

Electronically tunable current-mode multifunction filter using current-controlled current follower transconductance amplifier

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Abstract: A new electronically tunable current-mode multifunction universal filter with three inputs and one output based on current-controlled current follower transconductance amplifier is presented. The proposed filter can implement low-pass, band-pass, high-pass, band-stop and all-pass transfer functions with a single topology. For implementation of these transfer functions, no passive component-matching conditions, no inverted input signal requirements and high-output impedance are required. Also the proposed filter offers electronic control of the natural angular frequency, low active and passive sensitivities and use of grounded capacitors which is ideal for integrated circuit implementation. The proposed universal biquadratic filter has been used for implementing sixth-order filters. PSPICE simulation results confirm the presented theory.

Keywords: Universal filter; current follower transconductance amplifier (CFTA); current-controlled CFTA (CCFTA); current-mode circuit; high-order filter

Elektronsko nastavljiv multifunkcijski filter v tokovnem načinu za uporabo v tokovno krmiljenem transkonduktančnem ojačevalniku

Izveček: Predstavljen je nov elektronsko nastavljiv multifunkcijski filter v tokovnem načinu s tremi vhodi in enim izhodom na osnovi tokovno krmiljenega transkonduktančnega ojačevalnika. Predlagan filter lahko vsebuje, nizko pasovno, pasovno, visoko pasovno, pasovno blokirno in vse-propustno prenosno funkcijo v enojni topologiji. Za implementacijo propustnih funkcij ne potrebujemo pogoje usklajenosti pasivnih komponent, invertiranih vhodnih signalov ali visokih izhodnih impedanc. Filter prav tako omogoča elektronski nadzor naravne kotne frekvence, nizko aktivno in pasivno občutljivost in ozemljene kondenzatorje, ki so idealni za implementacijo v integrirana vezja. Predlagan filter je bil uporabljen kot filter šestega reda in simuliran v okolju PSPICE.

Ključne besede: univerzalni filter; tokovni transkonduktančni ojačevalnik; tokovno krmiljenje; filter visokega reda

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1 Introduction

The universal biquadratic filters are classified as second-order filters that typically implement five filtering functions with a single topology such as low-pass (LP), band-pass (BP), high-pass (HP), band-stop (BS) and

all-pass (AP) transfer functions. The biquadratic filters can be used in electronic and communication systems such as phase locked loop (PLL), touch-tone telephone tone decoder, cross-over network for a three-way high-fidelity loudspeaker [1]. It is also well-known that

biquadratic filter can be used for implementing high-order filters [2]. As a result, many universal biquadratic filters are reported; see, for example [3]-[26]. Considering the input and output terminals, these filters can be classified in three categories, that are a single-input multiple-output (SIMO) filter, a multiple-output single-input (MISO) filter and a multiple-input multiple-output (MIMO) filter. When a single signal is applied at the input of a SIMO filter, filtering functions such as LP, BP, HP, BS and AP functions can be obtained at each output terminal. Thus, a SIMO filter can generate the response to several filtering functions without changing the input terminal and without requiring additional circuitry. Unfortunately, a SIMO filter normally requires several active and passive devices, if five standard filtering functions are implemented. Compared with SIMO filter, MISO and MIMO filters require fewer active and passive components, because the filtering function is selected by appropriately applying the input signals and/or selecting the output signals. However, if five filtering functions are required, additional summing and subtraction amplifiers are needed. This requirement is difficult especially for voltage-mode (VM) filters where addition and subtraction voltage amplifiers are required using several passive components. Fortunately, this problem is not present in current-mode (CM) filters, because summing and subtracting currents can be implemented in a straightforward manner. Moreover, multiple copies of an input signal can be easily implemented with multiple-output current mirrors.

