Single CFTA Based Current-Mode Universal Biquad Filter

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Abstract: This paper introduces a new current-mode (CM) universal biquad filter structure with optimum number of active and passive elements. In the design, the proposed circuit uses a single active element namely, current follower trans-conductance amplifier (CFTA) and two grounded capacitors as passive elements. The main feature of the proposed circuit is that it can realize all five standard filtering functions such as low pass (LP), band pass (BP), high pass (HP), band stop (BS) and all pass (AP) responses across an explicit high impedance output terminal through the appropriate selection of three inputs. In addition, the same circuit is also capable to simultaneously realize three filtering functions (LP, BP and HP) by the use of single current input signal. Moreover, the proposed structure is suited for low voltage, low power operations and offers the feature of electronic tunability of pole-frequency and quality factor. Further to extend the utility of the proposed circuit block higher order current-mode filters are also realized through direct cascading. A detailed non-ideal and parasitic study is also included. The performance of the circuits has been examined using standard 0.25 μm CMOS parameters from TSMC.

مكبر أحادي ناقل للتوصيلية تابع للتيار لمرشح الهوائيات العالمي معتمداً علي حالة التيار

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الملخص: تعرض هذه الورقة نمط تياري عالمي لبنية مرشح كهربائي هوائي جديد مع الاستخدام الأمثل للعناصر النشطة والخاملة. وفي التصميم، فإن مجمع الدائرة الكهربائية يستخدم عنصر مفرد فعال (مضخم ناقلية عبور تابع التيار) مع مكثفين كهربائيين أرضيين كعناصر خاملة. إن فائدة الدائرة المقترحة تبدأ مع التيار المدخل المفرد، والتي يمكن من خلالها تحقيق استجابات تمرير بطيئة، ومتحدة، وعالية، على التوالي. علاوة على ذلك، مع التيارات المدخلة الثلاثية، فإنه يمكن أيضا تحقيق جميع وظائف الترشيح القياسية الخمسة عبر محطة انتاج محددة عالية المقاومة. وتناسب البنية المقترحة الجهد المنخفض، والعمليات المنخفضة الطاقة، وتقدم ميزة قابلية التوليف الإلكترونية لقطب التردد وعامل الجودة. بالإضافة الى اطالة المنفعة من مجمع الدائرة المقترحة، فإن المرشحات العالية الرتبة تتحقق من خلال التوصيل التعاقبي المباشر. كما تم بالتفصيل احتواء دراسة غير المثالية والطفيلية. وقد تم فحص أداء الدوائر الكهربائية باستخدام متغيرات معيار ٠,٢٥ ميكرون.

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1. Introduction

Analog circuit designers have paid a significant accent on the study and synthesis of current-mode (CM) continuous-time (CT) filter structures using variants of current conveyors suited for high performance analog signal processing applications (Toumazou *et al*. 1990; Kacar and Kuntman 2011). In the last few decades, circuit designers primarily focused on designing low cost, low supply voltage operable simple filter structures realizing all the standard filtering responses. These filter structures are termed as universal filters. Further, in the perspective of less chip area, cost reduction and lower average power consumption, the concluding remarks state that the circuit topology must be intended for an optimum use of active and passive elements in the biquad design (Satansup and Tangsrirat 2014). Another concern pertinent to the integrated circuit realization point of view is the electronic tuning aspects of filter parameters used for the fine adjustment of filter parameters against large manufacturing tolerances, which can also be used to compensate for deviations in the circuit performance parameters due to process tolerance, port parasitic, temperature drift, imperfections and aging *etc*. (Karybakas and Papazoglou 1999; Metin 2007). In the process, enormous papers on the synthesis of current-mode biquad filter circuits having one or more active elements are reported in the available literature (Keskin *et al*. 2006; Tomar *et al*. 2014a). The circuits reported in references are capable of realizing all the five standard filtering functions in current-mode either simultaneously (Cicekoglu 2001; Maheshwari and Khan 2004; Maheshwari and Khan 2005) or one at a time (Singh *et al*. 2014; Tomar *et al*. 2014b; Singh *et al*. 2012) or both way (Wang *et al*. 2001) without changing the circuit structure. Beside it, most of the circuit structure except few (Cicekoglu 2001; Wang *et al*. 2001) is resister less and canonical. However, all these circuits require more than one active element to realize filtering functions. With a view to being adaptable towards the simplicity, low cost, lower power consumption and space saving of the circuits, it is beneficial to design the biquad filter by the use of single active element and minimum number of passive elements. Therefore, we have focused here only on the single active element based current-mode filters from the available literature (Arsian *et al*. 2006; Prasad *et al*. 2009) for the comparative purpose and hence, the detail features of all

single active elements based current-mode biquad filters are compared and summarized in Table 1. Some of the circuits among the current-mode filters based on single active element as discussed in Table 1 are configured as single input three output (SITO) (Mangkalakecree *et al*. 2009; Prasad *et al*. 2009) and remaining are configured as three input single output (TISO) (Arsian *et al*. 2006; Satansup *et al*. 2014). The SITO filters can realize at the most three filtering functions simultaneously whereas TISO filters can realize all five filtering functions but one function explicitly at a time. Unfortunately, most of the filter circuits (Arsian *et al*. 2006; Chang *et al*. 2007; Prasad *et al*. 2009; Tangsrirat 2010) and except (Satansup *et al*. 2012; Satansup *et al*. 2014) use at least three passive elements in the design.

Beside this, some of the circuit configurations also employ floating passive elements (Arsian *et al*. 2006; Biolek *et al*. 2009; Chang *et al*. 2007; Prasad *et al*. 2009; Sirirat *et al*. 2010; Singh *et al*. 2013; Tangsrirat 2010; Yuce *et al*. 2004) and, hence, are not favorable for IC implementation. Moreover, few circuits require inverted current input signals and/or component matching constraints (Arsian *et al*. 2006; Chang *et al*. 2007; Chaturvedi and Maheswari 2011; Prasad *et al*. 2009; Sirirat *et al*. 2010; Satansup *et al*. 2014) too for the filtering functions realization. Furthermore, none of the CM filters endorsed in Table 1 can be configured as single input multiple output (SIMO) as well as multiple input single output (MISO) simultaneously from the same structure and are inept to operate at low supply voltages with lower power consumption.

The current-mode filter structure proposed in this paper consists of single CFTA and two grounded capacitors and can realize LP, BP, HP, BS and AP filter functions in three input single output (TISO) and LP, BP, and HP filter functions in single input three output (SITO) configurations. The active element CFTA is slightly modified to get dual electronic tunability feature. The proposed structure is a resistor-less CMOS topology which neither requires inverted current input nor matching constraints for the realization of any filter function(s). With suitable impedance levels the proposed filter blocks are directly cascaded to obtain higher order filters (Kacar and Kuntman 2011; Tomar *et al*. 2014a). The proposed circuits are evaluated through P-SPICE programs on cadence tools using 0.25µm TSMC CMOS parameters (Prommee *et al*. 2009).

Table 1. Comparison among recent single active element based current-mode filters (Prasad *et al*. 2009; Arsian *et al*. 2006).

