. \tag{4.2.56}$$

The non-ideal transfer function of the proposed filter becomes

$$H_{NI-LP}(s) = \frac{I_{out}}{V_{in}} = \frac{\dfrac{\alpha_p}{\alpha_n \beta} \dfrac{G_1 G_2 G_5}{C_3 C_4}}{s^2 + s\dfrac{\alpha_n C_4 G_1 + \alpha_n C_4 G_2 + C_3 G_5}{\alpha_n C_3 C_4} + \dfrac{G_5(G_1 + G_2)}{\alpha_n C_3 C_4}}. \tag{4.2.57}$$

Ideal and non-ideal ω_0 and Q for this filter are expressed as

$$\omega_0 = \sqrt{\frac{G_5(G_1 + G_2)}{C_3 C_4}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{G_5(G_1 + G_2)}{\alpha_n C_3 C_4}}, \tag{4.2.58}$$

$$Q = \frac{\sqrt{C_3 C_4 G_5 (G_1 + G_2)}}{C_3 G_5 + C_5 G_1 + C_5 G_2}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_3 C_4 G_5 (G_1 + G_2)}}{\alpha_n C_4 G_1 + \alpha_n C_4 G_2 + C_3 G_5}. \tag{4.2.59}$$

If passive elements are chosen as

$$y_1 = G_1,\ y_2 = G_2,\ y_3 = sC_3,\ y_4 = G_4,\ y_5 = sC_5,$$

the following transfer function is obtained.

$$\frac{V_{out}}{V_{in}} = \frac{sC_5 G_1 G_2}{(G_1 + G_2 + sC_3)(G_4 + sC_5)}. \tag{4.2.60}$$

And re-organization of Eq. (4.2.60) will result in a BP filter characteristic equation in the form

$$H_{BP}(s) = \frac{I_{out}}{V_{in}} = \frac{\frac{sG_1G_2}{C_3}}{s^2 + s\frac{C_3G_4 + C_5G_1 + C_5G_2}{C_3C_5} + \frac{G_4(G_1 + G_2)}{C_3C_5}}. \tag{4.2.61}$$

By taking into account the non-ideal effects of CDBA, modified transfer function becomes

$$H_{NI-BP}(s) = \frac{I_{out}}{V_{in}} = \frac{s\frac{\alpha_p G_1G_2}{\beta C_3}}{s^2 + s\frac{\alpha_n C_3G_4 + C_5G_1 + C_5G_2}{C_3C_5} + \frac{\alpha_n G_4(G_1 + G_2)}{C_3C_5}}. \tag{4.2.62}$$

Ideal and non-ideal ω_0 and Q are given as follows;

$$\omega_0 = \sqrt{\frac{G_4(G_1 + G_2)}{C_3C_5}}, \qquad \omega_{0_{NI}} = \sqrt{\frac{\alpha_n G_4(G_1 + G_2)}{C_3C_5}}, \tag{4.2.63}$$

$$Q = \frac{\sqrt{C_3C_5G_4(G_1 + G_2)}}{C_3G_4 + C_5G_1 + C_5G_2}, \qquad Q_{NI} = \frac{\sqrt{\alpha_n C_3C_5G_4(G_1 + G_2)}}{\alpha_n C_3G_4 + C_5G_1 + C_5G_2}. \tag{4.2.64}$$

Following admittances will give our last topology;

$$y_1 = sC_1,\ y_2 = G_2,\ y_3 = G_3,\ y_4 = G_4,\ y_5 = sC_5.$$

With these selections, the transfer function in Eq. (4.2.54) becomes

$$\frac{V_{out}}{V_{in}} = \frac{s^2C_1C_5G_2}{(sC_1 + G_2 + G_3)(G_4 + sC_5)}. \tag{4.2.65}$$

And with the final arrangement we obtain a HP filter;

$$H_{HP}(s) = \frac{I_{out}}{V_{in}} = \frac{s^2G_2}{s^2 + s\frac{C_1G_4 + C_5G_2 + C_5G_3}{C_1C_5} + \frac{G_4(G_2 + G_3)}{C_1C_5}}. \tag{4.2.66}$$

The transfer function of the proposed filter taking the non-ideal effects into account becomes

$$H_{NI-HP}(s)=\frac{I_{out}}{V_{in}}=\frac{s^2\,\frac{\alpha_p G_2}{\beta}}{s^2+s\,\frac{\alpha_n C_1 G_4+C_5 G_2+C_5 G_3}{C_1 C_5}+\frac{\alpha_n G_4(G_2+G_3)}{C_1 C_5}}. \tag{4.2.67}$$

Ideal and non-ideal ω_0 and Q are obtained as

$$\omega_0=\sqrt{\frac{G_4(G_2+G_3)}{C_1 C_5}}, \qquad \omega_{0_{NI}}=\sqrt{\frac{\alpha_n G_4(G_2+G_3)}{C_1 C_5}}, \tag{4.2.68}$$

$$Q=\frac{\sqrt{C_1 C_5 G_4(G_2+G_3)}}{C_1 G_4+C_5 G_2+C_5 G_3}, \quad Q_{NI}=\frac{\sqrt{\alpha_n C_1 C_5 G_4(G_2+G_3)}}{\alpha_n C_1 G_4+C_5 G_2+C_5 G_3}. \tag{4.2.69}$$

Up to this point, filter characteristics depending on mode of operation with various admittance selections are given. A summary of the presented filter functions with these admittance selections in CM-VM and TIM-TAM are given in Table 4.1 and Table 4.2, respectively.

Table 4.1 CM and VM filters with appropriate admittances.

Mode	Filter	Admittance	Transfer Function
CM	LP	$y_2 = G_2$, $y_3 = sC_3$, $y_4 = sC_4$, $y_5 = G_5$.	$H_{LP}(s) = \frac{I_{out}}{I_{in}} = \frac{\frac{G_2G_5}{C_3C_4}}{s^2 + s\frac{C_3G_5 + C_4G_2}{C_3C_4} + \frac{G_2G_5}{C_3C_4}}$
	BP	$y_2 = G_2$, $y_3 = sC_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{BP}(s) = \frac{I_{out}}{I_{in}} = \frac{s\frac{G_2}{C_3}}{s^2 + s\frac{C_3G_4 + C_5G_2}{C_3C_5} + \frac{G_2G_4}{C_3C_5}}$
	HP	$y_2 = sC_2$, $y_3 = G_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{HP}(s) = \frac{I_{out}}{I_{in}} = \frac{s^2}{s^2 + s\frac{C_2G_4 + C_5G_3}{C_2C_5} + \frac{G_3G_4}{C_2C_5}}$
VM	LP	$y_1 = G_1$, $y_2 = G_2$, $y_3 = sC_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{LP}(s) = \frac{V_{out}}{V_{in}} = \frac{\frac{G_1G_2}{C_3C_5}}{s^2 + s\frac{C_3G_4 + C_5G_1 + C_5G_2}{C_3C_5} + \frac{G_4(G_1 + G_2)}{C_3C_5}}$
	BP	$y_1 = sC_1$, $y_2 = G_2$, $y_3 = sC_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{BP}(s) = \frac{V_{out}}{V_{in}} = \frac{\frac{sC_1G_2}{C_5(C_1 + C_3)}}{s^2 + s\frac{C_1G_4 + C_3G_4 + C_5G_2}{C_5(C_1 + C_3)} + \frac{G_2G_4}{C_5(C_1 + C_3)}}$
	HP	$y_1 = sC_1$, $y_2 = sC_2$, $y_3 = G_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{HP}(s) = \frac{V_{out}}{V_{in}} = \frac{\frac{s^2C_1C_2}{C_5(C_1 + C_2)}}{s^2 + s\frac{C_1G_4 + C_2G_4 + C_5G_3}{C_5(C_1 + C_2)} + \frac{G_3G_4}{C_5(C_1 + C_2)}}$

Table 4.2 TIM and TAM filters with appropriate admittances.

Mode	Filter	Admittance	Transfer Function
TIM	LP	$y_2 = G_2$, $y_3 = sC_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{LP}(s) = \dfrac{V_{out}}{I_{in}} = \dfrac{\dfrac{G_2}{C_3C_5}}{s^2 + s\dfrac{C_3G_4 + C_5G_2}{C_3C_5} + \dfrac{G_2G_4}{C_3C_5}}$
	BP	$y_2 = sC_2$, $y_3 = G_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{BP}(s) = \dfrac{V_{out}}{I_{in}} = \dfrac{\dfrac{s}{C_5}}{s^2 + s\dfrac{C_2G_4 + C_5G_3}{C_2C_5} + \dfrac{G_3G_4}{C_2C_5}}$
	HP	Not Applicable	Not Applicable
TAM	LP	$y_1 = G_1$, $y_2 = G_2$, $y_3 = sC_3$, $y_4 = sC_4$, $y_5 = G_5$.	$H_{LP}(s) = \dfrac{I_{out}}{V_{in}} = \dfrac{\dfrac{G_1G_2G_5}{C_3C_4}}{s^2 + s\dfrac{C_4G_1 + C_4G_2 + C_3G_5}{C_3C_4} + \dfrac{G_5(G_1 + G_2)}{C_3C_4}}$
	BP	$y_1 = G_1$, $y_2 = G_2$, $y_3 = sC_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{BP}(s) = \dfrac{I_{out}}{V_{in}} = \dfrac{\dfrac{sG_1G_2}{C_3}}{s^2 + s\dfrac{C_3G_4 + C_5G_1 + C_5G_2}{C_3C_5} + \dfrac{G_4(G_1 + G_2)}{C_3C_5}}$
	HP	$y_1 = sC_1$, $y_2 = G_2$, $y_3 = G_3$, $y_4 = G_4$, $y_5 = sC_5$.	$H_{HP}(s) = \dfrac{I_{out}}{V_{in}} = \dfrac{s^2G_2}{s^2 + s\dfrac{C_1G_4 + C_5G_2 + C_5G_3}{C_1C_5} + \dfrac{G_4(G_2 + G_3)}{C_1C_5}}$

4.3 SENSITIVITY ANALYSIS

The ideal and non-ideal sensitivities of ω_0 and Q with respect to passive components are calculated using the formulation shown in Appendix A (Sagbas, 2004).

All of the realized filter functions are free from non-ideal voltage gain constant β at w-terminal and non-ideal current gain constant α_p at p-terminal. Thus the sensitivities with respect to α_p and β are equal to zero. Other sensitivities are obtained as follows:

For CM LP case, by using Eq. (4.2.9) and (4.2.10) we obtain the following sensitivities;

$$S_{G_2}^{\omega} = S_{G_5}^{\omega} = -S_{C_3}^{\omega} = -S_{C_4}^{\omega} = \frac{1}{2}, \tag{4.3.1}$$

$$S_{G_2}^{\omega_{NI}} = S_{G_5}^{\omega_{NI}} = -S_{C_3}^{\omega_{NI}} = -S_{C_4}^{\omega_{NI}} = -S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.2}$$

$$S_{G_2}^{Q} = S_{C_4}^{Q} = \frac{1}{2} - \frac{C_4 G_2}{C_3 G_5 + C_4 G_2}, \tag{4.3.3}$$

$$S_{G_2}^{Q_{NI}} = S_{C_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n C_4 G_2}{C_3 G_5 + \alpha_n C_4 G_2}, \tag{4.3.4}$$

$$S_{C_3}^{Q} = S_{G_5}^{Q} = \frac{1}{2} - \frac{C_3 G_5}{C_3 G_5 + C_4 G_2}, \tag{4.3.5}$$

$$S_{C_3}^{Q_{NI}} = S_{G_5}^{Q_{NI}} = \frac{1}{2} - \frac{C_3 G_5}{C_3 G_5 + \alpha_n C_4 G_2}. \tag{4.3.6}$$

From the above calculations, it is deduced that all ideal and non-ideal sensitivities are equal or smaller than 0.5 in magnitude.

Due to the same ω_0 and Q, sensitivities for CM BP and TIM LP filters are same as in the following:

$$S_{G_2}^{\omega} = S_{G_4}^{\omega} = -S_{C_3}^{\omega} = -S_{C_5}^{\omega} = \frac{1}{2}, \tag{4.3.13}$$

$$S_{G_2}^{\omega_{NI}} = S_{G_4}^{\omega_{NI}} = -S_{C_3}^{\omega_{NI}} = -S_{C_5}^{\omega_{NI}} = S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.14}$$

$$S_{C_3}^{Q} = S_{G_4}^{Q} = \frac{1}{2} - \frac{C_3 G_4}{C_3 G_4 + C_5 G_2}, \tag{4.3.15}$$

$$S_{C_3}^{Q_{NI}} = S_{G_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n C_3 G_4}{\alpha_n C_3 G_4 + C_5 G_2}, \tag{4.3.16}$$

$$S_{C_5}^{Q} = S_{G_2}^{Q} = \frac{1}{2} - \frac{C_5 G_2}{C_3 G_4 + C_5 G_2}, \tag{4.3.17}$$

$$S_{C_5}^{Q_{NI}} = S_{G_2}^{Q_{NI}} = \frac{1}{2} - \frac{C_5 G_2}{\alpha_n C_3 G_4 + C_5 G_2}. \tag{4.3.18}$$

Note that all ideal and non-ideal sensitivities are bounded by ∓ 0.5.

The above similarity occurs for CM HP and TIM BP filters. Thus we obtain following sensitivities;

$$S_{G_3}^{\omega} = S_{G_4}^{\omega} = -S_{C_2}^{\omega} = -S_{C_5}^{\omega} = \frac{1}{2}, \tag{4.3.19}$$

$$S_{G_3}^{\omega_{NI}} = S_{G_4}^{\omega_{NI}} = -S_{C_2}^{\omega_{NI}} = -S_{C_5}^{\omega_{NI}} = S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.20}$$

$$S_{C_2}^{Q} = S_{G_4}^{Q} = \frac{1}{2} - \frac{C_2 G_4}{C_2 G_4 + C_5 G_3}, \tag{4.3.21}$$

$$S_{C_2}^{Q_{NI}} = S_{G_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n C_2 G_4}{\alpha_n C_2 G_4 + C_5 G_3}, \tag{4.3.22}$$

$$S_{C_5}^{Q} = S_{G_3}^{Q} = \frac{1}{2} - \frac{C_5 G_3}{C_2 G_4 + C_5 G_3}, \tag{4.3.23}$$

$$S_{C_5}^{Q_{NI}} = S_{G_3}^{Q_{NI}} = \frac{1}{2} - \frac{C_5 G_3}{\alpha_n C_2 G_4 + C_5 G_3}. \tag{4.3.24}$$

Again, the above sensitivities are equal or smaller than 0.5 in absolute value.

For VM LP and TAM BP filter functions we obtain the following sensitivities;

$$S_{G_1}^{\omega} = S_{G_1}^{\omega_{NI}} = \frac{1}{2}\frac{G_1}{G_1 + G_2}, \tag{4.3.25}$$

$$S_{G_2}^{\omega} = S_{G_2}^{\omega_{NI}} = \frac{1}{2}\frac{G_2}{G_1 + G_2}, \tag{4.3.26}$$

$$S_{G_4}^{\omega} = S_{G_4}^{\omega_{NI}} = -S_{C_3}^{\omega} = -S_{C_3}^{\omega_{NI}} = -S_{C_5}^{\omega} = -S_{C_5}^{\omega_{NI}} = S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.27}$$

$$S_{G_1}^{Q} = \frac{1}{2}\frac{G_1}{G_1 + G_2} - \frac{C_5 G_1}{C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.28}$$

$$S_{G_1}^{Q_{NI}} = \frac{1}{2}\frac{G_1}{G_1 + G_2} - \frac{C_5 G_1}{\alpha_n C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.29}$$

$$S_{G_2}^{Q} = \frac{1}{2}\frac{G_2}{G_1 + G_2} - \frac{C_5 G_2}{C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.30}$$

$$S_{G_2}^{Q_{NI}} = \frac{1}{2}\frac{G_2}{G_1 + G_2} - \frac{C_5 G_2}{\alpha_n C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.31}$$

$$S_{C_3}^{Q} = S_{G_4}^{Q} = \frac{1}{2} - \frac{C_3 G_4}{C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.32}$$

$$S_{C_3}^{Q_{NI}} = S_{G_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n C_3 G_4}{\alpha_n C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.33}$$

$$S_{C_5}^{Q} = \frac{1}{2} - \frac{C_5 (G_1 + G_2)}{C_3 G_4 + C_5 G_1 + C_5 G_2}, \tag{4.3.34}$$

$$S_{C_5}^{Q_{NI}} = \frac{1}{2} - \frac{C_5\left(G_1 + G_2\right)}{\alpha_n C_3 G_4 + C_5 G_1 + C_5 G_2}. \tag{4.3.35}$$

Then again, all ideal and non-ideal sensitivities are bounded by ∓ 0.5.