Current-mode (CM) signal processing circuits received considerable attention because this technique offers several advantages compared to voltage-mode (VM) signal processing circuits, such as greater signal bandwidth, wider dynamic range and especially simpler circuitry [27], [28]. Considering the universal filters in [3-26], the circuits in [3-12] are VM filters while the circuits in [13-26] are CM filters. This paper is focused on the CM filters which are supposed to use only a single active device and grounded capacitors. Several active devices have been used to realize CM universal filters; see, for example [13-33]. The CM filters in [13-17] use second-generation current conveyor (CCII) as the active element. However, these structures suffer from a lack of electronic tuning capability. The CM filters with an electronic tuning capability can be realized using operational transconductance amplifiers (OTAs) [18-20] and second-generation current-controlled current conveyors (CCCII) [21-26], but these circuits use a large number of active elements.

Recently, a new current-mode active device with two current inputs and two kinds of current output referred to as a current differencing transconductance amplifier (CDTA), has been proposed [29]. This device is a

synthesis of the well-known advantages of the current differencing buffered amplifier (CDBA) [30] and the transconductance amplifier (TA) to facilitate the implementation of current-mode analog signal processing circuits. Some current-mode universal filters using CD-TAs as active elements have been reported in technical literature, see, for example [31-34]. However, these reported filters require more than one CDTA. Moreover, some configurations do not exploit the full capability of the CDTA when typically one of two input terminals of the CDTA is floated and not used [31-33]. Unfortunately, this can cause noise injection in a monolithic circuit [35].

More recently, a new active element with one current input and two kinds of current outputs, the so-called "current follower transconductance amplifier (CFTA)", has been introduced [36]. It is obtained by modifying the original CDTA. It is similar to the CDTA except for current input. The current input of CFTA is operated as a current follower. CFTA-based universal filters were already proposed [37-48]. However, the reported circuits in [37-46] require an excessive number of active components while reported circuits in [47], [48] provide only three filtering functions and some output current terminals do not exhibit high output impedance, thus additional current followers are needed for avoiding the loading problem. Active filters employing only a few active components have a lower power consumption and smaller chip area when implemented as an IC. Also the use of grounded capacitors is suitable for IC implementation [49].

Several current-mode universal filters using a single active element have been proposed in the technical literature; see, for example [50-55]. However, the reported filters suffer from one or more of the following disadvantages: (i) lack of electronic tuning capability [50-52], (ii) use of either floating capacitors or floating resistors [50-53], (iii) cannot provide five standard filtering functions [50], [53], [54], (iv) do not exploit the full capability of the active device when the y-terminal of the current conveyor is not used [55].

In this paper, a new electronically tunable current-mode universal filter employing only a modified CFTA and two grounded capacitors, is presented. The proposed circuit can implement LP, BP, HP, BS, and AP filtering functions simultaneously, by appropriately applying the input signals. For realizing these filtering functions, no passive component-matching conditions and no inverted input signal are required. Also the natural angular frequency (ω_0) can be electronically controlled. The proposed universal filter has been used to realize high-order filters as application examples. PSPICE simulation results confirm the characteristics

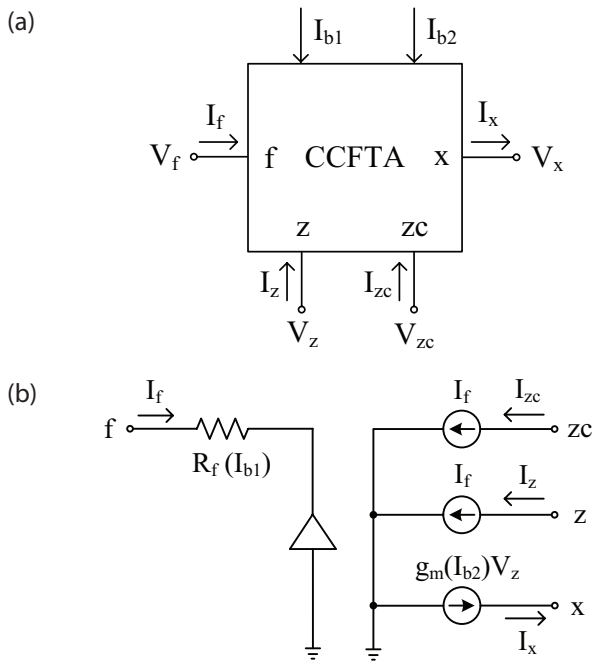


Figure 1: CCFTA: (a) circuit symbol, (b) equivalent circuit.

of the proposed circuit. The comparison between the proposed filter and some previous work is summarized in Table 1. From Table 1 it can be seen that when compared with CCII-based filters in [13-17], the proposed filter provides an electronic tuning capability whereas when compared with filter structures that enjoy an electronic tuning capability in [19-33], the proposed filter uses fewer active elements and when compared with CFTA-based filters in [37-46], the proposed filter uses fewer active and passive elements. Also when

compared with the filters using a single active element in [47-55], the proposed filter provides five standard filtering functions, electronic tuning capability, the use of grounded capacitors and high-output impedance.