Attributes / Reference	Active Element	Config.	Passive Elements	Functions Realized	Floating Passive Elements	Explicit Outputs	Electronic Tunability of ω and Q	Designed Pole Frequency	Matching Condition required	Inverted Input required	Power Supply	Power Consumed
(Prasad *et al*. 2009)	DOCDTA		2R+2C	LP,BP,HP	2R	1	YES	NA	YES	NO	± 1 V	NA
(Chaturvedi *et al*. 2011)	DVCC		2R+2C	LP,BP,HP	None	None	NO	9.7 MHz	YES	NO	± 2.5 V	4.9 mW
(Yuce *et al*. 2004)	CCII		2R+2C	LP,BP	1R	None	NO	NA	NO	NO	NA	NA
(Herencsar *et al*. 2009)	CCTA		2C+2R	LP, BP, HP	None	1	YES	1 MHz	NO	NO	± 2 V	NA
(Prasad *et al*. 2013)	VDTA	SIMO	1R+2C	LP,BP,HP	None	1	YES	NA	NO	NO	± 1 V	NA
(Singh *et al*. 2013)	MOCFTA		1R+2C	LP,BP,HP	1R	2	YES	1.35 MHz	NO	NO	± 1.5 V	1.18 mW
(Kacar *et al*. 2012)	FDCCII		2R+2C	LP,BP or HP,BP	None	2	NO	10 MHz	NO	NO	± 1.3 V	NA
(Biolek *et al*. 2009)	CDTA		1R+2C	LP,BP,HP	1R	1	YES	10 MHz	NO	NO	NA	NA
(Tangsrirat 2010)	CFTA		1R+2C	LP,BP,HP	1R	1	NO	159 KHz	NO	NO	± 3 V	
(Satansup *et al*. 2012)	VDTA		2C	LP,BP,HP	None	1	YES	3.03 MHz	NO	NO	± 2 V	NA
(Mangkalakecree *et al*. 2009)	CCCCTA		2C	LP,BP,HP	None	2	YES	572 KHz	NO	NO	± 1.5 V	1.42 mW
(Satansup *et al*. 2014)	VDTA		2C	LP, BP, HP, BS, AP	None	1	YES	3 MHz	NO	YES	± 2 V	1 mW
(Chang *et al*. 2007)	CCCII		1R+2C	LP, BP, HP, BS, AP	1R	1	YES	1 MHz	YES	YES	± 1.65 V	NA
(Kacar *et al*. 2012)	FDCCII	MISO	2R+2C	LP, BP, HP, BS, AP	None	None	NO	1.59 MHz	NO	NO	± 3 V	NA
(Pandey and Paul 2011)	DVCCTA		1R+2C	LP, BP, HP, BS, AP	None	None	NO	1.59 MHz	NO	NO	± 1.25 V	NA
(Sirirat *et al*. 2010)	CFTA		1R+2C	LP, BP, HP, BS, AP	1R	1	NO	1.12 MHz	YES	YES	± 3 V	NA
(Arsian *et al*. 2006)	CCII		2R+2C	LP, BP, HP, BS, AP	1C	1	NO	NA	YES	NO	NA	NA
Proposed	CFTA	Both SIMO/ MISO	2C	LP, BP, HP, BS, AP	None	None	YES	8 MHz	NO	NO	± 0.75 V	0.6 mW

2. Basics of CFTA

Since after its inception as current-mode active element and its terminal impedance suitability for current-mode analog signal processing applications, CFTA have been received considerable attention from the circuit designers (Satansup and Tangsrirat 2011; Tomar *et al.* 2014a; Singh *et al.* 2012; Singh and Maheswari 2013). The conventional CFTA can be further modified for additional advantage of dual electronic tunability. The modified CFTA with symbolic diagram is shown in Fig. 1 and can be described through the following set of equations
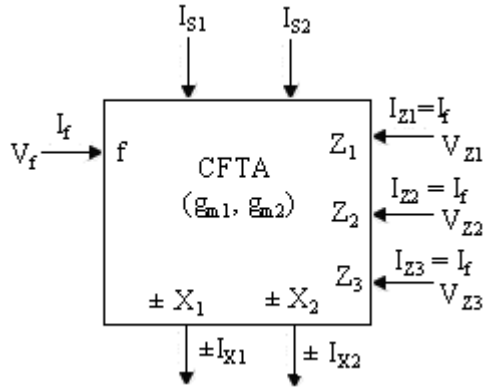
Figure 1. Schematic symbol of modified CFTA.

$$V_f = 0, I_{Z1} = I_{Z2} = I_{Z3} = I_f, I_{\pm X_1} = g_{m1} V_{Z1},$$
$$I_{\pm X_2} = g_{m2} V_{Z2} \tag{1}$$

Realization of the modified CFTA using CMOS is depicted in Fig. 2. The transconductance parameters g_{m1} and g_{m2} of the CMOS implemented CFTA (Tomar *et al.* 2014b; Singh *et al.* 2012) can be related with the biasing currents I_{S1} and I_{S2}, respectively as:
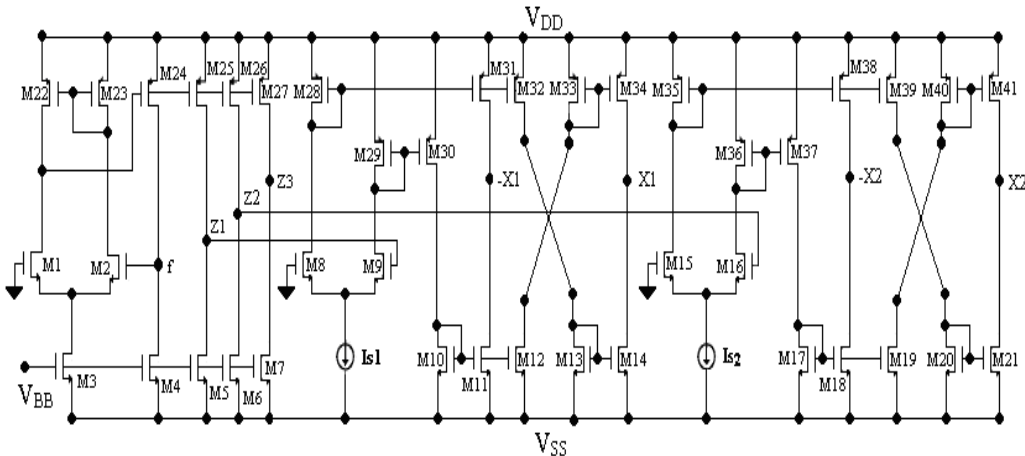
$$g_{m1} = \sqrt{\beta_n I_{S1}} \quad \text{and} \quad g_{m2} = \sqrt{\beta_n I_{S2}} \tag{2}$$

Where, $\beta_n = \mu_n C_{OX} (W/L)$ is the process parameter of NMOS transistors M_8-M_9, M_{15}-M_{16} forming differential pairs at the transconductance stages of employed CFTA.
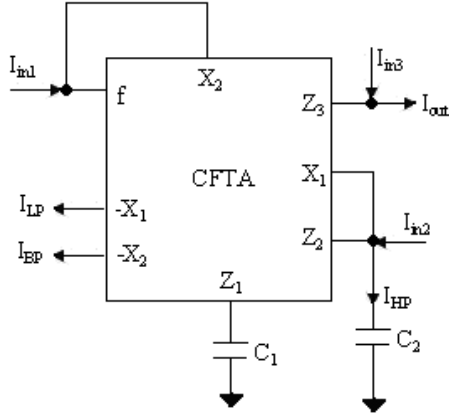
3. Proposed Current-Mode Filter