Sensitivities for VM BP case can be obtained as;

$$S_{C_1}^{\omega} = S_{C_1}^{\omega_{NI}} = \frac{1}{2}\frac{C_1}{\left(C_1 + C_3\right)^2}, \tag{4.3.36}$$

$$S_{C_3}^{\omega} = S_{C_3}^{\omega_{NI}} = \frac{1}{2}\frac{C_3}{\left(C_1 + C_3\right)^2}, \tag{4.3.37}$$

$$S_{G_2}^{\omega} = S_{G_4}^{\omega} = -S_{C_5}^{\omega} = S_{G_2}^{\omega_{NI}} = S_{G_4}^{\omega_{NI}} = -S_{C_5}^{\omega_{NI}} = S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.38}$$

$$S_{C_1}^{Q} = \frac{1}{2}\frac{C_1}{C_1 + C_3} - \frac{C_1 G_4}{C_1 G_4 + C_3 G_4 + C_5 G_2}, \tag{4.3.39}$$

$$S_{C_1}^{Q_{NI}} = \frac{1}{2}\frac{C_1}{C_1 + C_3} - \frac{\alpha_n C_1 G_4}{\alpha_n C_1 G_4 + \alpha_n C_3 G_4 + C_5 G_2}, \tag{4.3.40}$$

$$S_{G_2}^{Q} = S_{C_5}^{Q} = \frac{1}{2} - \frac{C_5 G_2}{C_1 G_4 + C_3 G_4 + C_5 G_2}, \tag{4.3.41}$$

$$S_{G_2}^{Q_{NI}} = S_{C_5}^{Q_{NI}} = \frac{1}{2} - \frac{C_5 G_2}{\alpha_n C_1 G_4 + \alpha_n C_3 G_4 + C_5 G_2}, \tag{4.3.42}$$

$$S_{C_3}^{Q} = \frac{1}{2}\frac{C_3}{C_1 + C_3} - \frac{C_3 G_4}{C_1 G_4 + C_3 G_4 + C_5 G_2}, \tag{4.3.43}$$

$$S_{C_3}^{Q_{NI}} = \frac{1}{2}\frac{C_3}{C_1 + C_3} - \frac{\alpha_n C_3 G_4}{\alpha_n C_1 G_4 + \alpha_n C_3 G_4 + C_5 G_2}, \tag{4.3.44}$$

$$S_{G_4}^{Q} = \frac{1}{2} - \frac{G_4\left(C_1 + C_3\right)}{C_1 G_4 + C_3 G_4 + C_5 G_2}, \tag{4.3.45}$$

$$S_{G_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n G_4 (C_1 + C_3)}{\alpha_n C_1 G_4 + \alpha_n C_3 G_4 + C_5 G_2}. \tag{4.3.46}$$

Again, all of the above sensitivities are equal or smaller than 0.5 in magnitude.

ω_0 in Eq. (4.2.41) and Q in Eq. (4.2.42) for VM HP filter results in the following sensitivities;

$$S_{G_3}^{\omega} = S_{G_4}^{\omega} = -S_{C_5}^{\omega} = S_{G_3}^{\omega_{NI}} = S_{G_4}^{\omega_{NI}} = -S_{C_5}^{\omega_{NI}} = S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.47}$$

$$S_{C_1}^{\omega} = S_{C_1}^{\omega_{NI}} = +\frac{1}{2}\frac{C_1}{(C_1 + C_2)^2}, \tag{4.3.48}$$

$$S_{C_2}^{\omega} = S_{C_2}^{\omega_{NI}} = \frac{1}{2}\frac{C_2}{(C_1 + C_2)^2}, \tag{4.3.49}$$

$$S_{C_1}^{Q} = \frac{1}{2}\frac{C_1}{C_1 + C_2} - \frac{C_1 G_4}{C_1 G_4 + C_2 G_4 + C_5 G_3}, \tag{4.3.50}$$

$$S_{C_1}^{Q_{NI}} = \frac{1}{2}\frac{C_1}{C_1 + C_2} - \frac{\alpha_n C_1 G_4}{\alpha_n C_1 G_4 + \alpha_n C_2 G_4 + C_5 G_3}, \tag{4.3.51}$$

$$S_{C_2}^{Q} = \frac{1}{2}\frac{C_2}{C_1 + C_2} - \frac{C_2 G_4}{C_1 G_4 + C_2 G_4 + C_5 G_3}, \tag{4.3.52}$$

$$S_{C_2}^{Q_{NI}} = \frac{1}{2}\frac{C_2}{C_1 + C_2} - \frac{\alpha_n C_2 G_4}{\alpha_n C_1 G_4 + \alpha_n C_2 G_4 + C_5 G_3}, \tag{4.3.53}$$

$$S_{G_3}^{Q} = S_{C_5}^{Q} = \frac{1}{2} - \frac{C_5 G_3}{C_1 G_4 + C_2 G_4 + C_5 G_3}, \tag{4.3.54}$$

$$S_{G_3}^{Q_{NI}} = S_{C_5}^{Q_{NI}} = \frac{1}{2} - \frac{C_5 G_3}{\alpha_n C_1 G_4 + \alpha_n C_2 G_4 + C_5 G_3}, \tag{4.3.55}$$

$$S_{G_4}^{Q} = \frac{1}{2} - \frac{G_4 (C_1 + C_2)}{C_1 G_4 + C_2 G_4 + C_5 G_3}, \tag{4.3.56}$$

$$S_{G_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n G_4 (C_1 + C_2)}{\alpha_n C_1 G_4 + \alpha_n C_2 G_4 + C_5 G_3}, \tag{4.3.57}$$

It can be seen that all ideal and non-ideal sensitivities are bounded by ∓ 0.5.

Sensitivities of ω_0 and Q of TAM LP case are

$$S_{G_1}^{\omega} = S_{G_1}^{\omega_{NI}} = \frac{1}{2}\frac{G_1}{G_1 + G_2}, \tag{4.3.58}$$

$$S_{G_2}^{\omega} = S_{G_2}^{\omega_{NI}} = \frac{1}{2}\frac{G_2}{G_1 + G_2}, \tag{4.3.59}$$

$$S_{G_5}^{\omega} = -S_{C_3}^{\omega} = -S_{C_4}^{\omega} = S_{G_5}^{\omega_{NI}} = -S_{C_3}^{\omega_{NI}} = -S_{C_4}^{\omega_{NI}} = -S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.60}$$

$$S_{G_1}^{Q} = \frac{1}{2}\frac{G_1}{G_1 + G_2} - \frac{C_4 G_1}{C_3 G_5 + C_4 G_1 + C_4 G_2}, \tag{4.3.61}$$

$$S_{G_1}^{Q_{NI}} = \frac{1}{2}\frac{G_1}{G_1 + G_2} - \frac{\alpha_n C_4 G_1}{C_3 G_5 + \alpha_n C_4 G_1 + \alpha_n C_4 G_2}, \tag{4.3.62}$$

$$S_{G_2}^{Q} = \frac{1}{2}\frac{G_2}{G_1 + G_2} - \frac{C_4 G_2}{C_3 G_5 + C_4 G_1 + C_4 G_2}, \tag{4.3.63}$$

$$S_{G_2}^{Q_{NI}} = \frac{1}{2}\frac{G_2}{G_1 + G_2} - \frac{\alpha_n C_4 G_2}{C_3 G_5 + \alpha_n C_4 G_1 + \alpha_n C_4 G_2}, \tag{4.3.64}$$

$$S_{C_3}^{Q} = S_{G_5}^{Q} = \frac{1}{2} - \frac{C_3 G_5}{C_3 G_5 + C_4 G_1 + C_4 G_2}, \tag{4.3.65}$$

$$S_{C_3}^{Q_{NI}} = S_{G_5}^{Q_{NI}} = \frac{1}{2} - \frac{C_3 G_5}{C_3 G_5 + \alpha_n C_4 G_1 + \alpha_n C_4 G_2}, \tag{4.3.66}$$

$$S_{C_4}^{Q} = \frac{1}{2} - \frac{C_4 (G_1 + G_2)}{C_3 G_5 + C_4 G_1 + C_4 G_2}, \tag{4.3.67}$$

$$S_{C_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n C_4 (G_1 + G_2)}{C_3 G_5 + \alpha_n C_4 G_1 + \alpha_n C_4 G_2}, \tag{4.3.68}$$

For this case, it is seen that all ideal and non-ideal sensitivities are equal or smaller than 1 in magnitude.

Last, sensitivity calculations for ω_0 and Q of TAM HP filter are given as

$$S_{G_4}^{\omega} = -S_{C_1}^{\omega} = -S_{C_5}^{\omega} = S_{G_4}^{\omega_{NI}} = -S_{C_1}^{\omega_{NI}} = -S_{C_5}^{\omega_{NI}} = S_{\alpha_n}^{\omega_{NI}} = \frac{1}{2}, \tag{4.3.69}$$

$$S_{G_2}^{\omega} = S_{G_2}^{\omega_{NI}} = \frac{1}{2} \frac{G_2}{G_2 + G_3}, \tag{4.3.70}$$

$$S_{G_3}^{\omega} = S_{G_3}^{\omega_{NI}} = \frac{1}{2} \frac{G_3}{G_2 + G_3}, \tag{4.3.71}$$

$$S_{C_1}^{Q} = S_{G_4}^{Q} = \frac{1}{2} - \frac{C_1 G_4}{C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.72}$$

$$S_{C_1}^{Q_{NI}} = S_{G_4}^{Q_{NI}} = S_{\alpha_n}^{Q_{NI}} = \frac{1}{2} - \frac{\alpha_n C_1 G_4}{\alpha_n C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.73}$$

$$S_{G_2}^{Q} = \frac{1}{2} \frac{G_2}{G_2 + G_3} - \frac{C_5 G_2}{C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.74}$$

$$S_{G_2}^{Q_{NI}} = \frac{1}{2} \frac{G_2}{G_2 + G_3} - \frac{C_5 G_2}{\alpha_n C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.75}$$

$$S_{G_3}^{Q} = \frac{1}{2} \frac{G_3}{G_2 + G_3} - \frac{C_5 G_3}{C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.76}$$

$$S_{G_3}^{Q_{NI}} = \frac{1}{2} \frac{G_3}{G_2 + G_3} - \frac{C_5 G_3}{\alpha_n C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.77}$$

$$S_{C_5}^{Q} = \frac{1}{2} - \frac{C_5 (G_2 + G_3)}{C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.78}$$

$$S_{C_5}^{Q_{NI}} = \frac{1}{2} - \frac{C_5(G_2 + G_3)}{\alpha_n C_1 G_4 + C_5 G_2 + C_5 G_3}, \tag{4.3.79}$$

All sensitivities for this case are found to be limited by ∓ 0.5.

4.4 SIMUALTION AND EXPERIMENTAL RESULTS

The validity of proposed LP, BP and HP filters in VM, CM, TIM and TAM are simulated using PSpice simulation program. Simulation results are shown in sub-sections according to usage mode with selected design parameters. Lackness of an ideal current source impeded us to make experimental realization of simulated TIM, TAM and CM filter topologies and only VM filter topologies are experimented. Thus only the comparison of simulation and experimental results of VM filters is given.

In this work, due to the conventional implementation of CDBA with two AD844s (Acar and Ozoguz, 1999) has given us a noisy *w*-terminal voltage output, a new CDBA implementation by using three commercially available AD844s is realized for experimental purposes. This implementation is shown in Fig. 4.2.

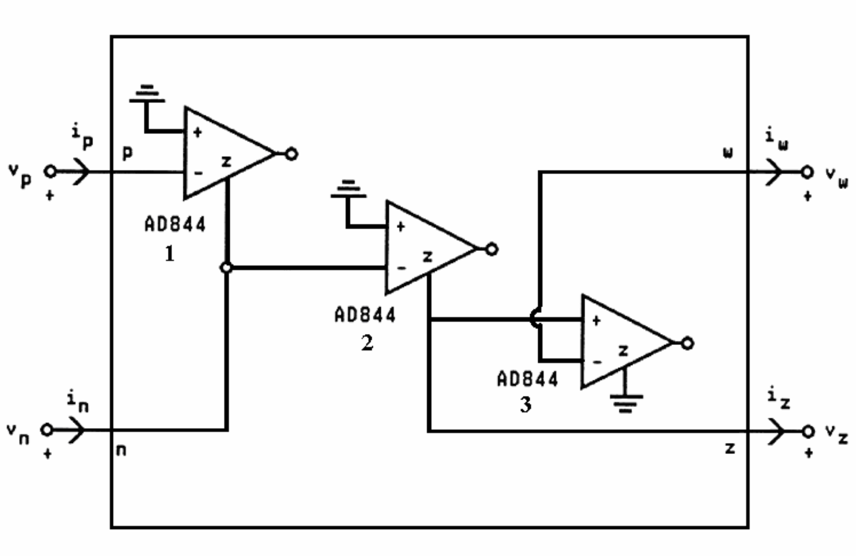

Figure 4.2 The proposed implementation of CDBA.

The new implementation uses input voltage relation of AD844 (Analog Devices, 1990) given in Eq. (2.3.2).

This relation gives us an opportunity to transfer the voltage at the *z*-terminal of the second AD844 into the *p*-terminal of the third AD844 as shown in Fig. 4.2. Since $V_p = V_n$, *n*-terminal of the third AD844 can be used as a *w*-terminal voltage output of CDBA which is free from the noise that arises in *w*-terminals of simulation by using two AD844s.

The simulations are carried out using the model parameters of AD844 from the built in library of PSpice simulation program and two AD844s are used to realize CDBA in PSpice. However, the experimental results are obtained by using three AD844s which is shown in Fig. 4.2.

4.4.1 CM Filters

Proposed CM filters are simulated with PSpice simulation program with the given design parameters in Table 4.3.

Table 4.3 Design parameters used for CM filters.

Filter Function	y_2	y_3	y_4	y_5
CM-LP	$R_2 = 1k\Omega$	$C_3 = 10nF$	$C_4 = 1nF$	$R_5 = 1k\Omega$
CM-BP	$R_2 = 100\Omega$	$C_3 = 10nF$	$R_4 = 10k\Omega$	$C_5 = 1nF$
CM-HP	$C_2 = 10nF$	$R_3 = 1k\Omega$	$R_4 = 1k\Omega$	$C_5 = 1nF$

Using the above passive elements, simulation and theoretical are conducted using symmetrical $\pm 5\,V$ supplies for an undamped natural frequency of $50\,kHz$. The simulation results shown in Figures 4.3-5 are obtained. All of the filter characteristics are redrawn in Fig. 4.6 together.

As seen in the mentioned figures, simulation of LP filter is matched very well with the theoretical plot. Due to the tracking errors, BP and HP filter simulation responses have a small discrepancy from the theoretical ones. BP characteristics exhibit differences due to mismatch in the center frequency and the bandwidth. Further difference at high frequencies for HP response is caused by parasitic capacitance effects at the terminals of CDBA.