2 Circuit realization

The circuit symbol and the equivalent circuit of the CCFTA are shown in Fig. 1 (a) and (b). The ideal characteristic of CCFTA can be described as

$$\begin{pmatrix} I_z \\ I_{zc} \\ I_x \\ V_f \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 & 1 \\ 0 & 0 & 0 & 1 \\ g_m & 0 & 0 & 0 \\ 0 & 0 & 0 & R_f \end{pmatrix} \begin{pmatrix} V_z \\ V_{zc} \\ V_x \\ I_f \end{pmatrix} \quad (1)$$

where R_f and g_m are the internal resistance at the f-terminal and the transconductance gain of the CCFTA, respectively. The properties of this device are similar to those of the CFTA [36], [42] except for the f-terminal of CCFTA has finite input resistance R_f . From Fig. 1(b), the parasitic resistance R_f can be controlled by adjusting the bias current I_{b1} . This property makes it different from conventional CFTA. From Fig. 1(a), the transconductance g_m can also be controlled by adjusting the bias current I_{b2} . The current I_z can be copied to current I_{zc} at the zc-terminal. This terminal may be called the z-copy terminal [36] and it can be realized both as plus- and minus-type zc terminal. Similarly, the plus- and minus-type x-terminals can also be obtained.

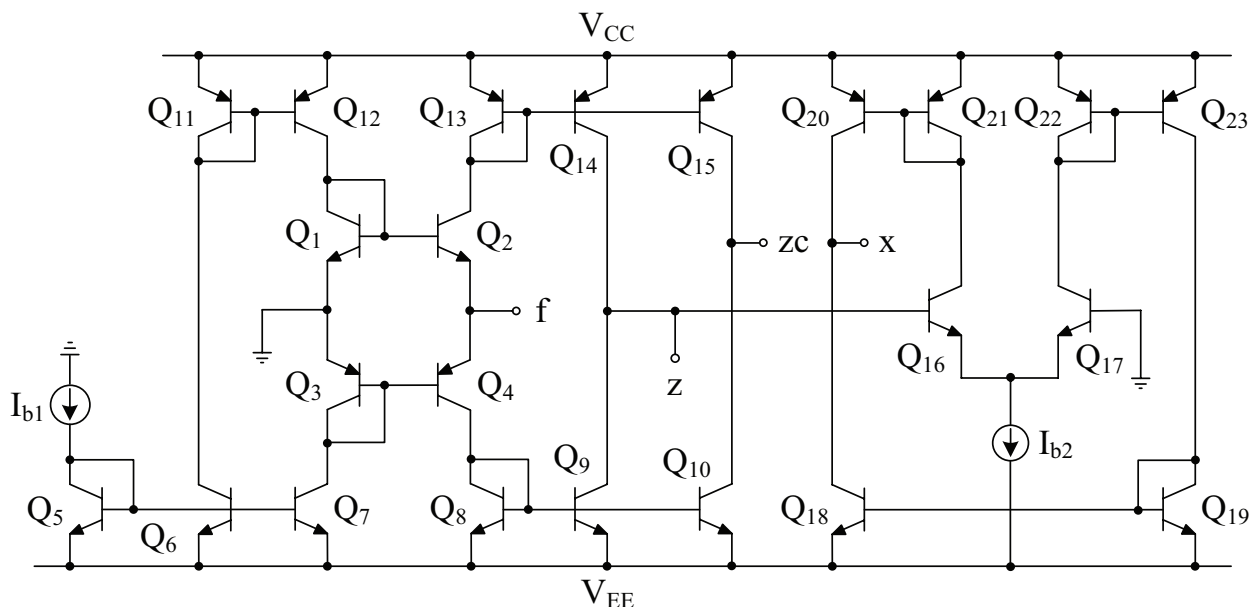


Figure 2: Bipolar implementation of the CCFTA.