The proposed current-mode biquad filter circuit as shown in Fig. 3 employs single CFTA as the active element and two grounded capacitors. The use of grounded capacitors is elegant for integrability (Bhusan and Newcomb 1967). The proposed circuit in Fig. 3 can be configured as either three input single output or single input three output configurations through appropriate selection of input and output current signal(s). The workability of the proposed current-mode filter structure can be described in following two different cases:

3.1 Case-I: TISO Filter Configuration

If the proposed circuit in Fig. 3 is analyzed with three current input signals (I_{in1}, I_{in2}, I_{in3}), the expression for current output (I_{out}) can be obtained as:

$$I_{out} = $$
$$\frac{-C_1 C_2 s^2 I_{in1} - g_{m2} C_1 s I_{in2} + (C_1 C_2 s^2 + g_{m2} C_1 s + g_{m1} g_{m2}) I_{in3}}{D(s)}$$
$$\tag{3}$$

Where, $D(s) = s^2 C_1 C_2 + g_{m2} C_1 s + g_{m1} g_{m2}$ $\tag{4}$

Figure 2. CMOS implementation of CFTA.

Figure 3. Proposed current-mode biquad filter.

It can be easily retrieved from equation (3) that different CM filtering responses can be obtained at current output I_{out} by selecting current inputs appropriately, as follows:

- For $I_{in1} = I_{in2} = I_{in3} = I_{in}$ a non-inverted LP response is obtained.

- For $I_{in1} = I_{in}$ and $I_{in2} = I_{in3} = 0$ an inverted HP response is obtained.

- For $I_{in2} = I_{in}$ and $I_{in1} = I_{in3} = 0$ an inverted BP response is obtained.

- For $I_{in2} = I_{in3} = I_{in}$ and $I_{in1} = 0$ a non-inverted BS response is obtained.

- For $I_{in2} = 2I_{in}$, $I_{in3} = I_{in}$ and $I_{in1} = 0$ a non-inverted AP response is obtained.

3.2 Case-II: SITO Configuration

With single current input ($I_{in2}=I_{in}$), the proposed circuit in Fig. 3 is capable to realize LP, BP and HP responses, simultaneously across three different current outputs I_{LP}, I_{BP}, I_{HP}. The transfer functions for received filtering responses are as follows:

$$\frac{I_{BP}}{I_{in}} = \frac{-g_{m2} C_1 s}{D(s)} \tag{5}$$

$$\frac{I_{LP}}{I_{in}} = \frac{g_{m1} g_{m2}}{D(s)} \tag{6}$$

$$\frac{I_{HP}}{I_{in}} = \frac{C_1 C_2 s^2}{D(s)} \tag{7}$$

From cases I and II, it can be remarked that proposed circuit in Fig. 3 can realize all the standard filter functions (LP, BP, HP, BS, and AP) across an explicit high impedance current output terminal by the appropriate selection of three current input signals. In addition, the same structure realizes LP, BP, HP responses, simultaneously by the use of single current input signal. Thus, the proposed circuit can be used as either TISO or SITO current-mode biquad filter without changing the structure and neither requires any component matching constraint nor inverted current input signal for any of the response realization.

The filter's characteristic parameters like pole frequency (ω), quality factor (Q) and band-width (BW) can be derived from the transfer function as:

$$\omega = \sqrt{\frac{g_{m1}g_{m2}}{C_1 C_2}} , Q = \sqrt{\frac{C_2 g_{m1}}{C_1 g_{m2}}} , \ BW = \frac{g_{m2}}{C_2} \tag{8}$$

On substituting trans-conductance parameters g_{m1} and g_{m2} from equation (2) we find that equation (8) turns out to be:

$$\omega = \sqrt{\frac{\beta_n (I_{S1}I_{S2})^{1/2}}{C_1 C_2}}, \ Q = \sqrt{\frac{C_2}{C_1} \left(\frac{I_{S1}}{I_{S2}}\right)^{1/2}} \tag{9}$$

From equation (9), it is evident that the pole frequency can be tuned precisely, without distressing quality factor, if the ratio of biasing currents I_{S1} and I_{S2} is kept to be constant. The band width (BW) of the proposed filter circuit can also be expressed as:

$$BW = \frac{(\beta_n I_{S2})^{1/2}}{C_2} \tag{10}$$

Equation (10) shows that the BW can solely be controlled by only biasing current I_{S2}. From equations (9) and (10), it is clearly seen that filter's parameters ω and Q can be electronically adjusted by regulating the only biasing current I_{S1} without influencing BW.

4. Non-ideal and Parasitic Effects

4.1 Non-Ideal Analysis

This section describes the non-ideal aspects of the proposed biquad structure. For this a non-ideal CFTA can be modeled through modified set of equations, as follows:

$$V_f = 0, I_{Z1} = I_{Z2} = I_{Z3} = \alpha I_f, I_{\pm X1} = \pm \gamma_1 g_{m1} V_{Z1},$$
$$I_{\pm X2} = \pm \gamma_2 g_{m2} V_{Z2} \qquad (11)$$

Where, $\alpha = 1 - \varepsilon$ with ε ($|\varepsilon| \ll 1$) represents the current tracking errors between f to Z_i (i =1, 2, 3) terminals of CFTA and depends on the ratio of PMOS trans-conductance. With the use of identical PMOS transistors at the input stage, α may approach to unity. Similarly, γ_1 and γ_2 are the trans-conductance inaccuracies between Z_1 to $\pm X_1$ and Z_2 to $\pm X_2$ terminals, respectively. On considering these non-ideal factors of the CFTA and re-analyzing the proposed filter of Fig. 3, we get the similar denominator for each current output expressions as:

$$D'(s) = s^2 C_1 C_2 + \alpha \gamma_2 g_{m2} C_1 s + \alpha \gamma_1 \gamma_2 g_{m1} g_{m2} \qquad (12)$$

With involved non-idealities, the modified ω and Q are obtained from equation (12) as:

$$\omega = \sqrt{\frac{\alpha \gamma_1 \gamma_2 g_{m1} g_{m2}}{C_1 C_2}}, Q = \sqrt{\frac{\gamma_1 C_2 g_{m1}}{\alpha \gamma_2 C_1 g_{m2}}} \qquad (13)$$

From equation (13), it can be seen that minor deviations may occur in ω and Q due to involved non-ideal factors. Though these slight deviations can be ignored as non-ideal factors α, γ_1 and γ_2 are imminent to one at operational frequency. However, to what an extent the filter's characteristic may actually be deviated due to possible changes in the design can be visualized by sensitivity coefficients which are as follows:

$$S^{\omega}_{\alpha, \gamma_1, \gamma_2, g_{m1}, g_{m2}} = \frac{1}{2}, \qquad S^{\omega}_{C_1, C_2} = -\frac{1}{2} \qquad (14)$$

$$S^{Q}_{\gamma_1, C_2, g_{m1}} = \frac{1}{2}, \qquad S^{Q}_{\gamma_2, C_1, g_{m2}} = -\frac{1}{2} \qquad (15)$$

Hence, we can state that the entire sensitivity coefficients are within 'half' in magnitude.