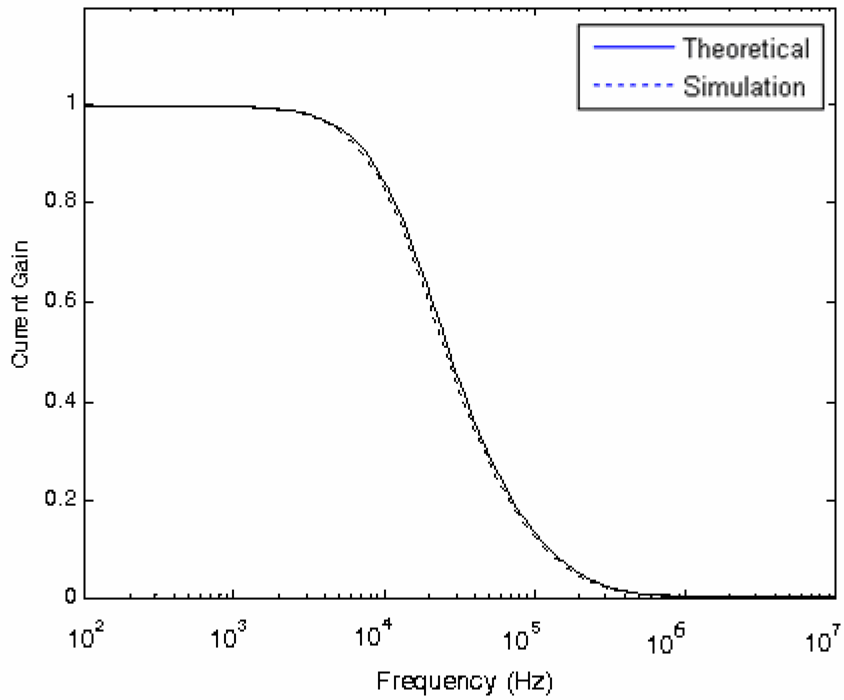

Figure 4.3 LP characteristics for CM topology.

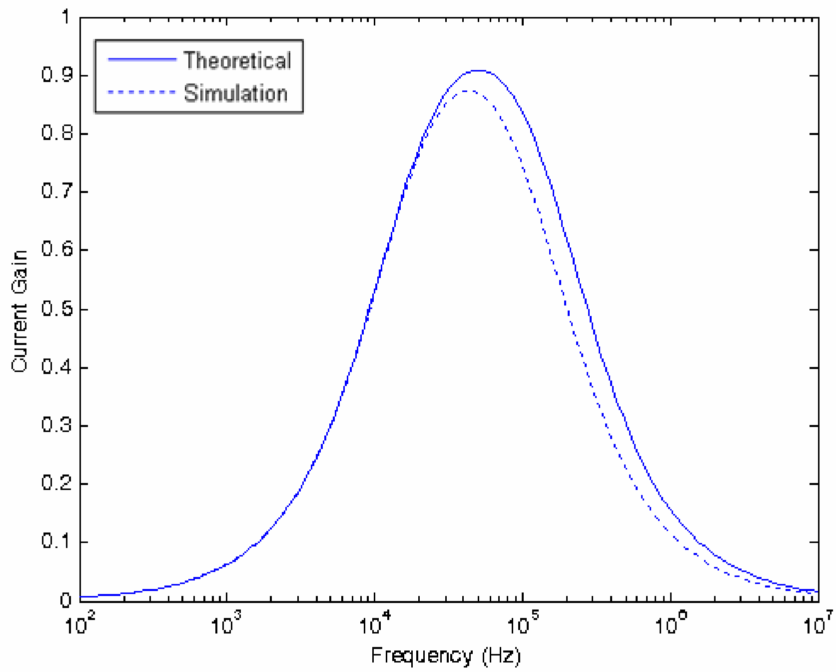

Figure 4.4 BP characteristics for CM topology.

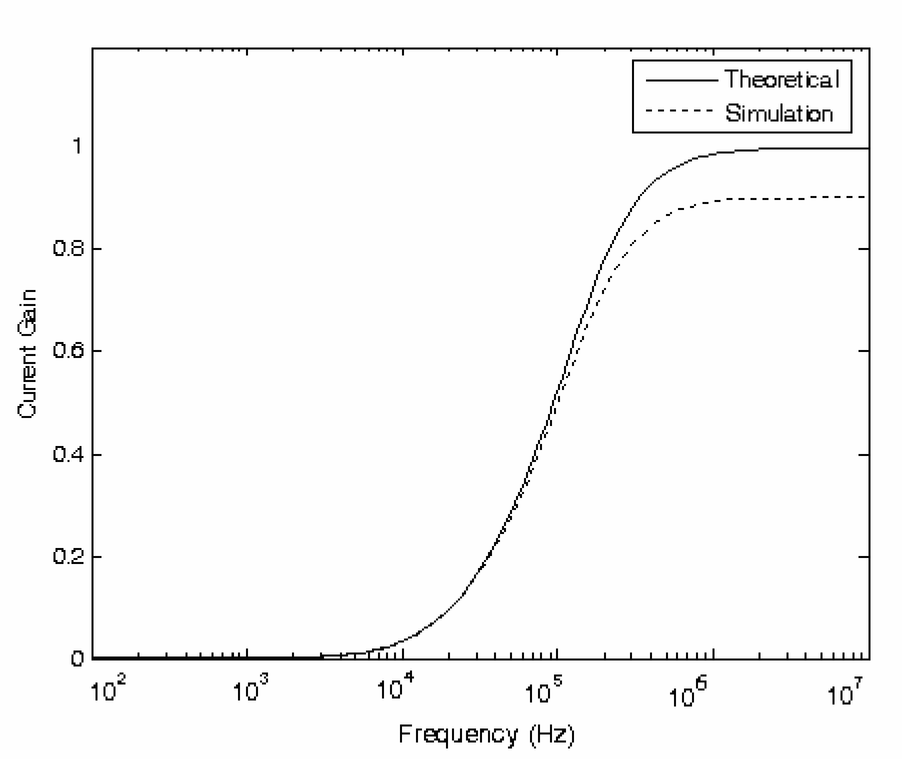

Figure 4.5 HP characteristics for CM topology.

4.4.2 VM Filters

The validity of proposed VM LP, BP and HP filters are shown in Fig. 4.7-4.8, and in Fig. 4.9 with simulation and experimental results together, respectively. As already mentioned in this section, due to the noise existence at the *w*-terminal, we use three AD844s to realize CDBA for experimental verifications. VM circuits are supplied with symmetrical voltages of $\mathrm{V_{cc}} = 5\ \mathrm{V}$ and $\mathrm{V_{ee}} = -5\ \mathrm{V}$, and are designed for using the parameters shown in Table 4.4.

Table 4.4 Design parameters used for VM filters.

Filter Function	y_1	y_2	y_3	y_4	y_5
VM-LP	$R_1 = 100\,\Omega$	$R_2 = 1\,k\Omega$	$C_3 = 1\,nF$	$R_4 = 1\,k\Omega$	$C_5 = 10\,nF$
VM-BP	$C_1 = 2.2\,nF$	$R_2 = 470\,\Omega$	$C_3 = 10\,nF$	$R_4 = 4.7\,k\Omega$	$C_5 = 1\,nF$
VM-HP	$C_1 = 1\,nF$	$C_2 = 100\,pF$	$R_3 = 1\,k\Omega$	$R_4 = 100\,k\Omega$	$C_5 = 100\,pF$

As seen in the figures 4.7, 4.8, the comparison of the simulation and experimental plots show that they are in good agreement. Over the 3 MHz as in Fig. 4.9, the discrepancies between the theoretical and simulation results start to higher values, this much error is expected since the internal capacitances of AD844s are much more affective at high frequencies.

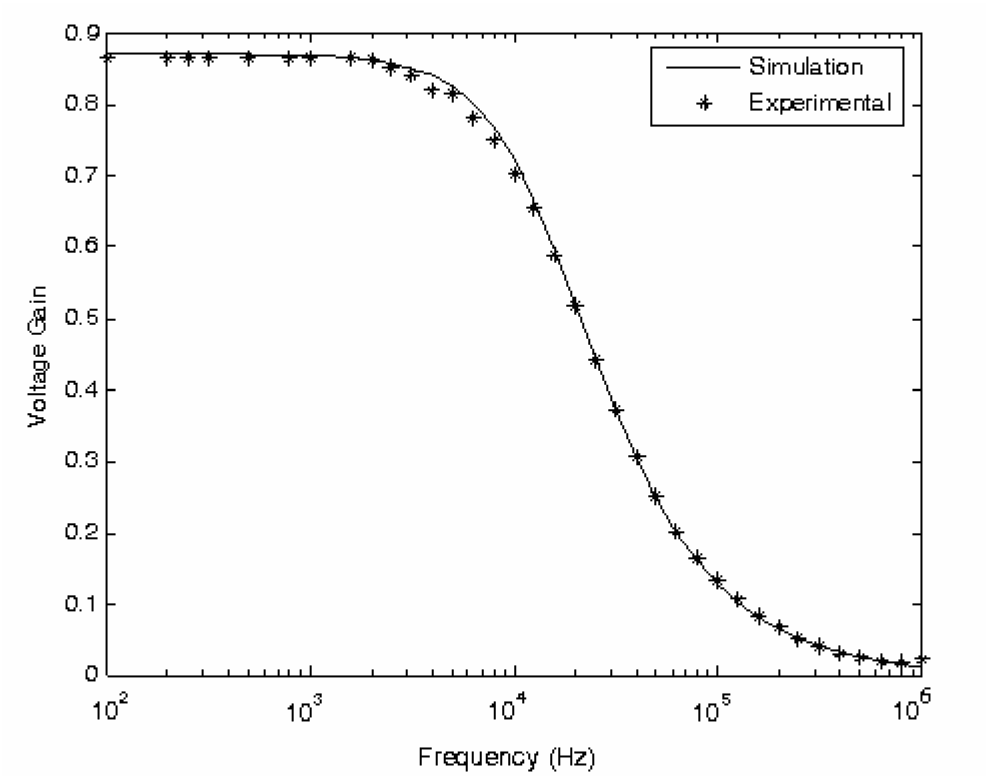

Figure 4.6 LP characteristics for VM topology.

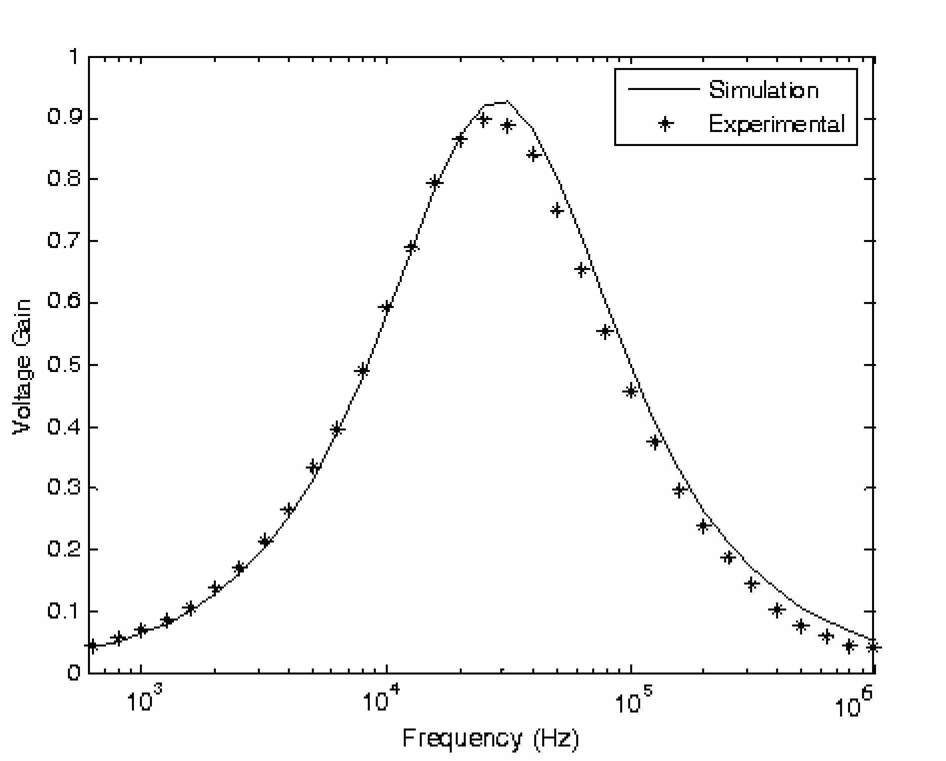

Figure 4.7 BP characteristics for VM topology.

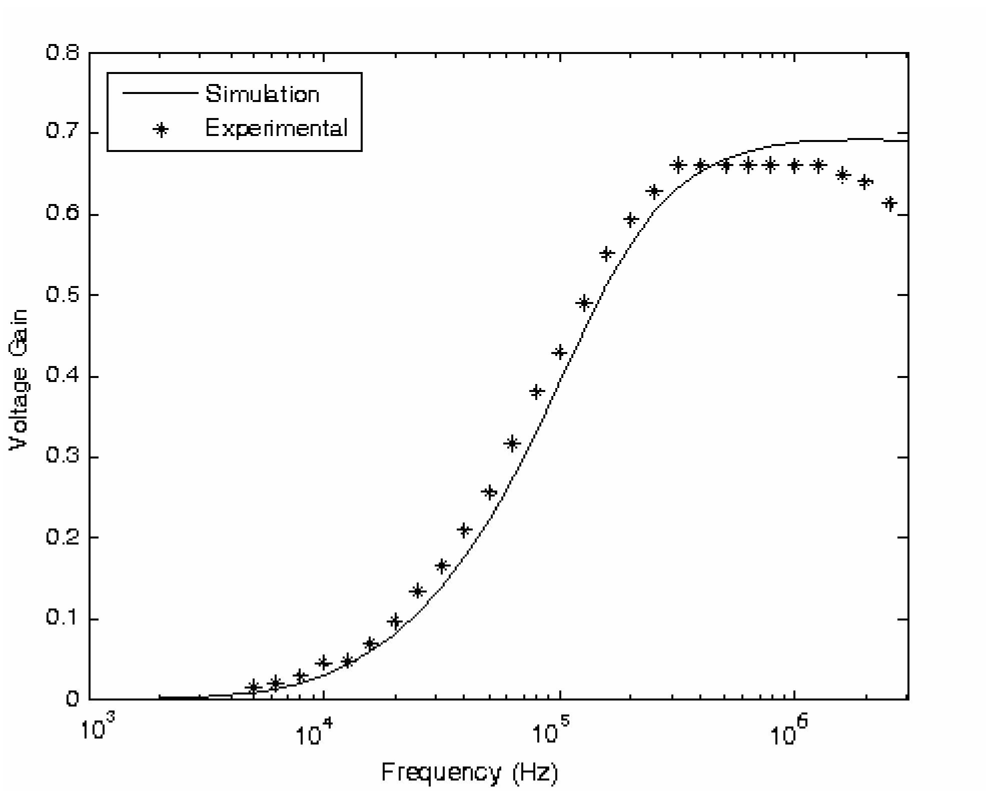

Figure 4.8 HP characteristics for VM topology.

4.4.3 TIM Filters

Proposed TIM filters are simulated with PSpice simulation program with the given design parameters in Table 4.5 and simulation results are given in Figures 4.10-12. Filters are biased with symmetrical $\pm 5\,V$ and an input current of $1\,mA$ is used in order to have V_{out} in the level of Volts.

Table 4.5 Design parameters used for TIM filters.

Filter Function	y_2	y_3	y_4	y_5
TIM-LP	$R_2 = 2\,k\Omega$	$C_3 = 2\,nF$	$R_4 = 1\,k\Omega$	$C_5 = 1\,nF$
TIM-BP	$C_2 = 2\,nF$	$R_3 = 1\,k\Omega$	$R_4 = 2\,k\Omega$	$C_5 = 1\,nF$

Though very small discrepancies exist, simulation results of LP and BP filters demonstrate a good agreement with the theoretical ones as shown in Figs. 4.9,10. A HP filter simulation is not possible due to the lackness of a suitable HP filter transfer function for TIM.