Table 1: Comparison of the proposed filter with existing CM filters (Continued).

Circuits	Technology or active device used	Number of active elements	Number of resistors (R) & capacitors (C)	Offers five standard filters simultaneously	Controlling ω_o without affecting Q	All-grounded passive components	High-output impedance	Operational frequency (simulated)
Proposed filters	HFA3096 (BJT)	1-CCFTA	2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [13]	AD844N (BJT)	6-CCII	6-R & 2-C	Yes	Yes	Yes	Yes	>1 MHz
Ref. [14] (Fig.8)	LM13600, LF356 (BJT)	10-CCII	8-R & 2-C	Yes	Yes	Yes	Yes	>10 MHz
Ref. [15] (Fig.2)	AD844N (BJT)	4-CCII	2-R & 2-C	Yes	Yes	No	Yes	>1 MHz
Ref. [16] (Fig.3)	Ad844N (BJT)	1-CCII	2-R & 2-C	No	Yes	No	No	>1 MHz
Ref. [17] (Fig.3a)	0.35 μm CMOS	1-FDCCII	2-R & 2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [18]	0.5 μm CMOS	4-OTA	2-C	Yes	Yes	No	Yes	>10 MHz
Ref. [19]	0.5 μm CMOS	2-OTA	2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [20]	LM13600 (BJT)	2-OTA	2-C	Yes	Yes	Yes	Yes	>10 MHz
Ref. [21]	ALA400 BJT	5-CCCII	2-C	Yes	Yes	Yes	Yes	>10 MHz
Ref. [22]	ALA400 BJT	4-CCCII	2-C	Yes	Yes	Yes	Yes	>2 MHz
Ref. [24]	ALA400 BJT	4-CCCII	2-C	Yes	Yes	Yes	Yes	>10 MHz
Ref. [26]	HFA3096 (BJT)	2-CCCII	2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [31]	0.5 μm CMOS	2-CDTA	2-C	No	Yes	Yes	No	>100 MHz
Ref. [32]	0.5 μm CMOS	2-CDTA	2-C	Yes	Yes	Yes	Yes	>10 MHz
Ref. [33]	0.35 μm CMOS	3-CDTA	1-R & 2-C	Yes	No	Yes	No	>10 MHz
Ref. [37]	ALA400 BJT	3-CFTA, 3-PCA	1-R & 2-C	No	Yes	Yes	Yes	>10 MHz
Ref. [38]	AD844N, MAX435	3-CFTA, 3-CMI	3-CFTA, 3-CMI	1-R & 2-C	Yes	Yes	Yes	>10 MHz
Ref. [39]	AD844N, MAX435	3-CFTA	1-R & 2-C	Yes	Yes	Yes	Yes	>10 MHz
Ref. [40]	AD844N, MAX435	3-CFTA	1-R & 2-C	Yes	No	Yes	No	>10 MHz
Ref. [41]	0.5 μm CMOS	3-CCFTA, 1-OTA	2-C	No	Yes	Yes	Yes	>100 MHz
Ref. [42]	0.35 μm CMOS	3-ZC-CFTA	1-R & 2-C	Yes	Yes	No	Yes	>100 MHz
Ref. [43]	AD844N, MAX435	3-CFTA	2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [44]	AD844N, MAX435	3-CFTA	1-R & 2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [45]	ALA400 BJT	3-CFTA	2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [46]	0.35 μm CMOS	3-ZC-CFTA	2-C	Yes	Yes	Yes	Yes	>100 MHz
Ref. [47]	ALA400 BJT	1-CFTA	1-R & 2-C	Yes	Yes	Yes	No	>10 MHz
Ref. [48]	0.5 μm CMOS	1-CCFTA	2-C	No	Yes	Yes	No	>100 MHz
Ref. [50]	CMOS	1-CCII	3-R & 2-C	No	No	No	No	>10 MHz
Ref. [51]	AD844N (BJT)	1-CCII	2-R & 2-C	No	Yes	No	Yes	>10 MHz
Ref. [52]	AD844N (BJT)	1-CFOA	2-R & 2-C	No	Yes	No	Yes	>10 MHz
Ref. [53]	0.35 μm CMOS	1-CDTA	1-R & 2-C	No	No	No	No	>10 MHz
Ref. [54]	0.5 μm CMOS	1-CDTA	2-R & 2-C	No	No	No	No	>100 MHz
Ref. [55]	ALA400 BJT	1-CCCCTA	2-C	Yes	Yes	Yes	Yes	>3 MHz