4.2 Parasitic Effects

In this section, the performance of the proposed current-mode filter circuit in the presence of diverse parasitic impedance effects of the CFTA are to be considered. A practical CFTA, similar to any other active including various ports parasitic is shown in Fig. 4. These are port Z1, Z_2 and Z_3 parasitic in the form of $R_{Z1}//C_{Z1}$, $R_{Z2}//C_{Z2}$ and $R_{Z3}//C_{Z3}$ respectively,

and ports X_1, X_2 parasitic in the form of $R_{X1}//C_{X1}$, $R_{X2}//C_{X2}$ and port 'f' parasitic in the form of R_f. In Fig. 4 the X_2 port parasitic R_{X2} and C_{X2} appear between the high impedance terminal and ground. To abolish their effect, the CFTA should be well designed to comprise a very low port parasitic R_f. Ideally, the value of R_f is zero and terminal 'f' is virtually grounded. Since these parasitic impedances are connected between true ground and virtual ground, these are, therefore, almost ineffective (Singh et al. 2012). The external capacitances C_1 and C_2 can be chosen to be much greater than the parasitic capacitances at the Z_1, Z_2, X_1 and X_2 terminals of CFTA i.e. C_1, $C_2 \gg C_{Z1}$, C_{Z2}, C_{X1}, C_{X2}.
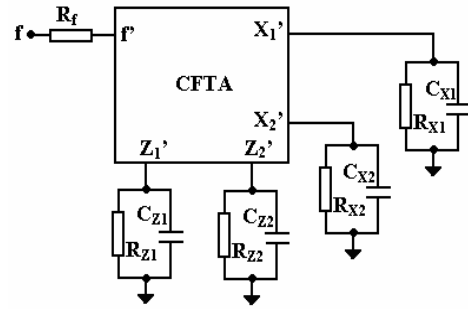

Figure 4. Modified CFTA with port parasitic.

To visualize the effects of various parasitic impedances on the performance of the proposed current-mode filter circuit, the proposed circuit is reanalyzed in the presence of various ports parasitic of CFTA. For this purpose, current-mode BP response is taken in to consideration. By taking the above discussion and parasitic effects into account and on further reanalyzing the proposed current-mode filter, we get following expression for the BP filter at I_{out}.

$$\frac{I_{BP}}{I_{in}} = \frac{\frac{g_{m2}}{C_2}(s + \frac{1}{C_1 R_{Z1}})}{s^2 + s\frac{g_{m2}}{C_2}[L] + \frac{g_{m1} g_{m2}}{C_1 C_2}[M]} \qquad (16)$$

Where, $L = 1 + \frac{1}{g_{m2} R_C} + \frac{C_2}{C_1} \frac{1}{g_{m2} R_{Z1}} \qquad (17)$

and $M = 1 + \frac{1}{g_{m1} R_{Z1}} + \frac{1}{g_{m1} g_{m2} R_C R_{Z1}} \qquad (18)$

With, $R_C = R_{Z2} // R_{X1} \qquad (19)$

And hence, filter parameters in equations (8) to (10) are changed to:

$$\omega^* = \omega\sqrt{M} \qquad (20)$$

$$\left(\frac{\omega}{Q}\right)^* = \frac{g_{m2}}{C_2}[L] = \frac{\omega}{Q}[L] \qquad (21)$$

and

$$Q^* = \sqrt{\frac{g_{m1}C_2}{g_{m2}C_1}}\left[\frac{\sqrt{M}}{L}\right] = Q\left[\frac{\sqrt{M}}{L}\right] \qquad (22)$$

In equations (16)-(22), undesirable factors are yielded due to various ports parasitic of CFTA. By considering $C_1=C_2$, and $g_{m1}=g_{m2}$ in the design, these undesirable factors can be eliminated or minimized and hence, the proposed filter may approach towards ideal response, if we choose the design criterion given in equation (23).

$$\frac{1}{g_{m2}R_{Z1}} << 1 , \quad \frac{1}{g_{m2}R_C} << 1 , \quad |\omega C_1| >> \frac{1}{R_{Z1}} \qquad (23)$$

5. Simulation Results

In this section, the current-mode universal filter in Fig. 3 has been simulated using PSPICE programs to authenticate the theoretical expectations. The employed CFTA as given in Fig. 1 was implemented using standard CMOS technology (TSMC 0.25μm process parameters) as shown in Fig. 2. The dimensions of MOS transistors were obtained as summarized in Table 2. The circuit was biased with power supply $V_{DD} = -V_{SS}= 0.75V$ and $V_{BB} = -0.32V$ and designed to realize biquadratic characteristics with f = 8 MHz at Q = 1. In design, the active and passive components value were determined as $I_{S1} = I_{S2}$ = 40 μA and $C_1 = C_2$ = 10 pF, which resulted in total average power consumption of about 0.6 mW. Fig. 5 shows the simulated and ideal current-gain responses of BP, LP and HP filter functions for the SITO configuration of the proposed current-mode filter circuit. Similarly, for the TISO configuration the simulated and ideal current-gain and phase responses for LP, HP, BP, BS and AP filtering functions are shown in Fig. 6. In Fig. 5 and Fig. 6 the difference between ideal and simulated results is clearly visible along with distortion (spike) beyond 670 MHz or so. Which is obviously due to involved non ideal factors and parasitic effects dominating at higher frequencies. However, at working frequencies these non ideal factors were found to be unified as discussed in section

4. For different filtering responses the pole frequency was measured as 7.94 MHz, which deviated only by 0.75% from the numerical designed value of 8 MHz.

To show the electronic tuning aspects of the proposed current-mode filter, the circuit is further simulated to obtain various BP responses at different sets of I_{S1} and I_{S2} in such a way so that $I_{S1}=I_{S2}$ = 5μA, 10μA, 30μA, 150μA, which result in to the pole frequency variation as 1.25 MHz, 3.22 MHz, 7.25 MHz and 12.03 MHz, respectively, at constant Q = 1. The simulated result showing the pole frequency tuning aspect independent of Q is shown in Fig. 7. Similarly, the simulation results showing the tuning feature of Q independent of f_0 for the SITO configuration is also shown in Fig. 8, which were obtained by maintaining the product of I_{S1} and I_{S2} to be a constant as (I_{S1} =10μA, I_{S2} =160 μA), (I_{S1} =16μA, I_{S2} =100 μA), (I_{S1} =32μA, I_{S2} =50 μA), and (I_{S1} =64μA, I_{S2} =25 μA) respectively.

In Fig. 9, the time-domain analysis of BP filter (SITO configuration) is displayed, which was obtained by applying a sinusoidal input of 80μA peak to peak at 7.94 MHz. Similarly, the time-domain analysis of LP filtering function (TISO configuration) is also shown in Fig. 10, which was obtained by applying a sinusoidal input of 80μA peak to peak at 200 MHz. Next, the THDs of the LP response of the proposed circuit were also measured by applying a current input sinusoidal signal of peak amplitude of 40 μA and having varying frequency from 50 KHz to 900 KHz.

It can be observed that the THD figures are adequate and not more than 3%, causes the output signal is not distorted significantly for an applied input signal of varying frequency as shown in Fig. 11. To observe the effect of passive component mismatching on the filter's performance, Monte-Carlo analysis has been performed for 100 samples.