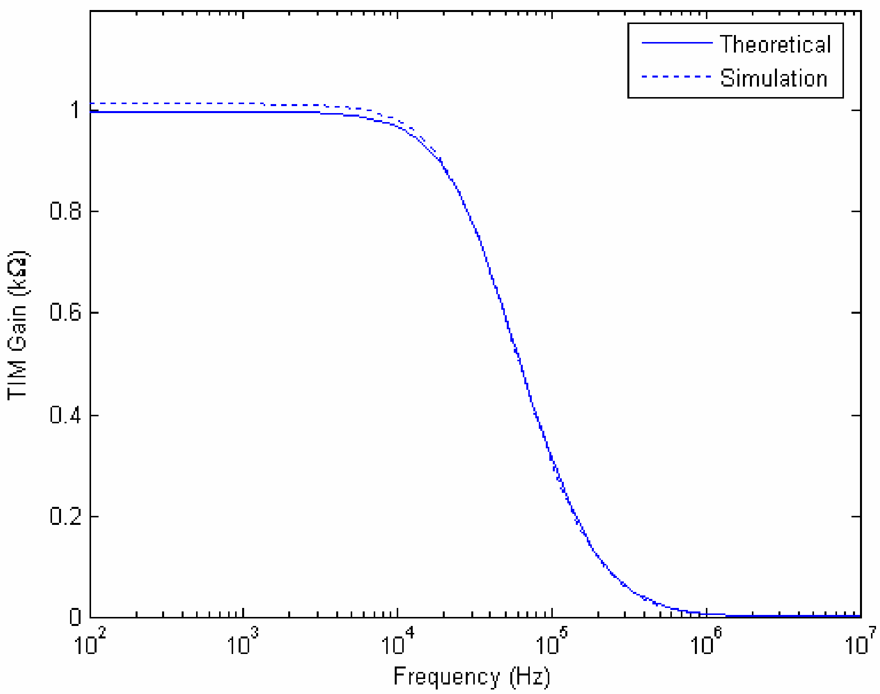

Figure 4.9 LP characteristics for TIM topology.

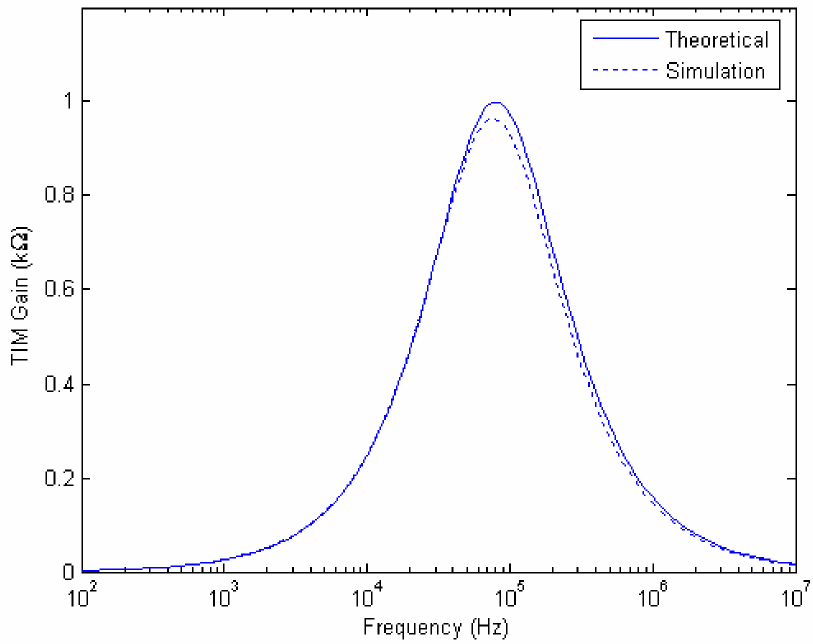

Figure 4.10 BP characteristics for TIM topology.

4.4.4 TAM Filters

Proposed TAM filters are simulated with PSpice simulation program with symmetrical $\pm 5\,V$ supply voltages and an input voltage of 1 V to obtain an output current in the level of miliamperes. The design parameters for TAM filters are given in Table 4.6 and simulation results are shown in Figs. 4.11-13.

Table 4.6 Design parameters used for TAM filters.

Filter Function	y_1	y_2	y_3	y_4	y_5
TAM-LP	$R_1 = 800\,\Omega$	$R_2 = 200\,\Omega$	$C_3 = 10\,nF$	$C_4 = 100\,pF$	$R_5 = 10\,k\Omega$
TAM-BP	$R_1 = 800\,\Omega$	$R_2 = 200\,\Omega$	$C_3 = 10\,nF$	$R_4 = 10\,k\Omega$	$C_5 = 100\,pF$
TAM-HP	$C_1 = 10\,nF$	$R_2 = 800\,\Omega$	$R_3 = 200\,\Omega$	$R_4 = 10\,k\Omega$	$C_5 = 100\,pF$

Non-linear effects again cause discrepancies on filter responses given in Figs. 4.11-13. In the LP case the maximum discrepancy occurs at lower frequencies and the simulation results are lower than 5% from the expected (theoretical) results. However, for the BP case the validitation shows very weak coherence at the center frequency. The results obtained by the simulation are smaller 15% for the gain, 7% for the center frequency than the theoretical ones. The same weakness also appears in Fig. 4.13 for the HP case, the simulation results are 11% lower than the theoretical results at high frequencies. These discrepancies are expected due to the non-zero input resistances (50-65 ohm) and output resistance (15 ohm) of AD844 used for the simulation of CDBA. (Analog Devices, 1990); and comparable resistance values (minimum 200 ohm) are used in the simulations. Frequency dependent discrepancies are not effective at the considered band (0-10 Mhz) for the parasitic capacitances of AD844 (maximum 4 pF) are much smaller than the ones used in the circuit.

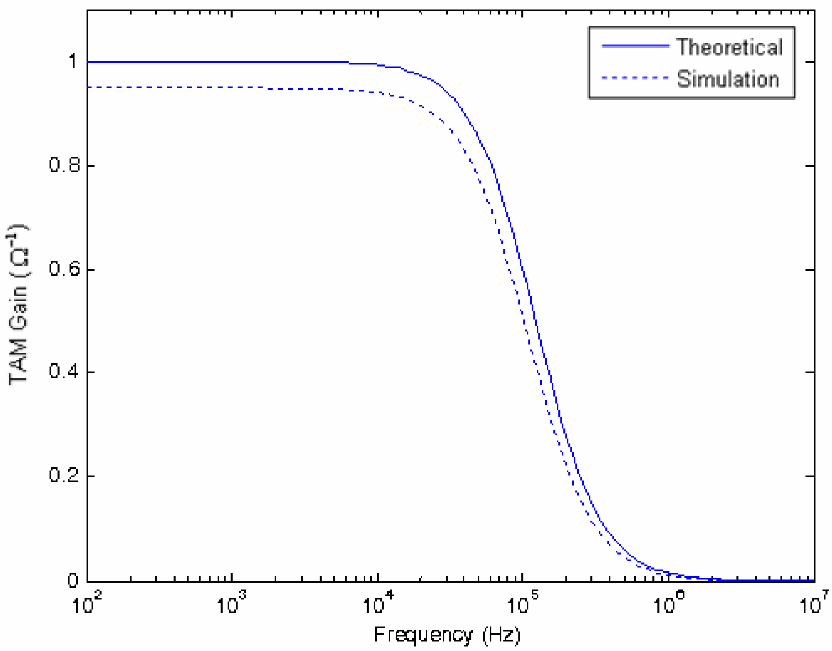

Figure 4.11 LP characteristics for TAM topology.

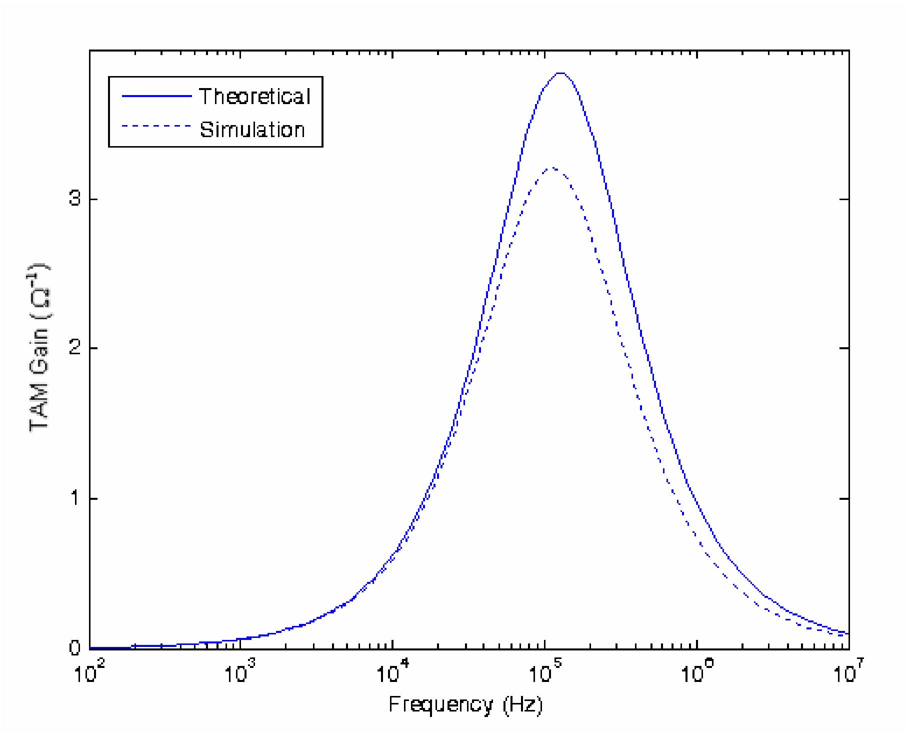

Figure 4.12 BP characteristics for TAM topology.

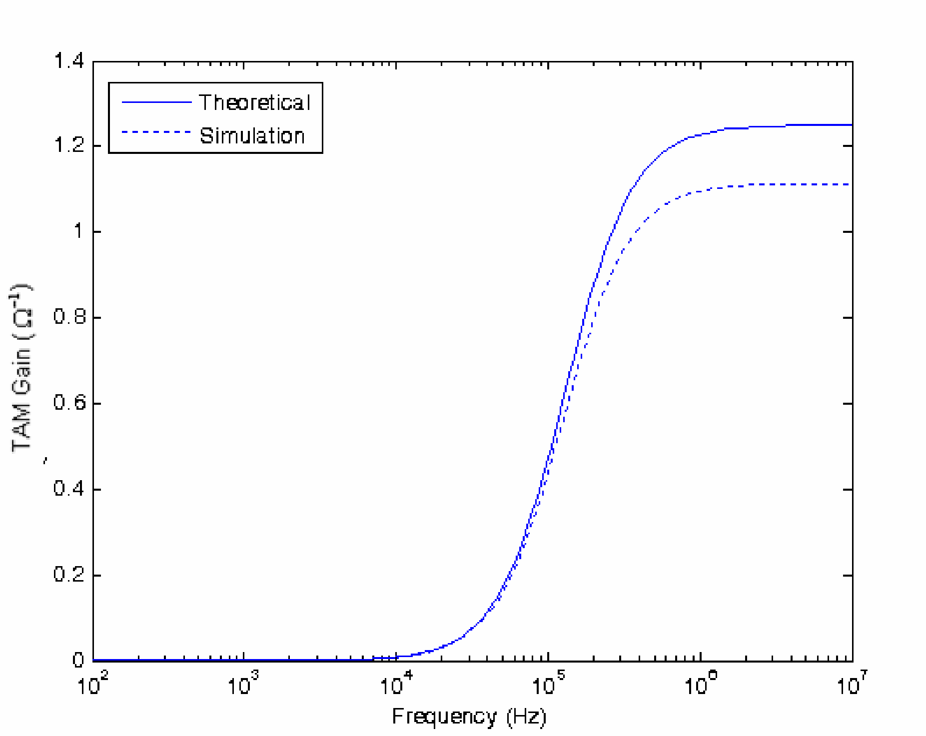

Figure 4.13 HP characteristics for TAM topology.

CHAPTER 5

A NEW VOLTAGE-MODE SINGLE INPUT MULTIPLE OUTPUT MULTI-FUNCTION FILTER USING CDBAs

In this chapter, a SIMO filter configuration that realizes all five basic filter transfer characteristics is given. Eight topologies of the proposed configuration are investigated by changing two admittances and a single topology that realizes a universal filter is presented. A summary of previously reported multi-function configurations are given as an introductory section. Circuit topologies, their analyses and a table of realized filters are given in the second section of this chapter. The third section is devoted to sensitivity analysis of obtained filters. Finally, PSpice simulations of the topologies are given and the results are compared with the theoretical ones.

5.1 INTRODUCTION

Several VM multi-function filter configurations are proposed using CDBA. Koksal and Sagbas realized a second-order VM universal filter using seven resistors, two capacitors and four CDBAs (Koksal and Sagbas, 2007). In 2005, Sagbas and Koksal presented a new VM multi-function filter that uses eight resistors, two capacitors and four CDBAs (Sagbas and Koksal, 2005). A realization of multiple-output biquadratic filters is given in the same year (Tangsrirat and Surakampontorn, 2005). This filter configuration has two topologies and uses three CDBAs, seven resistors-two capacitors or five resistors-four capacitors. The configuration presented in this chapter is an improved version of the last one.

Presented configuration offers the following advantages;

- One passive element is reduced from the previous work (Tangsrirat and Surakampontorn, 2005).
- A single topology using six resistors, two capacitors and three CDBAs realizes all five basic filter functions by altering two resistors only.
- Although, *w*-terminals are conventionally used for voltage outputs, in this configuration *z*-terminals can be used as voltage outputs and the voltage buffer stage in CDBAs can be eliminated for a fully integrated circuit implementation.
- Considering configurations using three CDBAs, a minimum number of active and/or passive elements are used in the given topologies for certain cases.

5.2 CIRCUIT CONFIGURATIONS AND THEIR ANALYSES

Proposed configuration for a SIMO VM multi-function filter is given in Fig. 5.1. First sub-section demonstrates the derivation of the general transfer function using conventional circuit analysis techniques. In the second sub-section, several filter transfer functions are obtained by appropriate passive element topologies and resulting filter transfer functions are summarized by a table.

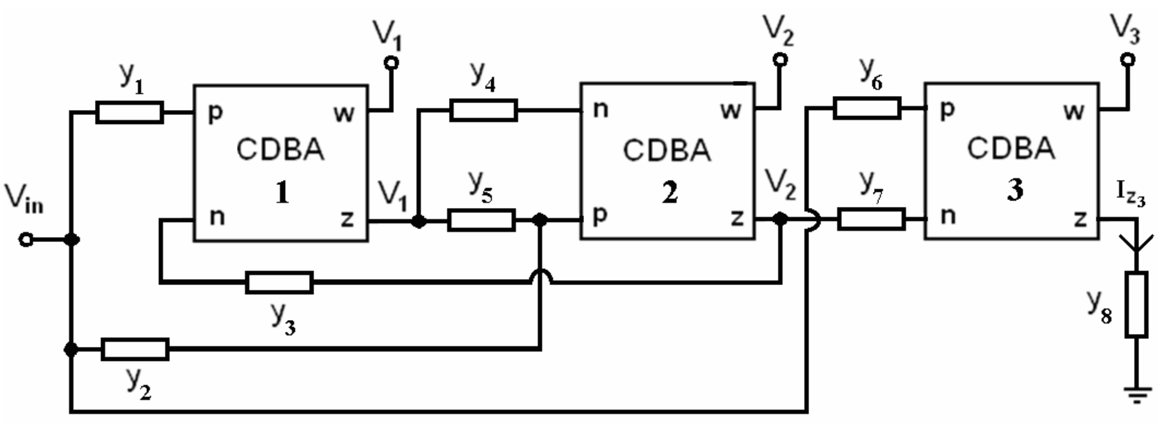

Figure 5.1 Proposed SIMO VM multi-function configuration.