The bipolar implementation CCFTA that was used in this work is shown in Fig. 2. It should be noted that if CMOS implementation of CCFTA is required, the bipolar junction transistors in Fig. 2 can be replaced by MOS transistors counterparts. Assuming that transistors, Q_1 to Q_4 , in Fig. 2 are identical, the resistance at f-terminal (R_f) can be expressed [56] as

$$R_f = \frac{V_T}{2I_{b1}} \tag{2}$$

where V_T is the thermal voltage.

Assuming transistors Q_{16} and Q_{17} are identical, the transconductance gain (g_m) can be expressed as

$$g_m = \frac{I_{b2}}{2V_T} \tag{3}$$

The multiple-output plus/minus CCFTA can be obtained by adding additional current mirrors and cross-coupled current mirrors to obtain plus- and minus-type outputs $\pm zc$ and $\pm x$ [24]. It should be noted from Fig. 2 that there are two parasitic parameters available for implementing universal filter, meaning that passive de-

vices such as resistors are not required. Therefore the CCFTA-based universal filter can be tuned electronically. If Fig. 2 is implemented using CMOS technology, the values of R_f and g_m in (2) and (3), are proportional to the square root of the bias current. This, however, changes the electronic tunability of the CCFTA-based universal filter in the sense that the tuning range is no longer linear.

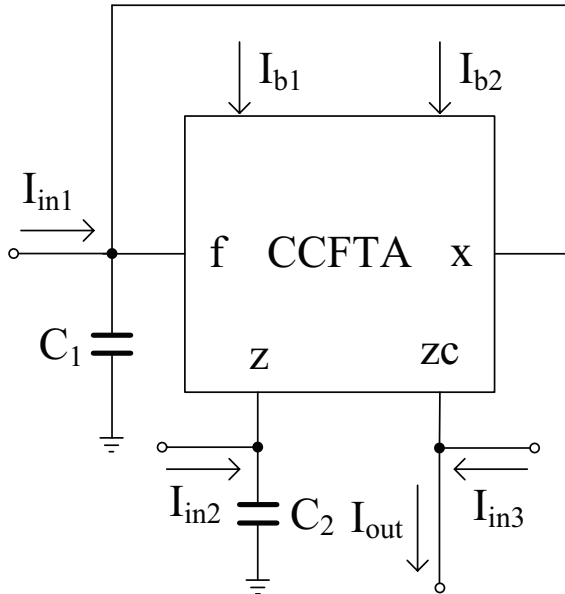


Figure 3: Proposed universal multifunction filter using CCFTA.

The proposed current-mode universal multifunction filter using minimum number of active and passive components is shown in Fig. 3. This filter is developed from a previously reported filter in [57]. The circuit consists of only one CCFTA and two grounded capacitors which is the main advantage of proposed circuit. It should be noted that the proposed circuit uses grounded capacitors which is ideal for IC implementation [49]. Assuming I_{in1} , I_{in2} and I_{in3} are input currents, using nodal analysis and CCFTA characteristic given in (1), current output I_{out} of the proposed filter can be expressed as

$$I_{out} = \frac{D(s) I_{in3} - g_m I_{in2} - sC_2 I_{in1}}{D(s)} \quad (4)$$

where $D(s) = s^2 R_f C_1 C_2 + sC_2 + g_m$.