For this, the BP filter of SITO configuration was simulated by selecting the values of capacitors C_1 and C_2 with 5% Gaussian deviation. The statistical results in the form of plots are depicted in Fig. 12, the simulated mean, median and standard deviations were respectively, 8.24 MHz, 8.22 MHz and 164.66 KHz, which reveal that with respect to the simulated pole frequency of 7.94, the proposed filter is less sensitive to the change in capacitors value and thus offers reasonable passive sensitivity.
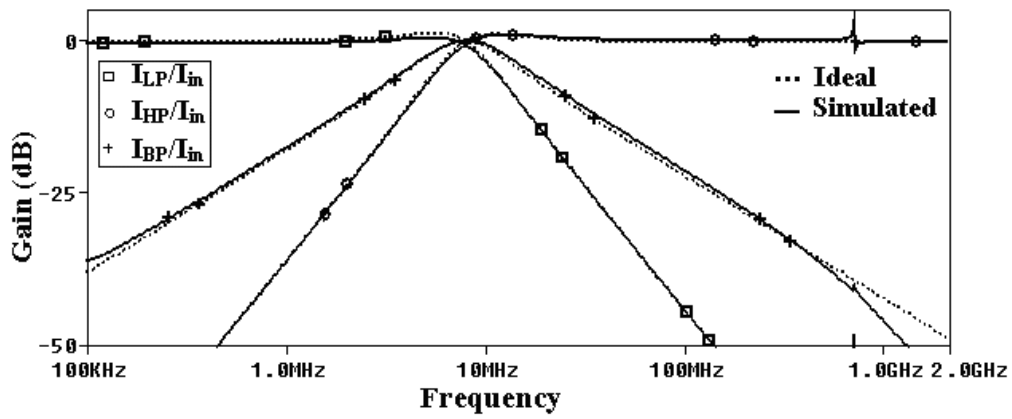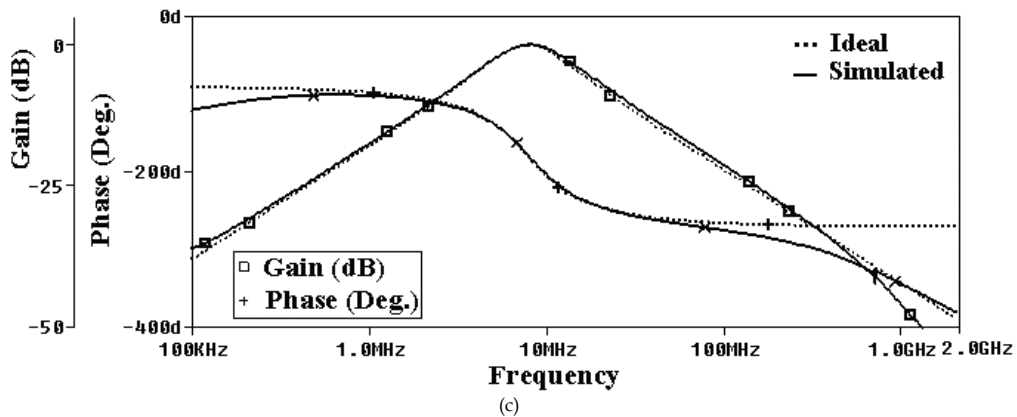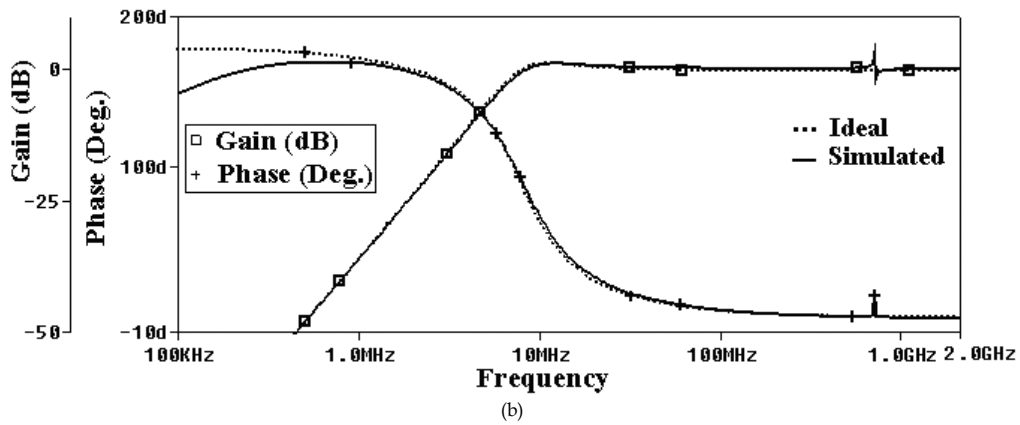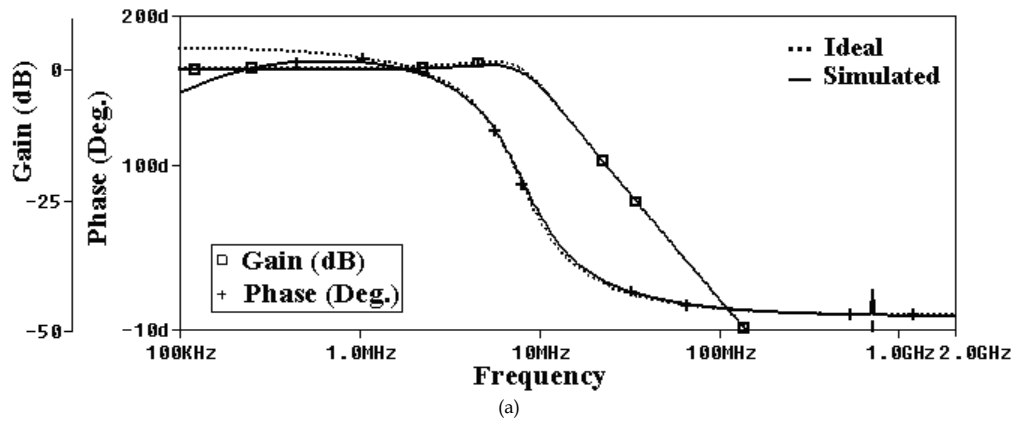
Figure 5. Current-gain responses of LP, HP and BP filters for SITO configuration of the proposed filter in Figure 3.
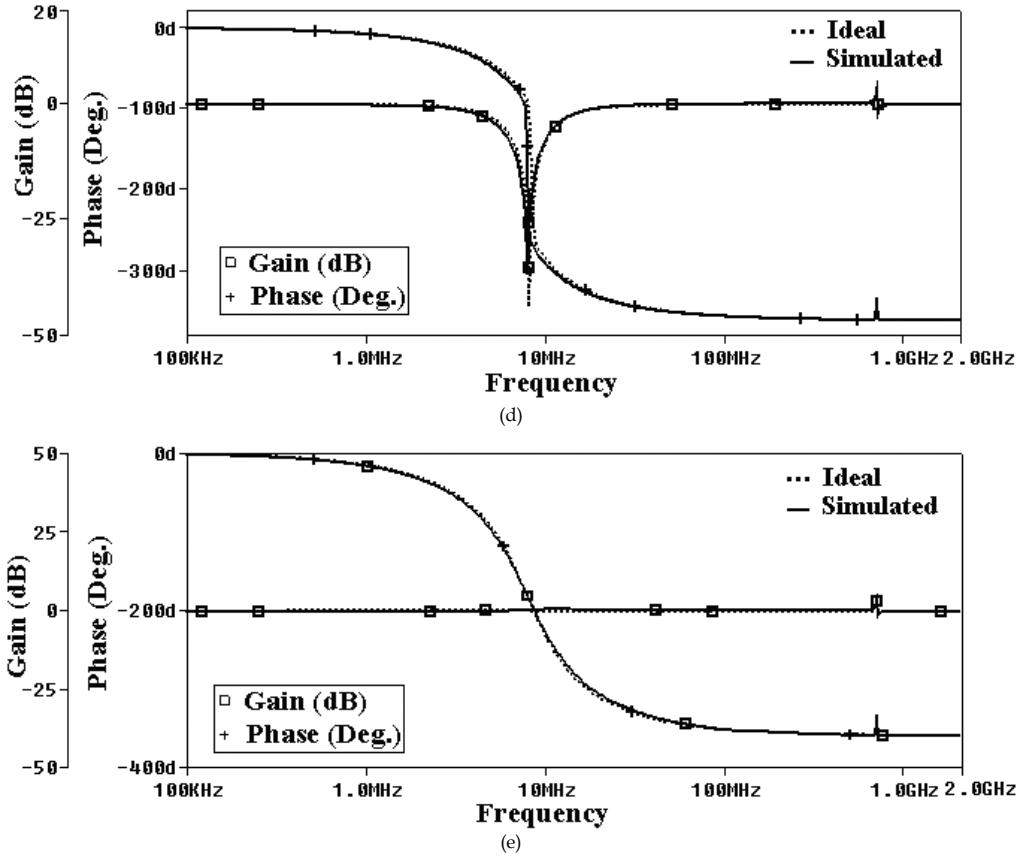