5.2.1 Derivation of General Voltage Transfer Functions

Considering CDBA-1 in Fig. 5.1, by using Eq. (2.1.2) and Eq. (2.1.3) we obtain,

$$V_1(y_4 + y_5) = V_{in}y_1 - V_2 y_3. \tag{5.2.1}$$

Likewise, it is easy to obtain

$$V_2(y_3 + y_7) = V_{in}y_2 - V_1(y_4 - y_5) \tag{5.2.2}$$

by using the terminal equations for CDBA-2. These two equalities can be written in the matrix form as

$$\begin{bmatrix} y_4 + y_5 & y_3 \\ y_4 - y_5 & y_3 + y_7 \end{bmatrix}\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} V_{in}. \tag{5.2.3}$$

The general transfer function matrix is easily obtained as

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \frac{1}{\Delta}\begin{bmatrix} y_3 + y_7 & -y_3 \\ y_5 - y_4 & y_4 + y_5 \end{bmatrix}\begin{bmatrix} y_1 \\ y_2 \end{bmatrix} V_{in}, \tag{5.2.4}$$

where,

$$\Delta = (y_3 + y_7)(y_4 + y_5) - y_3(y_4 - y_5) = 2y_3y_5 + y_7(y_4 + y_5). \tag{5.2.5}$$

From Eq. (5.2.4) we obtain the transfer function for V_1 explicitly as follows

$$\frac{V_1}{V_{in}} = \frac{y_1(y_3 + y_7) - y_2y_3}{2y_3y_5 + y_7(y_4 + y_5)}, \tag{5.2.6}$$

After denominator arrangement, the resultant transfer function is

$$\frac{V_1}{V_{in}} = \frac{y_1(y_3 + y_7) - y_2y_3}{2y_3y_5 + y_4y_7 + y_5y_7}. \tag{5.2.7}$$

Non-ideal transfer function can be obtained by using Eq. (2.1.5) as follows

$$\frac{V_{1_{NI}}}{V_{in}}=\frac{1}{\beta_1}\frac{\alpha_{p_1}y_1(y_3+y_7)-\alpha_{p_2}\alpha_{n_1}y_2y_3}{y_3y_4(1-\alpha_{n_1}\alpha_{n_2})+y_3y_5(1+\alpha_{p_2}\alpha_{n_1})+y_7(y_4+y_5)}. \tag{5.2.8}$$

In the same way, we can find the transfer function for V_2 by using Eq. (5.2.4);

$$\frac{V_2}{V_{in}}=\frac{y_1(y_5-y_4)+y_2(y_4+y_5)}{2y_3y_5+y_7(y_4+y_5)}. \tag{5.2.9}$$

After arrangements at the denominator, the above transfer function is written as

$$\frac{V_2}{V_{in}}=\frac{y_1(y_5-y_4)+y_2(y_4+y_5)}{2y_3y_5+y_4y_7+y_5y_7}, \tag{5.2.10}$$

Non-ideal transfer function can be obtained to be

$$\frac{V_{2_{NI}}}{V_{in}}=\frac{1}{\beta_2}\frac{\alpha_{p_1}y_1(\alpha_{p_2}y_5-\alpha_{n_2}y_4)+\alpha_{p_2}y_2(y_4+y_5)}{y_3y_4(1-\alpha_{n_1}\alpha_{n_2})+y_3y_5(1+\alpha_{p_2}\alpha_{n_1})+y_7(y_4+y_5)}. \tag{5.2.11}$$

Now by using Eq. (2.1.2) and Eq. (2.1.3) for the third CDBA, we obtain

$$I_{z_3}=V_{in}y_6-V_2y_7. \tag{5.2.12}$$

And the current flowing through *z*-terminal of third CDBA is

$$I_{z_3}=V_3y_8. \tag{5.2.13}$$

Substituting Eq. (5.2.13) into Eq. (5.2.12) we get

$$V_3y_8=V_{in}y_6-V_2y_7. \tag{5.2.14}$$

Furthermore, if we substitute Eq. (5.2.9) into Eq. (5.2.14), we confront with the following equality;

$$V_3y_8=V_{in}y_6-V_{in}\frac{y_1(y_5-y_4)+y_2(y_4+y_5)}{2y_3y_5+y_7(y_4+y_5)}y_7. \tag{5.2.15}$$

Then, the transfer function for V_3 is

$$\frac{V_3}{V_{in}} = \frac{2y_3y_5y_6 + y_6y_7(y_4 + y_5) - y_1y_7(y_5 - y_4) - y_2y_7(y_4 + y_5)}{y_8[2y_3y_5 + y_7(y_4 + y_5)]}, \tag{5.2.16}$$

Re-arrangement results in the final voltage transfer function below;

$$\frac{V_3}{V_{in}} = \frac{2y_3y_5y_6 + y_6y_7(y_4 + y_5) - y_1(y_5y_7 - y_4y_7) - y_2(y_4y_7 + y_5y_7)}{y_8[2y_3y_5 + y_7(y_4 + y_5)]}. \tag{5.2.17}$$

Non-ideal transfer function for V_3 / V_{in} is obtained as follows

$$\frac{V_{3_{NI}}}{V_{in}} = \frac{\alpha_{p_3} y_3 y_6 \left[y_4\left(1-\alpha_{n_1}\alpha_{n_2}\right) + y_5\left(1+\alpha_{p_2}\alpha_{n_1}\right)\right] - \alpha_{n_3} y_7 \left[y_1 \alpha_{p_1}\left(\alpha_{p_2} y_5 - \alpha_{n_2} y_4\right) + y_2 y_7 \alpha_{p_2}\left(y_4 + y_5\right)\right]}{\beta_3 y_8 \left[y_3 y_4\left(1-\alpha_{n_1}\alpha_{n_2}\right) + y_3 y_5\left(1+\alpha_{p_2}\alpha_{n_1}\right) + y_7\left(y_4 + y_5\right)\right]}. \tag{5.2.18}$$

5.2.2 Analyses of General Voltage Transfer Functions

Having obtained the general transfer functions for the voltage outputs V_1, V_2 and V_3, we consider which kind of filter characteristics can be obtained using these outputs. A general denominator characteristic can be expressed for the voltage output transfer functions with the passive element selections given as

$$y_3 = sC_3,\; y_4 = G_4,\; y_5 = sC_5,\; y_6 = G_6,\; y_7 = G_7,\; y_8 = G_8.$$

By using Eq. (5.2.7), we have the following ideal output voltage transfer function for V_1 output,

$$\frac{V_1}{V_{in}} = \frac{y_1(sC_3 + G_7) - y_2(sC_3)}{2s^2C_3C_5 + sC_5G_7 + G_4G_7}. \tag{5.2.19}$$

And for the non-ideal case it becomes

$$\frac{V_{1_{NI}}}{V_{in}} = \frac{1}{\beta_1} \frac{\alpha_{p_1} y_1(sC_3 + G_7) - \alpha_{p_2}\alpha_{n_1} y_2(sC_3)}{s^2C_3C_5\left(1+\alpha_{p_2}\alpha_{n_1}\right) + s\left[C_3G_4\left(1-\alpha_{n_1}\alpha_{n_2}\right) + C_5G_7\right] + G_4G_7}. \tag{5.2.20}$$

In the same way, we can find the voltage transfer function for V_2 by using Eq. (5.2.10),

$$\frac{V_2}{V_{in}} = \frac{y_1(sC_5 - G_4) + y_2(G_4 + sC_5)}{2s^2C_3C_5 + sC_5G_7 + G_4G_7}. \tag{5.2.21}$$

The non-ideal transfer equation for V_2 becomes

$$\frac{V_{2_{NI}}}{V_{in}} = \frac{1}{\beta_2}\frac{\alpha_{p_1}y_1(\alpha_{p_2}sC_5 - \alpha_{n_2}G_4) + \alpha_{p_2}y_2(G_4 + sC_5)}{s^2C_3C_5(1 + \alpha_{p_2}\alpha_{n_1}) + s[C_3G_4(1 - \alpha_{n_1}\alpha_{n_2}) + C_5G_7] + G_4G_7}. \tag{5.2.22}$$

The last transfer function is written using Eq. (5.2.17) as follows

$$\frac{V_3}{V_{in}} = \frac{2s^2C_3C_5G_6 + G_6G_7(G_4 + sC_5) - y_1(sC_5G_7 - G_4G_7) - y_2(G_4G_7 + sC_5G_7)}{G_8(2s^2C_3C_5 + G_4G_7 + sC_5G_7)}. \tag{5.2.23}$$

By using Eq. (5.2.20), non-ideal response becomes

$$\frac{V_{3_{NI}}}{V_{in}} = \frac{\alpha_{p_3}sC_3G_6[G_4(1-\alpha_{n_1}\alpha_{n_2}) + sC_5(1+\alpha_{p_2}\alpha_{n_1})] - \alpha_{n_3}G_7[y_1\alpha_{p_1}(\alpha_{p_2}sC_5 - \alpha_{n_2}G_4) + y_2\alpha_{p_2}(G_4 + sC_5)]}{\beta_3G_8[s^2C_3C_5(1+\alpha_{p_2}\alpha_{n_1}) + sC_3G_4(1-\alpha_{n_1}\alpha_{n_2}) + G_7(G_4 + sC_5)]}. \tag{5.2.24}$$

Now, all three voltage transfer functions can be analyzed in eight combinations of y_1 and y_2 admittances. We consider each case and derive resulting transfer functions for every output in the following sub-sections.

5.2.2.1 Case 1: $y_1 = G_1$ and $y_2 = G_2$

For the given admittances, the transfer functions given in Eqs. (5.2.19), (5.2.21) and (5.2.23) will be

$$\frac{V_1}{V_{in}} = \frac{s\dfrac{G_1 - G_2}{2C_5} + \dfrac{G_1G_7}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}. \tag{5.2.25}$$

$$\frac{V_2}{V_{in}} = \frac{s\dfrac{G_1 + G_2}{2C_3} + \dfrac{G_4(G_2 - G_1)}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}. \tag{5.2.26}$$

$$\frac{V_3}{V_{in}} = \frac{s^2 \frac{G_6}{G_8} - s\frac{G_7(G_1 + G_2 - G_6)}{2C_3G_8} + \frac{G_7(G_4G_6 + G_1G_4 - G_2G_4)}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}. \tag{5.2.27}$$

respectively. A LP filter characteristic can be obtained easily by choosing $G_1 = G_2$ for Eq. (5.2.25). Furthermore, for the same condition $G_1 = G_2$ a BP filter characteristic can also be obtained from Eq. (5.2.26). The last transfer function in Eq. (5.2.27) gives a BR filter function for $G_1 + G_2 = G_6$ and an AP filter function for $G_1 = G_2 = G_6 = G_8$.

5.2.2.2 Case 2: $y_1 = G_1$ *and* $y_2 = 0$

Voltage transfer functions in Eqs. (5.2.19), (5.2.21) and (5.2.23) will be in the following forms with the given admittance choices

$$\frac{V_1}{V_{in}} = \frac{s\frac{G_1}{2C_5} + \frac{G_1G_7}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}, \tag{5.2.28}$$

$$\frac{V_2}{V_{in}} = \frac{s\frac{G_1}{2C_3} - \frac{G_1G_4}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}, \tag{5.2.29}$$

$$\frac{V_3}{V_{in}} = \frac{s^2 \frac{G_6}{G_8} + s\frac{G_6G_7 - G_1G_7}{2C_3G_8} + \frac{G_4G_6G_7 + G_1G_4G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}, \tag{5.2.30}$$

respectively. Although it is not in the traditional form, Eqs. (5.2.28) and (5.2.29) give minimum and non-minimum phase LP filter functions, respectively. Also the last voltage transfer function demonstrates a BR filter characteristic with the condition $G_1 = G_6$.

5.2.2.3 Case 3: $y_1 = 0$ and $y_2 = G_2$

Voltage transfer function in Eqs. (5.2.19), (5.2.21) and (5.2.23) yield

$$\frac{V_1}{V_{in}} = \frac{-s\dfrac{G_2}{2C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.31}$$

$$\frac{V_2}{V_{in}} = \frac{s\dfrac{G_2}{2C_3} + \dfrac{G_2G_4}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.32}$$

$$\frac{V_3}{V_{in}} = \frac{s^2\dfrac{G_6}{G_8} - s\dfrac{G_7(G_2 - G_6)}{2C_3G_8} + \dfrac{G_7G_4(G_6 - G_2)}{2C_3C_5G_8}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.33}$$

respectively. The transfer function in Eq. (5.2.31) is a traditional IBP filter characteristic. Eq. (5.2.32) gives a non-traditional LP filter. If we choose $G_2 = G_6$, *s*-term and constant term in the nominator of Eq. (5.2.33) become zero, thus we obtain a HP filter transfer function.

5.2.2.4 Case 4: $y_1 = sC_1$ and $y_2 = sC_2$

This case reduces Eqs. (5.2.19), (5.2.21) and (5.2.23) to

$$\frac{V_1}{V_{in}} = \frac{s^2\dfrac{C_1 - C_2}{2C_5} + s\dfrac{C_1G_7}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.34}$$

$$\frac{V_2}{V_{in}} = \frac{s^2\dfrac{C_1 + C_2}{2C_3} + s\dfrac{G_4(C_2 - C_1)}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.35}$$

$$\frac{V_3}{V_{in}}=\frac{s^2\left(\frac{G_6}{G_8}-G_7\frac{C_1+C_2}{2C_3G_8}\right)+s\frac{G_7\left(C_5G_6+C_1G_4-C_2G_4\right)}{2C_3C_5G_8}+\frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2+s\frac{G_7}{2C_3}+\frac{G_4G_7}{2C_3C_5}}, \quad (5.2.36)$$

respectively. The transfer functions in Eqs. (5.2.34) and (5.2.35) show non-traditional HP filter characteristics. With $C_1 > C_2$ the second one becomes a non-minimum phase function. In particular, the condition $C_1 = C_2$ makes these functions a BP and HP filter transfer functions, respectively. By the conditions $C_5G_6 + C_1G_4 = C_2G_4$ and $2C_3G_6 > G_7(C_1 + C_2)$, we obtain a BR filter from Eq. (5.2.36); further this equation yields a LP filter if $2C_3G_6 = G_7(C_1 + C_2)$. These conditions are simplified if we choose $G_4 = G_6 = G_7$.

5.2.2.5 Case 5: $y_1 = sC_1$ *and* $y_2 = 0$

In a similar manner we obtain the following transfer functions for this choice;

$$\frac{V_1}{V_{in}}=\frac{s^2\frac{C_1}{2C_5}+s\frac{C_1G_7}{2C_3C_5}}{s^2+s\frac{G_7}{2C_3}+\frac{G_4G_7}{2C_3C_5}}, \quad (5.2.37)$$

$$\frac{V_2}{V_{in}}=\frac{s^2\frac{C_1}{2C_3}-s\frac{G_4C_1}{2C_3C_5}}{s^2+s\frac{G_7}{2C_3}+\frac{G_4G_7}{2C_3C_5}}, \quad (5.2.38)$$

$$\frac{V_3}{V_{in}}=\frac{s^2\left(\frac{G_6}{G_8}-G_7\frac{C_1}{2C_3G_8}\right)+s\frac{G_7\left(C_5G_6+C_1G_4\right)}{2C_3C_5G_8}+\frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2+s\frac{G_7}{2C_3}+\frac{G_4G_7}{2C_3C_5}}. \quad (5.2.39)$$

In this case, for Eqs. (5.2.37) and (5.2.38) a non-traditional HP filter is obtained, the second being non-minimum phase. For $2C_3G_6 = C_1G_7$ the last one yields a non-traditional LP filter characteristic.

5.2.2.6 Case 6: $y_1 = 0$ ***and*** $y_2 = sC_2$

In a similar manner to the previous cases we obtain

$$\frac{V_1}{V_{in}} = \frac{-s^2 \dfrac{C_2}{2C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.40}$$

$$\frac{V_2}{V_{in}} = \frac{s^2 \dfrac{C_2}{2C_3} + s\dfrac{C_2G_4}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.41}$$

$$\frac{V_3}{V_{in}} = \frac{s^2\left(\dfrac{G_6}{G_8} - \dfrac{G_7C_2}{2C_3G_8}\right) + s\dfrac{G_7\left(C_5G_6 - C_2G_4\right)}{2C_3C_5G_8} + \dfrac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}. \tag{5.2.42}$$

The first of above equations gives a classical inverting HP filter characteristic. Although the second one does not seem to be a conventional filter transfer function, it behaves like a HP filter due to dominant *s*-terms. It is possible to obtain a non-traditional LP filter characteristic from the last one by the balance condition $2C_3G_6 = C_2G_7$. Further if we choose $C_5G_6 = C_2G_4$ then the filter becomes a traditional LP type. On the other hand if $C_5G_6 = C_2G_4$ whilst $2C_3G_6 > C_2G_7$, then Eq. (5.2.42) yields a BR filter.

5.2.2.7 Case 7: $y_1 = G_1$ ***and*** $y_2 = sC_2$

Again, we use Eqs. (5.2.19), (5.2.21) and (5.2.23) to derive

$$\frac{V_1}{V_{in}} = \frac{-s^2 \dfrac{C_2}{2C_5} + s\dfrac{G_1}{2C_5} + \dfrac{G_1G_7}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.43}$$

$$\frac{V_2}{V_{in}} = \frac{s^2\dfrac{C_2}{2C_3} + s\dfrac{C_5G_1 + C_2G_4}{2C_3C_5} - \dfrac{G_1G_4}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.44}$$

$$\frac{V_3}{V_{in}} = \frac{s^2(\dfrac{G_6}{G_8} - \dfrac{C_2G_7}{2C_3G_8}) + s\dfrac{G_7(C_5G_6 - C_5G_1 - C_2G_4)}{2C_3C_5G_8} + \dfrac{G_4G_6G_7 + G_1G_4G_7}{2C_3C_5G_8}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.45}$$

respectively. The first and second equations do not obey the forms of the traditional types of filter characteristics. The last equation behaves like a BR filter for the condition $C_5G_6 = C_5G_1 + C_2G_4$ and $2C_3G_6 > C_2G_7$, a non-traditional type LP filter for $2C_3G_6 = C_2G_7$, and an ordinary LP filter by the further condition $C_5G_6 = C_5G_1 + C_2G_4$.

5.2.2.8 Case 8: $y_1 = sC_1$ *and* $y_2 = G_2$

In a similar manner, we finally obtain

$$\frac{V_1}{V_{in}} = \frac{s^2\dfrac{C_1}{2C_5} + s\dfrac{C_1G_7 - C_3G_2}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.46}$$

$$\frac{V_2}{V_{in}} = \frac{s^2\dfrac{C_1}{2C_3} + s\dfrac{C_5G_2 - C_1G_4}{2C_3C_5} + \dfrac{G_2G_4}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}, \tag{5.2.47}$$

$$\frac{V_3}{V_{in}} = \frac{s^2\left(\dfrac{G_6}{G_8} - \dfrac{C_1G_7}{2C_3G_8}\right) + s\dfrac{C_5G_6G_7 + C_1G_4G_7 - C_5G_2G_7}{2C_3C_5G_8} + \dfrac{G_4G_6G_7 - G_2G_4G_7}{2C_3C_5G_8}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}. \tag{5.2.48}$$

With this choice, the first of the above equations yields a non-traditional HP filter; further if $C_1G_7 = C_3G_2$ this filter becomes a traditional type HP filter. Eq. (5.2.47) results in a BR transfer function if $C_5G_1 = C_1G_4$. Eq. (5.2.48) also yields a BR characteristic for $C_5(G_2 - G_6) = C_1G_4$, $G_6 < G_2$ and $2C_3G_6 > C_1G_7$; it yields a BP characteristic by the conditions $2C_3G_6 = C_1G_7$ and $G_2 = G_6$, and it yields a LP characteristic if $2C_3G_6 = C_1G_7$ and $C_5G_6 + C_1G_4 = C_5G_2$. If only the condition $2C_3G_6 = C_1G_7$ is satisfied the filter is non-traditional LP type, if only the condition $G_2 = G_6$ is satisfied the filter is a non-traditional HP type.

Hitherto, all combinational transfer functions with the fixed denominator are obtained. Having the same denominator, all these transfer functions have the same ideal and non-ideal ω_0 and Q as follows;

$$\omega = \sqrt{\frac{G_4G_7}{2C_3C_5}}, \qquad \omega_{NI} = \sqrt{\frac{G_4G_7}{C_3C_5(1+\alpha_{p_2}\alpha_{n_1})}}, \tag{5.2.49}$$

$$Q = \sqrt{\frac{2C_3G_4}{C_5G_7}}, \qquad Q_{NI} = \frac{\sqrt{C_3C_5G_4G_7(1+\alpha_{p_2}\alpha_{n_1})}}{C_3G_4(1-\alpha_{n_1}\alpha_{n_2})+C_5G_7}. \tag{5.2.50}$$

A summary of transfer functions realized in the above cases with the chosen conditions are classified in Table 5.1 according to the types of the filter characteristics. In this table all the filter types are traditional and all the transfer functions are minimum phase, otherwise stated.

Table 5.1 Realized filter functions with required conditions.

Filter	Case	Condition	Transfer Function
LP	1	$G_1 = G_2$	$\frac{V_1}{V_{in}} = \frac{\frac{G_1G_7}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
	2	Minimum phase (non-traditional)	$\frac{V_1}{V_{in}} = \frac{s\frac{G_1}{2C_5} + \frac{G_1G_7}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}},$
		Non-minimum phase (non-traditional)	$\frac{V_2}{V_{in}} = \frac{s\frac{G_1}{2C_3} - \frac{G_1G_4}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}},$
	3	(non-traditional)	$\frac{V_2}{V_{in}} = \frac{s\frac{G_2}{2C_3} + \frac{G_2G_4}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}},$
	4	$C_5G_6 + C_1G_4 = C_2G_4$ and $2C_3G_6 = G_7(C_1 + C_2)$	$\frac{V_3}{V_{in}} = \frac{\frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}},$
	5	$2C_3G_6 = C_1G_7$ (non-traditional)	$\frac{V_3}{V_{in}} = \frac{s\frac{G_7(C_5G_6 + C_1G_4)}{2C_3C_5G_8} + \frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}.$
	6	$2C_3G_6 = C_2G_7$ (non-traditional)	$\frac{V_3}{V_{in}} = \frac{s\frac{G_7(C_5G_6 - C_2G_4)}{2C_3C_5G_8} + \frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}.$
		$2C_3G_6 = C_2G_7$ and $C_5G_6 = C_2G_4$	$\frac{V_3}{V_{in}} = \frac{\frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$

Table 5.1 (continued) Realized filter functions with required conditions.

Filter	Case	Condition	Transfer Function
LP	7	$2C_3G_6 = C_2G_7$ (non-traditional)	$\frac{V_3}{V_{in}} = \frac{s\frac{G_7(C_5G_6 - C_5G_1 - C_2G_4)}{2C_3C_5G_8} + \frac{G_4G_6G_7 + G_1G_4G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$,
		$2C_3G_6 = C_2G_7$ and $C_5G_6 = C_5G_1 + C_2G_4$	$\frac{V_3}{V_{in}} = \frac{\frac{G_4G_6G_7 + G_1G_4G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$,
	8	$2C_3G_6 = C_1G_7$ (non-traditional)	$\frac{V_3}{V_{in}} = \frac{s\frac{C_5G_6G_7 + C_1G_4G_7 - C_5G_2G_7}{2C_3C_5G_8} + \frac{G_4G_6G_7 - G_2G_4G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$.
		$2C_3G_6 = C_1G_7$ and $C_5G_6 + C_1G_4 = C_5G_2$	$\frac{V_3}{V_{in}} = \frac{\frac{G_4G_6G_7 - G_2G_4G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$.
BP	1	$G_1 = G_2$	$\frac{V_2}{V_{in}} = \frac{s\frac{G_2}{C_3}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
	3		$\frac{V_1}{V_{in}} = \frac{-s\frac{G_2}{2C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
	4	$C_1 = C_2$	$\frac{V_1}{V_{in}} = \frac{s\frac{C_1G_7}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
	8	$2C_3G_6 = C_1G_7$ and $G_2 = G_6$	$\frac{V_3}{V_{in}} = \frac{s\frac{C_1G_4G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$

Table 5.1 (continued) Realized filter functions with required conditions.

Filter	Case	Condition	Transfer Function
HP	3	$G_2 = G_6$	$\frac{V_3}{V_{in}} = \frac{s^2 \frac{G_6}{G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}}$
	4	$C_1 = C_2$	$\frac{V_2}{V_{in}} = \frac{s^2 \frac{C_1 + C_2}{2C_3}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}}$
		Min. or non-min. phase (non-traditional)	$\frac{V_1}{V_{in}} = \frac{s^2 \frac{C_1 - C_2}{2C_5} + s\frac{C_1 G_7}{2C_3 C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}},$
		Min. or non-min. phase (non-traditional)	$\frac{V_2}{V_{in}} = \frac{s^2 \frac{C_1 + C_2}{2C_3} + s\frac{G_4 (C_2 - C_1)}{2C_3 C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}},$
	5	(non-traditional)	$\frac{V_1}{V_{in}} = \frac{s^2 \frac{C_1}{2C_5} + s\frac{C_1 G_7}{2C_3 C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}},$
		Non-minimum phase (non-traditional)	$\frac{V_2}{V_{in}} = \frac{s^2 \frac{C_1}{2C_3} - s\frac{G_4 C_1}{2C_3 C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}},$
	6		$\frac{V_1}{V_{in}} = \frac{-s^2 \frac{C_2}{2C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}}$
		(non-traditional)	$\frac{V_2}{V_{in}} = \frac{s^2 \frac{C_2}{2C_3} + s\frac{C_2 G_4}{2C_3 C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4 G_7}{2C_3 C_5}}$

Table 5.1 (continued) Realized filter functions with required conditions.

Filter	Case	Condition	Transfer Function
HP	8	(non-traditional)	$\frac{V_1}{V_{in}} = \frac{s^2 \frac{C_1}{2C_5} + s\frac{C_1G_7 - C_3G_2}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$,
		$C_1G_7 = C_3G_2$	$\frac{V_1}{V_{in}} = \frac{s^2 \frac{C_1}{2C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
		$G_2 = G_6$ (non-traditional)	$\frac{V_3}{V_{in}} = \frac{s^2\left(\frac{G_6}{G_8} - \frac{C_1G_7}{2C_3G_8}\right) + s\frac{C_5G_6G_7 + C_1G_4G_7 - C_5G_2G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$.
AP	1	$G_1 = G_2 = G_6 = G_8$	$\frac{V_3}{V_{in}} = \frac{s^2 - s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
BR	1	$G_1 + G_2 = G_6$	$\frac{V_3}{V_{in}} = \frac{s^2\frac{G_6}{G_8} + \frac{G_7(G_4G_6 + G_1G_4 - G_2G_4)}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
	2	$G_1 = G_6$	$\frac{V_3}{V_{in}} = \frac{s^2\frac{G_6}{G_8} + \frac{G_1G_4G_7}{C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$
	4	$C_5G_6 + C_1G_4 = C_2G_4$ and $2C_3G_6 > G_7(C_1 + C_2)$	$\frac{V_3}{V_{in}} = \frac{s^2\left(\frac{G_6}{G_8} - G_7\frac{C_1 + C_2}{2C_3G_8}\right) + \frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$,
	6	$C_5G_6 = C_2G_4$ and $2C_3G_6 > C_2G_7$	$\frac{V_3}{V_{in}} = \frac{s^2\left(\frac{G_6}{G_8} - \frac{G_7C_2}{2C_3G_8}\right) + \frac{G_4G_6G_7}{2C_3C_5G_8}}{s^2 + s\frac{G_7}{2C_3} + \frac{G_4G_7}{2C_3C_5}}$.

Table 5.1 (continued) Realized filter functions with required conditions.

Filter	Case	Condition	Transfer Function
BR	7	$C_5G_6 = C_5G_1 + C_2G_4$ and $2C_3G_6 > C_2G_7$	$\dfrac{V_3}{V_{in}} = \dfrac{s^2(\dfrac{G_6}{G_8} - \dfrac{C_2G_7}{2C_3G_8}) + \dfrac{G_4G_6G_7 + G_1G_4G_7}{2C_3C_5G_8}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}$,
	8	$C_5G_2 = C_1G_4$	$\dfrac{V_2}{V_{in}} = \dfrac{s^2\dfrac{C_1}{2C_3} + \dfrac{G_2G_4}{2C_3C_5}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}$
		$C_5(G_2 - G_6) = C_1G_4$, $G_6 < G_2$ and $2C_3G_6 > C_1G_7$	$\dfrac{V_3}{V_{in}} = \dfrac{s^2\left(\dfrac{G_6}{G_8} - \dfrac{C_1G_7}{2C_3G_8}\right) + \dfrac{G_4G_6G_7 - G_2G_4G_7}{2C_3C_5G_8}}{s^2 + s\dfrac{G_7}{2C_3} + \dfrac{G_4G_7}{2C_3C_5}}$.

5.2.3 A Universal Filter Topology

A universal filter topology using minimum number of capacitors (two virtually grounded capacitors $y_3 = sC_3$, $y_5 = sC_5$) and six resistors ($y_1 = 1/R_1$, $y_2 = 1/R_2$, $y_4 = 1/R_4$, $y_6 = 1/R_6$, $y_7 = 1/R_7$, $y_8 = 1/R_8$) can be obtained from Fig. 5.1. By using Cases 1 and 3, the type of the filter response of this topology can be controlled easily by removing one or the other of the two resistors ($y_1 = 1/R_1$ and $y_2 = 1/R_2$). The resulting traditional second order LP, BP, HP, AP and BR transfer functions are extracted from Table 5.1 and re-tabled in Table 5.2 with the required conditions. In certain cases, this general topology using two capacitors and five or six resistors has less number of active and/or passive components than the proposed three or more CDBA-based topologies in the literature (Sagbas and Koksal, 2005), (Koksal and Sagbas, 2007), (Tangsrirat and Surakampontorn, 2005). For example, at least four CDBAs are required for the AP filter realization in (Sagbas and Koksal, 2005) and (Koksal and Sagbas, 2007).

Table 5.2 Universal filter topology transfer functions.