(a)

(b)

(c)

(d)

(e)

Figure 6. Simulated and Ideal Current-gain and phase responses for TISO configuration of the proposed filter in Figure 3 (a) LP (b) HP (c) BP (d) BS and (e) AP filter functions.
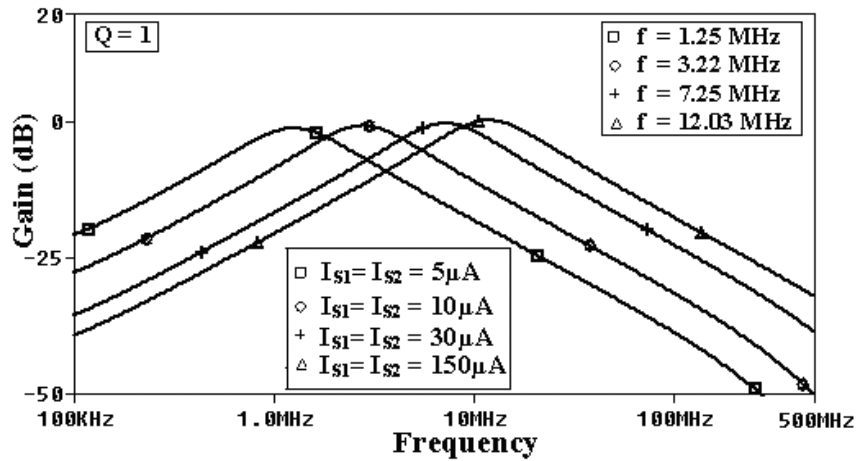

Figure 7. Simulated BP responses of SITO configuration of the proposed filter in Figure 3, showing variation in pole frequency, and keeping Q = 1.
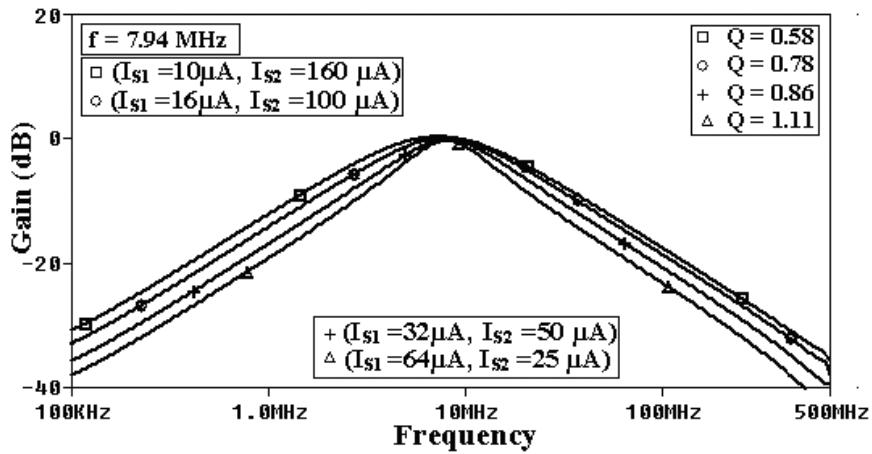
Figure 8. Simulated BP responses of SITO configuration showing variation in Q, by keeping pole frequency at 7.94 MHz.
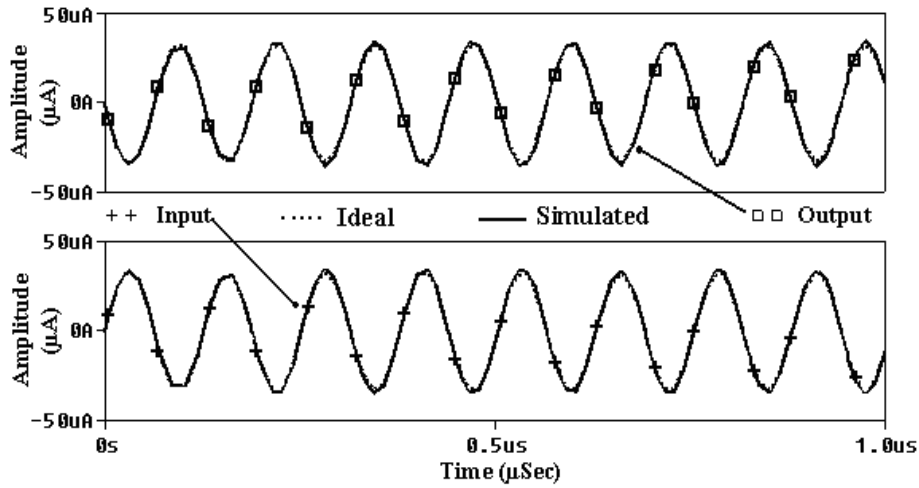

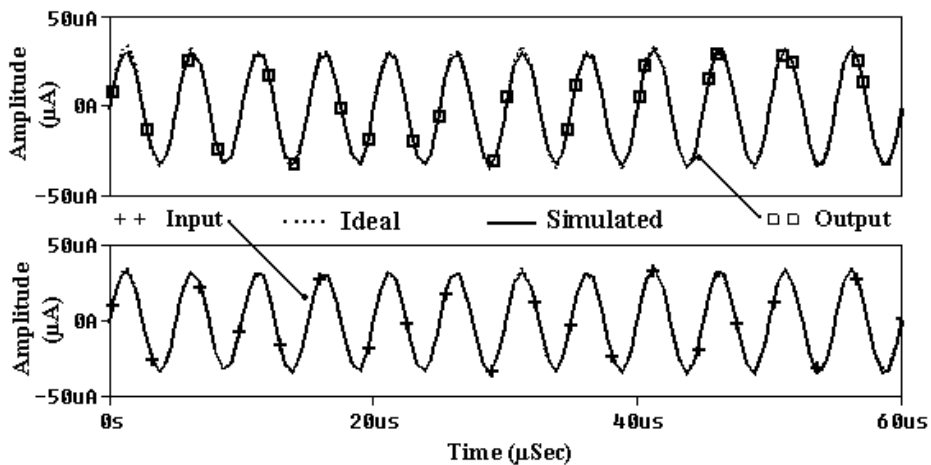**Figure 9.** Time domain response for BP filter of SITO configuration of the proposed filter in Figure 3.