Filter	Output	Condition	Transfer Function
LP	V_1	$R_1 = R_2$	$\dfrac{V_1}{V_{in}} = \dfrac{\dfrac{1}{2C_3C_5R_1R_7}}{s^2 + s\dfrac{1}{2C_3R_7} + \dfrac{1}{2C_3C_5R_4R_7}}$
BP	V_2	$R_1 = R_2$	$\dfrac{V_2}{V_{in}} = \dfrac{s\dfrac{1}{C_3R_2}}{s^2 + s\dfrac{1}{2C_3R_7} + \dfrac{1}{2C_3C_5R_4R_7}}$
HP	V_3	$R_1 = 0$ and $R_2 = R_6$	$\dfrac{V_3}{V_{in}} = \dfrac{s^2\dfrac{R_8}{R_6}}{s^2 + s\dfrac{1}{2C_3R_7} + \dfrac{1}{2C_3C_5R_4R_7}}$
AP	V_3	$R_1 = R_2 = R_6 = R_8$	$\dfrac{V_3}{V_{in}} = \dfrac{s^2 - s\dfrac{1}{2C_3R_7} + \dfrac{1}{2C_3C_5R_4R_7}}{s^2 + s\dfrac{1}{2C_3R_7} + \dfrac{1}{2C_3C_5R_4R_7}}$
BR	V_3	$R_2 = 0$ and $R_1 = R_6$	$\dfrac{V_3}{V_{in}} = \dfrac{s^2\dfrac{R_8}{R_6} + \dfrac{R_8}{C_3C_5R_1R_4R_7}}{s^2 + s\dfrac{1}{2C_3R_7} + \dfrac{1}{2C_3C_5R_4R_7}}$

5.3 SENSITIVITY ANALYSIS

The ideal sensitivities of ω_0 and Q of the filter obtained in this chapter with respect to passive components are calculated from Eqs. (5.2.49) and (5.2.50) using the formulation shown in Appendix A (Sagbas, 2004) as follows;

$$S_{G_1}^{\omega} = S_{G_2}^{\omega} = S_{C_1}^{\omega} = S_{C_2}^{\omega} = S_{G_6}^{\omega} = S_{G_8}^{\omega} = 0, \tag{5.3.1}$$

$$S_{G_4}^{\omega} = S_{G_7}^{\omega} = -S_{C_3}^{\omega} = -S_{C_5}^{\omega} = \frac{1}{2}, \tag{5.3.2}$$

$$S_{G_1}^{Q} = S_{G_2}^{Q} = S_{C_1}^{Q} = S_{C_2}^{Q} = S_{G_6}^{Q} = S_{G_8}^{Q} = 0, \tag{5.3.3}$$

$$S_{C_3}^{Q} = S_{G_4}^{Q} = -S_{C_5}^{Q} = -S_{G_7}^{Q} = \frac{1}{2}, \tag{5.3.4}$$

From the above calculations, it can be seen that all sensitivities are zero or smaller than 1 in magnitude. In similar manner, the non-ideal sensitivities can be found as

$$S_{G_1}^{\omega_{NI}} = S_{G_2}^{\omega_{NI}} = S_{C_1}^{\omega_{NI}} = S_{C_2}^{\omega_{NI}} = S_{G_6}^{\omega_{NI}} = S_{G_8}^{\omega_{NI}} = 0, \tag{5.3.5}$$

$$S_{\alpha_{p1}}^{\omega_{NI}} = S_{\alpha_{p3}}^{\omega_{NI}} = S_{\alpha_{n2}}^{\omega_{NI}} = S_{\alpha_{n3}}^{\omega_{NI}} = S_{\beta_1}^{\omega_{NI}} = S_{\beta_2}^{\omega_{NI}} = S_{\beta_3}^{\omega_{NI}} = 0, \tag{5.3.6}$$

$$S_{G_4}^{\omega_{NI}} = S_{G_7}^{\omega_{NI}} = -S_{C_3}^{\omega_{NI}} = -S_{C_5}^{\omega_{NI}} = \frac{1}{2}, \tag{5.3.7}$$

$$S_{\alpha_{p2}}^{\omega_{NI}} = \frac{1}{2}\frac{\alpha_{p2}}{\alpha_{p_2}\alpha_{n_1}+1}, \tag{5.3.8}$$

$$S_{\alpha_{n_1}}^{\omega_{NI}} = \frac{1}{2}\frac{\alpha_{n_1}}{\alpha_{p_2}\alpha_{n_1}+1}, \tag{5.3.9}$$

$$S_{G_1}^{Q_{NI}} = S_{G_2}^{Q_{NI}} = S_{C_1}^{Q_{NI}} = S_{C_2}^{Q_{NI}} = S_{G_6}^{Q_{NI}} = S_{G_8}^{Q_{NI}} = 0, \tag{5.3.10}$$

$$S_{\alpha_{p1}}^{Q_{NI}} = S_{\alpha_{p3}}^{Q_{NI}} = S_{\alpha_{n2}}^{Q_{NI}} = S_{\alpha_{n3}}^{Q_{NI}} = S_{\beta_1}^{Q_{NI}} = S_{\beta_2}^{Q_{NI}} = S_{\beta_3}^{Q_{NI}} = 0, \tag{5.3.11}$$

$$S_{C_3}^{Q_{NI}} = S_{G_4}^{Q_{NI}} = \frac{1}{2} - \frac{C_3G_4\left(1-\alpha_{n_1}\alpha_{n_2}\right)}{C_3G_4\left(1-\alpha_{n_1}\alpha_{n_2}\right)+C_5G_7}, \tag{5.3.12 }$$

$$S_{C_5}^{Q_{NI}} = S_{G_7}^{Q_{NI}} = \frac{1}{2} - \frac{C_5G_7}{C_3G_4\left(1-\alpha_{n_1}\alpha_{n_2}\right)+C_5G_7}, \tag{5.3.13}$$

$$S_{\alpha_{p2}}^{Q_{NI}} = \frac{1}{2}\frac{\alpha_{p_2}\alpha_{n_1}}{\alpha_{p_2}\alpha_{n_1}+1}, \tag{5.3.14}$$

$$S_{\alpha_{n_1}}^{Q_{NI}} = \frac{1}{2}\frac{\alpha_{p_2}\alpha_{n_1}}{\alpha_{p_2}\alpha_{n_1}+1} + \frac{\alpha_{n_1}\alpha_{n_2}C_3G_4\left(\sqrt{\alpha_{p_2}\alpha_{n_1}+1}\right)}{C_3G_4\left(1-\alpha_{n_1}\alpha_{n_2}\right)+C_5G_7}, \tag{5.3.15}$$

$$S_{\alpha_{n_2}}^{Q_{NI}} = \frac{\alpha_{n_1}\alpha_{n_2}C_3G_4}{C_3G_4\left(1-\alpha_{n_1}\alpha_{n_2}\right)+C_5G_7}. \tag{5.3.16}$$

Since α_{p_1}, α_{p_2}, α_{p_3}, α_{n_1}, α_{n_2}, α_{n_3}, β_1, β_2, and β_3 are equal to 1 approximately, the ideal and non-ideal sensitivities except $S_{\alpha_{n2}}^{Q_{NI}}$ will be smaller than $1/2$ in magnitude. And $S_{\alpha_{n2}}^{Q_{NI}}$ can be kept within reasonable limits by choosing the passive component values appropriately.

Nevertheless, ideal and non-ideal calculations demonstrate a good sensitivity performance with respect to passive components.

5.4 SIMULATION RESULTS

Some of filter topologies proposed in this chapter are simulated using PSpice. The simulations are carried out using the model parameters of Tübitak Yital 1.5µ and CMOS CDBA implementation proposed by Ayten (Ayten, 2003). Simulation bias voltages are chosen as $\mathrm{V_{cc}} = 3\,\mathrm{V}$ and $\mathrm{V_{ee}} = -3\,\mathrm{V}$. Design parameters are chosen appropriately in order to obtain $\omega_0 = 40\,kHz$ for all filters so that a comparison of filter characteristics would be done easily.

5.4.1 LP Filters

The analyzed general circuit configuration in Fig. 5.1 has given us five traditional and eight non-traditional LP filter characteristics. Only four of the mentioned LP filters, namely Case 1 (traditional), Case 2 (minimum phase non-traditional), Case 6 (traditional) and Case 7 (non-traditional) are simulated in order to give a general idea about the simulation results. If we choose design parameters as in Table 5.3, we obtain the simulation results shown in Fig. 5.2.

Table 5.3 LP filter design parameters.

	R_1 or C_1	R_2 or C_2	C_3	R_4	C_5	R_6	R_7	R_8
Case-1 (V_1)	$2k\Omega$	$2k\Omega$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-2 (V_1)	$2k\Omega$	0	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-6 (V_3)	0	$2nF$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-7 (V_3)	$4k\Omega$	$2nF$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$1.33k\Omega$

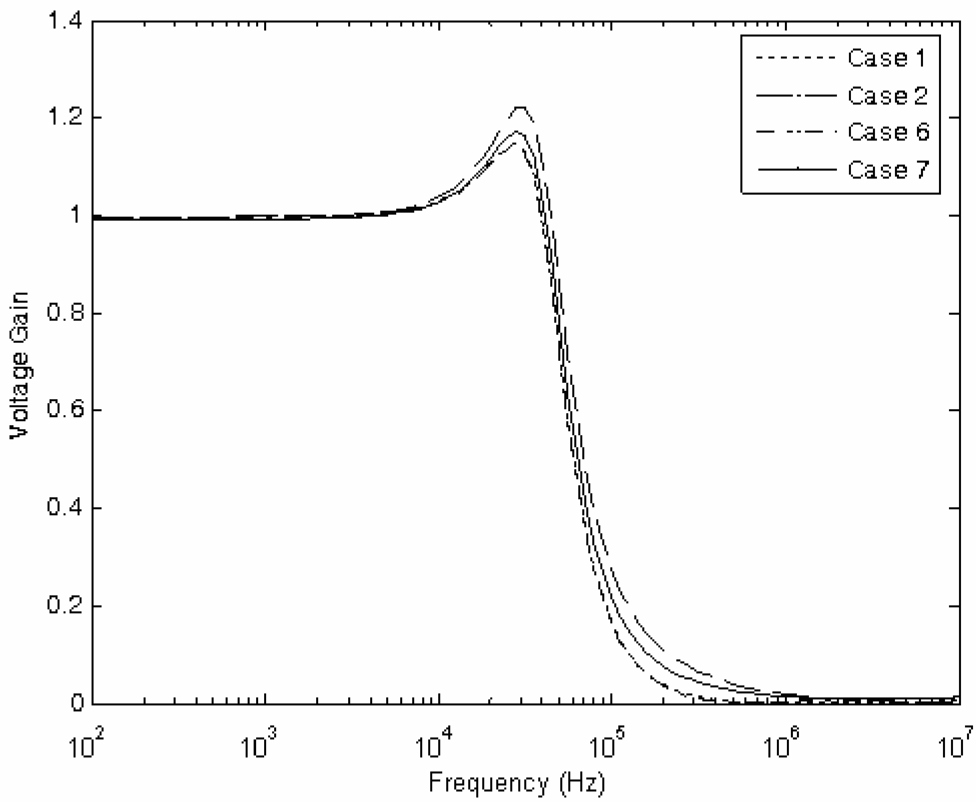

Figure 5.2 Simulation results of LP filters.

Having the same denominator polynomial, all these LP filters are of Chebyshev type and the simulation results well demonstrate this property in spite of non-idealities of CDBA.

5.4.2 BP Filters

Four BP filter realizations are possible for the configuration in concern as given in Table 5.1. BP simulation results for all of the four cases are shown in Fig. 5.3 by using the design parameters given in Table 5.4.

Table 5.4 BP filter design parameters.

	R_1 or C_1	R_2 or C_2	C_3	R_4	C_5	R_6	R_7	R_8
Case-1 (V_2)	$4k\Omega$	$4k\Omega$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-3 (V_1)	0	$1k\Omega$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-4 (V_1)	$2nF$	$2nF$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-8 (V_3)	$2nF$	$2k\Omega$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$

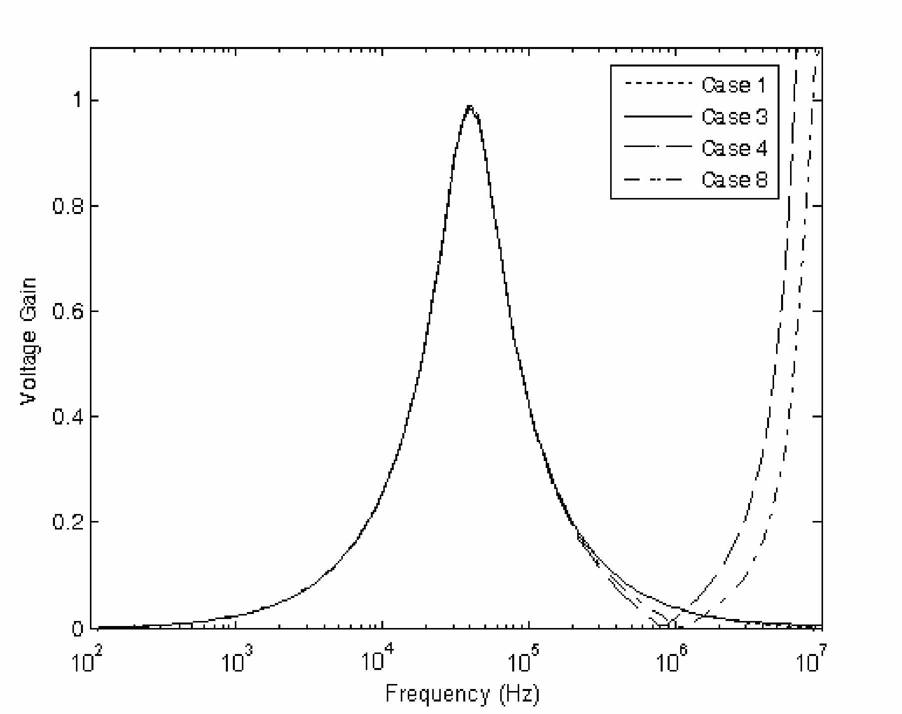

Figure 5.3 Simulation results of BP filters.

Considering Fig. 5.3, Case 1 and Case 3 show good BP filter characteristics with a quality factor of 1. Simulation results of Case 4 and Case 8 also demonstrate good BP filter characteristics up to 1 MHz with quality factor of 1. After 1 MHz, effects of parasitic capacitances increase and the balancing conditions canceling the coefficient of

s^2 term in the numerator of Eqs. (5.2.34) and (5.2.48) are not satisfied, resultantly gain increases at high frequencies.

5.4.3 HP Filters

The derived HP transfer functions are simulated for the traditional cases only by the design parameters given in Table 5.5. Fig. 5.4 shows simulation results.

Table 5.5 HP filter design parameters.

	R_1 or C_1	R_2 or C_2	C_3	R_4	C_5	R_6	R_7	R_8
Case-3 (V_3)	0	$2k\Omega$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-4 (V_2)	1 nF	1 nF	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-6 (V_1)	0	$4nF$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-8 (V_1)	$2nF$	$1k\Omega$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$

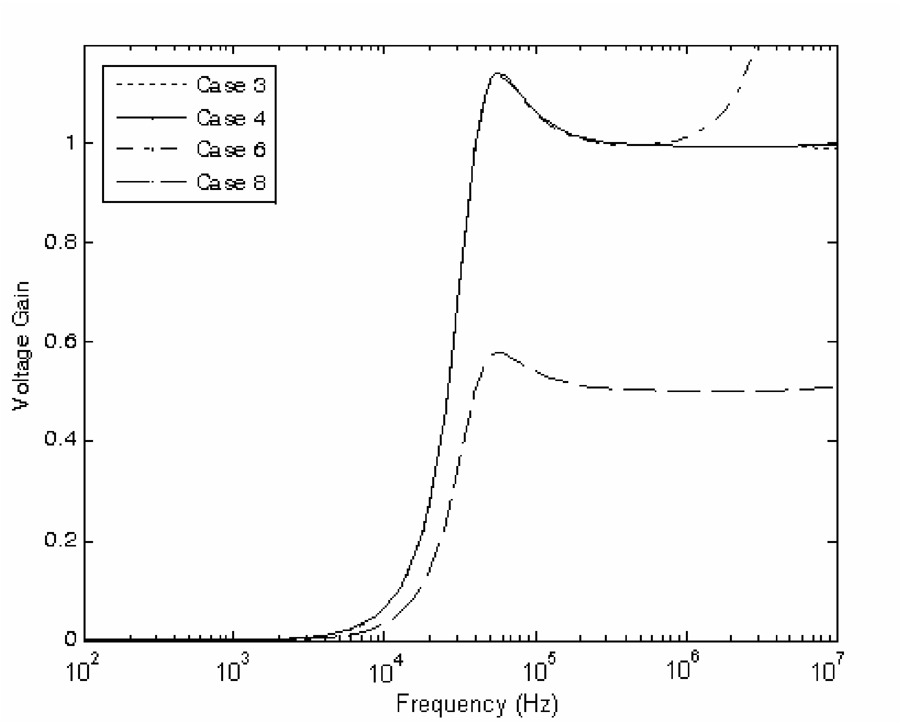

Figure 5.4 Simulation results of HP filters.