Figure 10. Time domain response for LP filter of TISO configuration of the proposed filter in Figure 3.
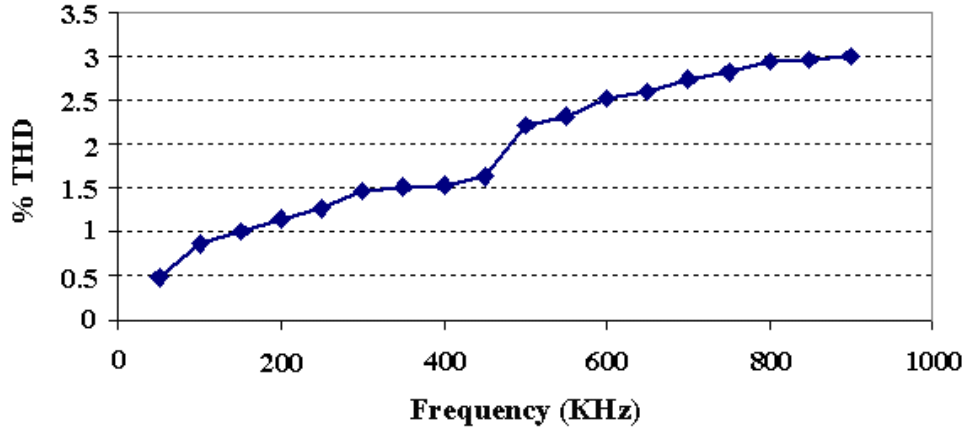
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Figure 11. THD performance of LP response of the SITO configuration for a sinusoidal current input of amplitude 40µA, for a frequency range 50 to 900 KHz.
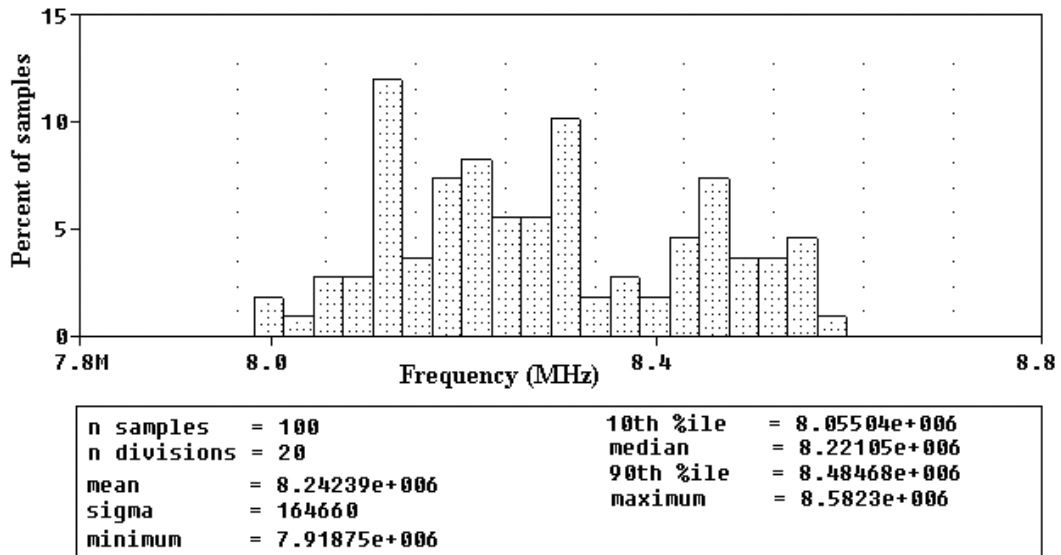

Figure 12. Statistical results of Monte-Carlo analysis for the BP response of SITO configuration of the proposed circuit with 5% deviation in capacitor values.

6. Application: Higher Order Filters

To further support the proposed CM filter topology in Fig. 3, it is used as a general purpose structure for realizing higher order current-mode filter (Kacar and Kuntman 2011; Tomar *et al*. 2014a). For example, nth order (where n = even) BP filter and HP filter are shown in Fig. 13 and Fig. 14, which are implemented through direct cascading (Maheshwari *et al*. 2004) of m = n/2, similar proposed blocks without using any impedance matching or coupling. Direct cascading facilitates with no lower cut-off frequency limitations and circuit can operate from low to High frequency range.

Simulations are next performed for the 4th, 6th and 8th order fourth-order filter to illustrate the frequency domain performance of the circuit and corresponding results are shown in Fig. 15 and Fig. 16. In this design, the component values for each of the filter block (1, 2…..m) were chosen as I_{Sm}= 40 µA and $C_{1m} = C_{2m} = 10$ pF. The voltage supply rails were used as $V_{DD} = -V_{SS} = 0.75$ V with a bias $V_{BB} = -0.32$ V. From different simulation results, it is evident that the higher order circuits realized through direct cascading are attractive choice and operated efficiently at low supply rails with least power consumption.
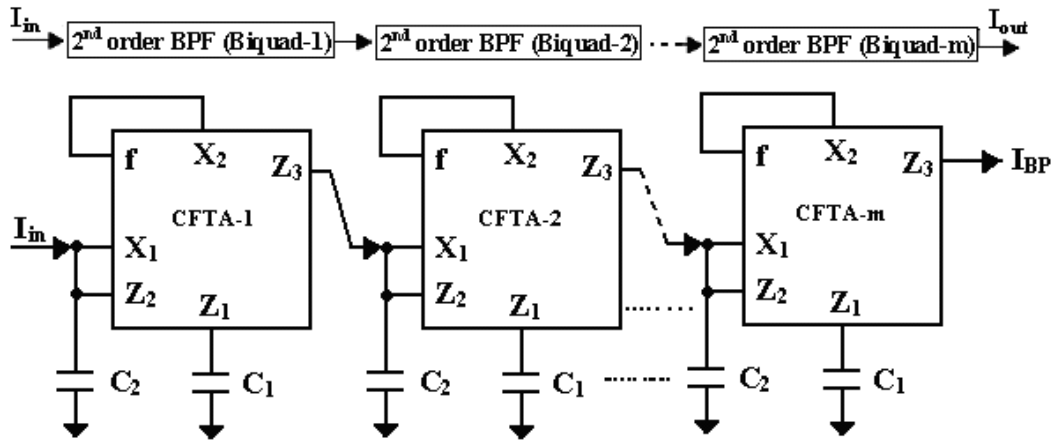
Figure 13. N^{th} - order band pass (BP) filter, derived through direct cascading of the proposed circuit blocks in Figure 3.
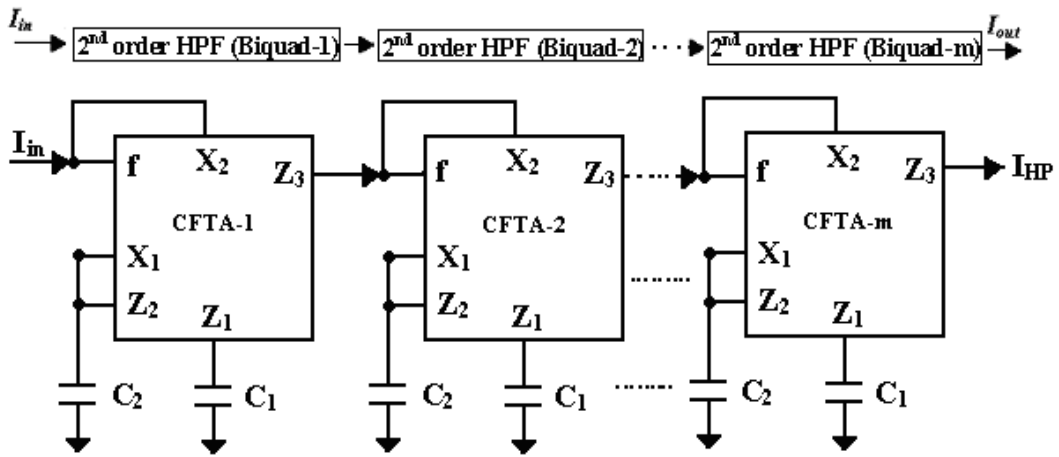

Figure 14. N^{th} - order high pass (HP) filter, derived through direct cascading of the proposed circuit blocks in Figure 3.

Table 2. Dimensions of MOS transistors in Figure 2.

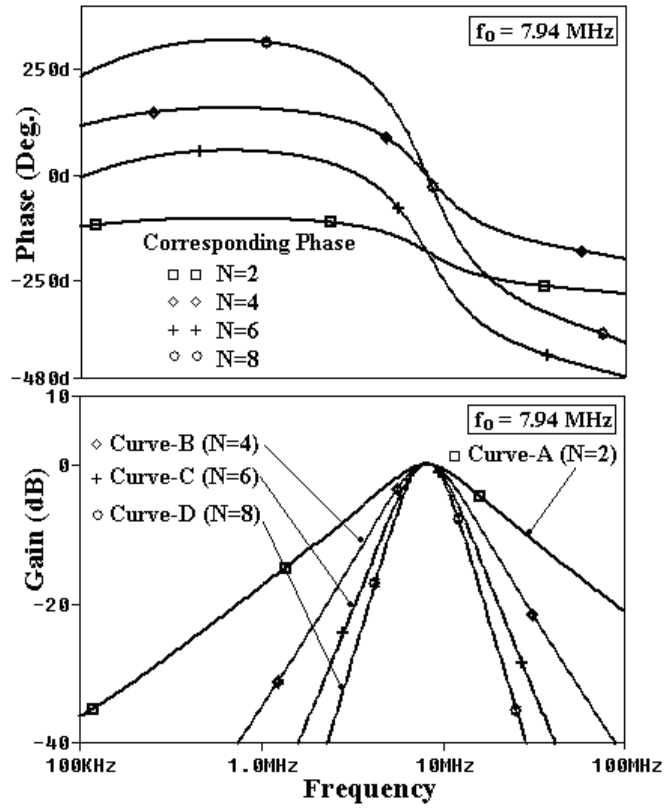| MOS Transistors	Dimensions W (μm)/L (μm)
M_1, M_2	1/0.25
M_3-M_7, M_{10}-M_{14}, M_{17}-M_{21}	3/0.25
M_8, M_9, M_{15}, M_{16}	15/0.25
M_{22}-M_{27}, M_{29}-M_{33}, M_{35}-M_{41}	5/0.25
M_{24}	10/0.25
M_{28}, M_{34}	4.2/0.25

Figure 15. Current-gain and corresponding phase response for the Nth-order (N = 2, 4, 6, 8) BP response of Figure 13.

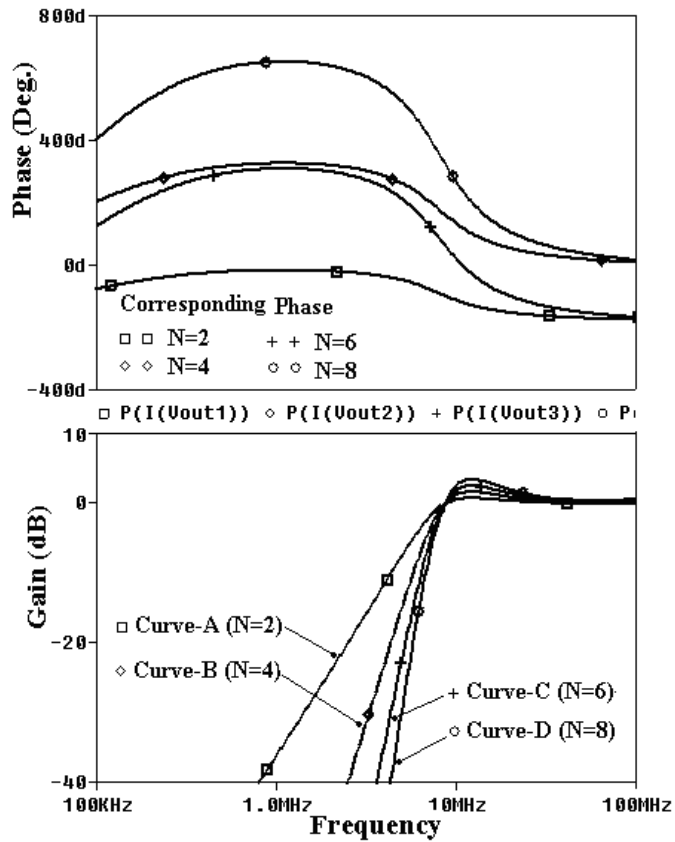**Figure 16**. Current-gain and corresponding phase response for the Nth-order (N = 2, 4, 6, 8), HP response of Figure 14.

7. Conclusion

In this paper, a single CFTA and two grounded capacitors based current-mode biquad filter is proposed and verified using simulation results. In addition, non ideal aspects and parasitic impedance effects of the CFTA were also considered in details. The simulation results agree quite well with the theoretical expectations.

The proposed circuit has a number of noticeable advantages as discussed below. The proposed circuit shows versatility by realizing LP, BP, HP, BS and AP functions in TISO as well as LP, BP, HP responses as the SITO configuration. The use of a single active element makes the circuit topology simpler. Moreover, the use of two grounded capacitors provides a canonical structure, leads to the option for integrated circuit implementation and for parasitic reduction. Moreover, the circuit neither requires inverted current input(s) nor any component matching conditions for response realization, and thus eliminates the need of complex external circuitry. Orthogonal control of Q and ω via biasing currents suited for practical adjustments in music and speech synthesis. The proposed circuit is a low voltage and low power structure, hence it is suited for battery operated systems. Expansion possibility of the proposed circuit for higher order realization through direct cascading validates its usefulness for analog signal processing. The CMOS compatibility of the circuit is fully suited for low cost integrated circuit implementation. Moreover, the circuit offer low active and passive sensitivity performance.

A careful inspection of Table 1 and introductory section of this study describes the superiority of the proposed filter block over other single active element based designs and finally concludes that none of the earlier reported current-mode filters offers all the above described features simultaneously.

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