The fact that almost the same transfer functions are obtained for both Case 3 and Case 4 is due to the design parameters given in Table 5.5 give the identical transfer functions for these two cases. Case 6 shows an expected characteristic up to 1 MHz, however an unexpected rise occurs after this frequency and the gain increases to a peak value which is slightly greater than 4 at about 10 MHz. Although the deviations between the theoretically expected and simulation results are expected at high frequencies due to non-ideal affects especially parasitic capacitances of CDBA, the resonant-like peak at 10 MHz could not be explained for this case. Low frequency gain of $1/2$ for Case 8 is expected from Eq. (5.2.46) with the component values shown in Table 5.5.

5.4.4 AP Filter

AP filter characteristic is obtained only from Case 1 for the design parameters given below and PSpice simulation and theoretical results are shown in Fig. 5.5.

Table 5.6 AP filter design parameters.

	R_1 or C_1	R_2 or C_2	C_3	R_4	C_5	R_6	R_7	R_8
Case-1 (V_3)	$2\,k\Omega$	$2\,k\Omega$	$1\,nF$	$2\,k\Omega$	$2\,nF$	$2\,k\Omega$	$2\,k\Omega$	$2\,k\Omega$

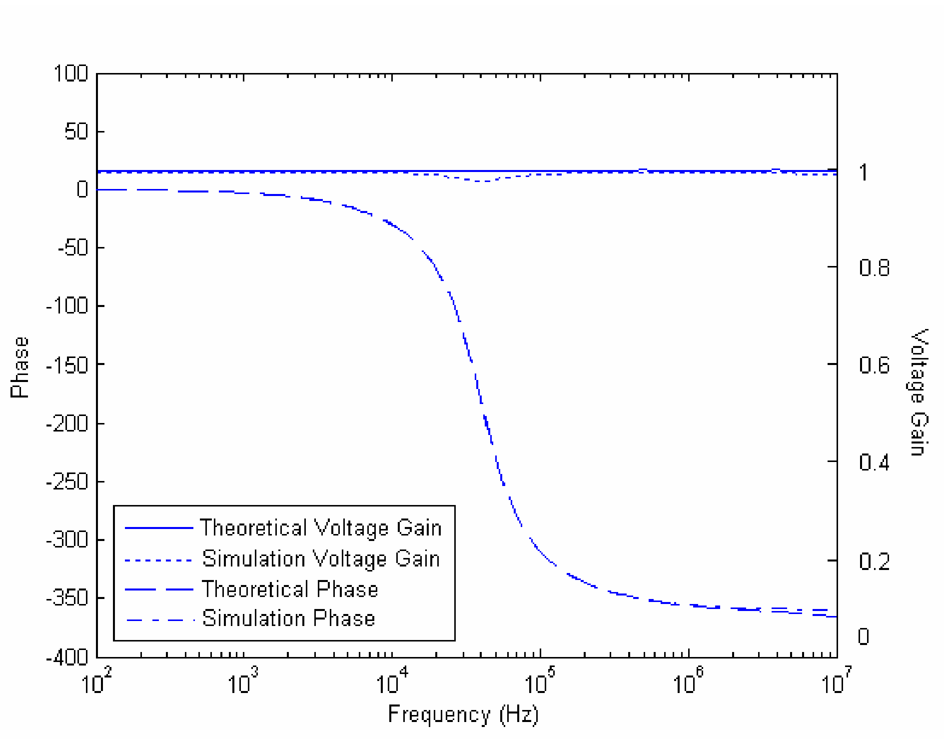

Figure 5.5 Simulation results of AP filter.

Non-ideal effects result in a very small difference between theoretical and simulation results for gain. Phase simulation characteristic is also in good agreement with theoretical response till 10 MHz where parasitic capacitance affects start.

5.4.5 BR Filters

Although seven BR filter functions are derived in Table 5.1, three of the least condition-based transfer functions are simulated with design parameters given in Table 5.7. Simulation results are demonstrated in Fig. 5.6.

Table 5.7 BR filter design parameters.

	R_1 or C_1	R_2 or C_2	C_3	R_4	C_5	R_6	R_7	R_8
Case-1 (V_3)	$4k\Omega$	$4k\Omega$	$1\,nF$	$2k\Omega$	$2\,nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-2 (V_3)	$2k\Omega$	0	$1\,nF$	$2k\Omega$	$2\,nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
Case-8 (V_2)	$2\,nF$	$2k\Omega$	$1\,nF$	$2k\Omega$	$2\,nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$

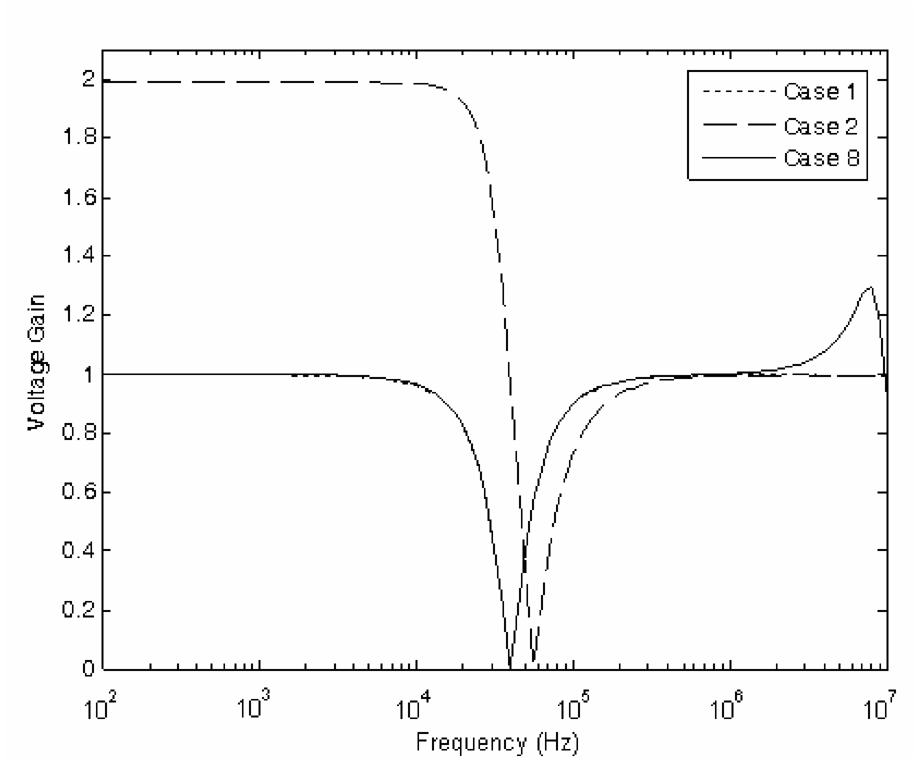

Figure 5.6 Simulation results of BR filters.

Case 1 gives a perfect result when compared with the theoretically expected one. Double gain at low frequencies for Case 2 occurs from the ratio of constant terms at nominator and denominator. An increase after 1 MHz and a reach to a peak at about 10 MHz for Case 8 exhibits similar behavior with the HP filter of Case 6 in Section 5.4.3 (Fig. 5.4). Another peculiar deviation occurs at the notch frequency for Case 2 which may be due to the parasitic capacitance effects and tracking errors of CDBA.

5.4.6 The Universal Filter

The universal filter topology described in Section 5.2.3 and summarized in Table 5.2 is tested by using the component values shown in Table 5.8 for all types of filter characteristics. All of the filters in this table have the same ω_0 of 40 kHz. PSpice simulation results are shown in Fig. 5.7.

Table 5.8 Universal filter design parameters.

Filter	Output	R_1	R_2	C_3	R_4	C_5	R_6	R_7	R_8
LP	V_1	$2k\Omega$	$2k\Omega$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
BP	V_2	$4k\Omega$	$4k\Omega$	$1\ nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
HP	V_3	0	$2k\Omega$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
AP	V_3	$2k\Omega$	$2k\Omega$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$
BR	V_3	$4k\Omega$	$4k\Omega$	$1nF$	$2k\Omega$	$2nF$	$2k\Omega$	$2k\Omega$	$2k\Omega$

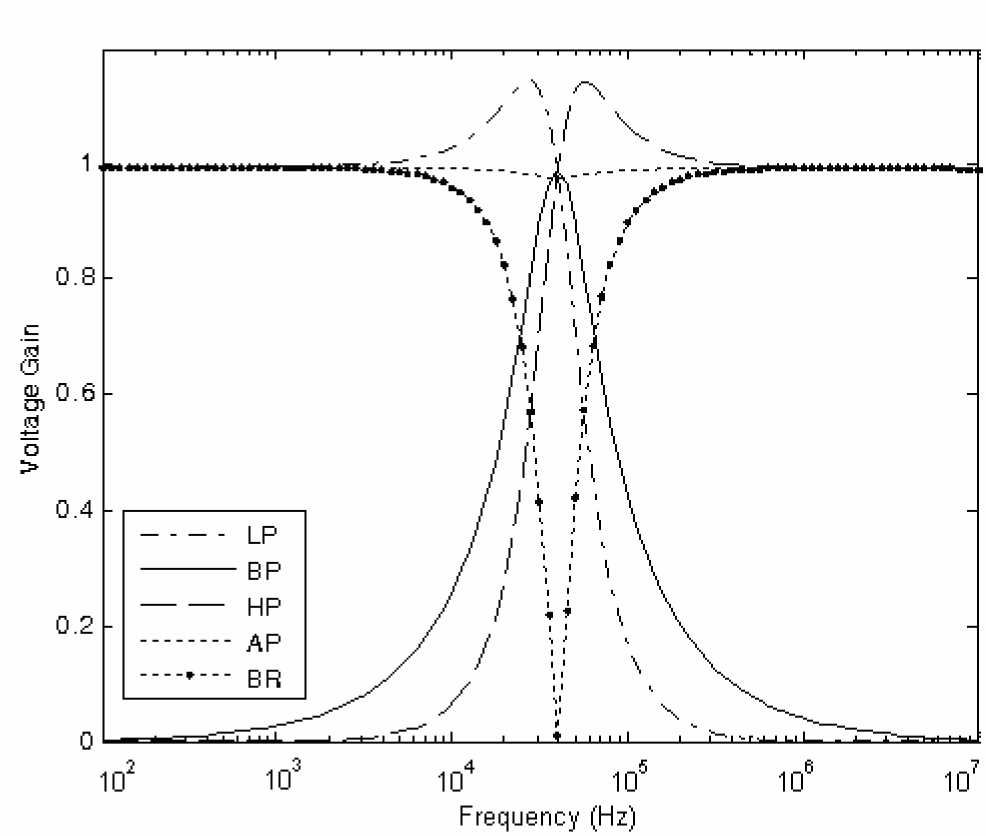

Figure 5.7 Filter characteristics of proposed universal filter.

Since the balancing conditions to obtain the desired type of filter characteristics are fairly simple and not stringent as to include the capacitances and resistances together, all the results behave quite well with respect to expected characteristics.

CHAPTER 6

CONCLUSIONS

This thesis presents design of active filters using a new active element called CDBA. A single CC-CDBA-based, a single CDBA-based and three CDBA-based filter configurations are proposed through the thesis.

In the third chapter, a single active device-based three new electronically tunable VM second-order universal filters are proposed. Each of the proposed filters realizes three basic second-order filter functions simultaneously: LP, BP, and HP. Besides they use two capacitors, one resistor and a single CC-CDBA. These are minimum number of active and passive elements in comparison to previously reported configurations in the literature (Abuelma'atti, 2000), (Salama et al., 2001), (Minae et al., 2001), (Ozcan et al., 2002), (Acar and Sedef, 2003), (Keskin, 2005), (Kilinc and Cam, 2004). The validity of the proposed filters is verified through PSpice simulations. Sensitivities of given filters are calculated below 0.5 in magnitude. ω_0 and Q can be adjusted electronically without changing the values of the passive components, and ω_0 and ω_0 / Q enjoy independent electronic tunability.

In chapter four, a new second-order multi-mode multifunction filter configuration is presented. Operation mode of the filter is decided by admittance choices for the given configuration. The configuration uses a single CDBA, four capacitors and four/five resistors which uses minimum active and/or passive elements in comparison to previously reported multi-mode configurations (Koksal and Sagbas, 2007) (Sagbas and Koksal, 2005). All of the characteristics of basic filters namely, LP, BP and HP filter are obtained in VM, CM, TIM and TAM except HP filter for TIM. The validity of proposed filters is verified through PSpice simulations. Ideal and non-ideal sensitivities of ω_0 and

Q are obtained below 1 in magnitude. Furthermore, a new three CFA-based CDBA implementation is presented for experimental purposes.

In chapter five, a SIMO filter configuration that realizes all five basic filter transfer characteristics is given. Eight topologies of proposed configuration are investigated by changing two admittances. PSpice simulations of given topologies are given. All the sensitivities are limited by 0.5 except the one in Eq. (5.3.15) which can be kept within reasonable limits by choosing the passive component values appropriately. Proposed configuration offers the following advantages: (i) One passive element is reduced from the previous work (Tangsrirat and Surakampontorn, 2005); (ii) A single topology using six resistors, two capacitors and three CDBAs realizes all five basic filter functions by altering two resistors only; (iii) Although, *w*-terminals are conventionally used for voltage outputs, *z*-terminals can be used as voltage outputs in this configuration, and the voltage buffer stage in CDBAs can be eliminated for a fully integrated circuit implementation (iv) Considering configurations using three CDBAs, a minimum number of active and/or passive elements are used in the given topologies for certain cases.

APPENDIX A

SENSITIVITY ANALYSIS

We define the measure of the change Δy in some performance characteristic y, resulting from a change Δx in a network parameter x to be the sensitivity of y with respect to x, given by (Johnson, 1976)

$$S_x^y = \underrightarrow{\frac{\Delta y / y}{\Delta x / x}} = \frac{x}{y}\frac{\Delta y}{\Delta x} \tag{A.1}$$

Thus the changes in x and y have been normalized. That is, Δy is divided by y and Δx by x so that S_x^y is a ratio of normalized changes or percentages. For example, if $S_x^y = 0.5$, then a 2 % changes in x will cause a 1 % changes in y.

Eq. (A.1) may be put in a more useful from by considering the Taylor's series,

$$y + \Delta y = y + \frac{\partial y}{\partial x_1}\Delta x_1 + \frac{\partial y}{\partial x_2}\Delta x_2 + \cdots + \frac{\partial y}{\partial x_n}\Delta x_n + \textit{high order terms} \tag{A.2}$$

where y is a function of $x_1, x_2, \cdots, x_n$. Neglecting the higher order terms and letting x_i vary while the order x_j, $j \neq i$, are fixed, we have

$$\Delta y \approx \frac{\partial y}{\partial x_i}\Delta x_i \tag{A.3}$$

so that by Eq. (A.1), $S_{x_i}^y$ can be approximately written as

$$S_{x_i}^{y} = \frac{x_i}{y}\frac{\partial y}{\partial x_i} \tag{A.4}$$

Theoretically Eq. (A.4) is only valid for small changes, but as a practical matter the sensitivity function is adequate for changes of network parameters up to and some times 10 percent.

As an example, consider the overall transfer function H of n cascaded sub-networks,

$$H = \prod_{i=1}^{n} H_i \tag{A.5}$$

The functions H_i are the transfer functions of the sub-networks. The sensitivity of H with respect to a sub-network function H_i is given by

$$S_{H_j}^{H} = \frac{H_j}{H}\frac{\partial H}{\partial H_j} = \frac{H_j}{\prod_{i=1}^{n} H_i}\prod_{i=1}^{n} H_i = 1 \tag{A.6}$$

Thus a relative change in H_j results in the same relative change in H.

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