From (4), the LP, BP, HP, BS and AP filters can be obtained as follows:

- The LP response can be obtained if $I_{in2} = I_{in}$ and $I_{in1} = I_{in3} = 0$.
- The BP response can be obtained if $I_{in1} = I_{in}$ and $I_{in2} = I_{in3} = 0$.
- The HP response can be obtained if $I_{in1} = I_{in2} = I_{in3} = I_{in}$.

- The BS response can be obtained if $I_{in1} = I_{in3} = I_{in}$ and $I_{in2} = 0$.
- The AP response can be obtained if $2I_{in1} = I_{in3} = I_{in}$ and $I_{in2} = 0$.

Therefore, the proposed filter in Fig. 3 can implement five standard filtering functions with a single topology. It should be noted that the realization requires no passive-matching condition and no inverted input signal. For obtaining HP, BS and AP responses, multiple- and/or double-input signals are required, but which can be easily obtained with a multiple-output current follower circuit. However, compared with LP and BP responses, HP, BS and AP responses may suffer from the input current mismatch because two identical input signals are required. This mismatch can disturb the operation of some responses, especially for obtaining the AP response when the condition of $2I_{in1} = I_{in3} = I_{in}$ is needed. This problem can be minimized by carefully designing the current follower. Also it should be noted that the current gains of the LP, HP and BP responses are equal to unity. If a filtering function with a current gain is required, additional active elements such as current amplifiers [58] are needed. The use of current amplifier at the input avoids the problem of input impedance dependency on the frequency.

The peak frequency ω_o and quality of BP filter $Q = \omega_o / BW$ is usually related, where BW is the bandwidth. It should be noted that the relation of Q and BW is inverse, thus the higher Q , the narrower BW of BP filter. Meanwhile, the peak frequency ω_o for the LP and HP filters will also increase with increasing the value of Q . The parameters ω_o and Q are calculated, respectively, as

$$\omega_o = \sqrt{\frac{g_m}{C_1 C_2 R_f}} \quad (5)$$

$$Q = \sqrt{R_f g_m \left(\frac{C_1}{C_2} \right)} \quad (6)$$

Using (2) and (3), the parameters ω_o and Q in (5) and (6) can be rewritten as

$$\omega_o = \frac{1}{V_T} \sqrt{\frac{I_{b1} I_{b2}}{C_1 C_2}} \quad (7)$$

$$Q = \frac{1}{2} \sqrt{\frac{I_{b2}}{I_{b1}} \left(\frac{C_1}{C_2} \right)} \quad (8)$$

Letting $I_{b1} = I_{b2} = I_b$, (7) and (8) simplify to

$$\omega_o = \frac{I_b}{V_T} \sqrt{\frac{1}{C_1 C_2}} \tag{9}$$

$$Q = \frac{1}{2} \sqrt{\frac{C_1}{C_2}} \tag{10}$$

From (9) and (10), the parameter ω_o can be tuned by adjusting the value of I_b whereas the parameter Q can be given by adjusting the ratio of C_1/C_2 . Therefore, the proposed filter can be controlled orthogonally for parameters ω_o and Q , but it cannot be controlled independently. It should be noted from (9) that if the bipolar implementation of CCFTA is used, parameter ω_o can be controlled linearity. For IC implementation, adjusting the value of capacitor for obtaining desired high Q -value is difficult, but it can be resolved using a capacitor bank formed by parallelly connected capacitors with switches. The value of the capacitor can then be varied by setting the switches.

3 Non-ideal analysis

In this section, the effects of CCFTA non idealities on the proposed filter performances have been analyzed. Taking into account the non-idealities of CCFTA, the CCFTA non-idealities can be obtained from

$$\begin{pmatrix} I'_z \\ I'_{z_c} \\ I'_x \\ V'_f \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 & \beta_z \\ 0 & 0 & 0 & \beta_{z_c} \\ g_m & 0 & 0 & 0 \\ 0 & 0 & 0 & R_f \end{pmatrix} \begin{pmatrix} V_z \\ V_{z_c} \\ V_x \\ I_f \end{pmatrix} \tag{11}$$

where β_z and β_{z_c} are respectively the non-ideal current transfer gains between f-z and f-zc terminals of the CCFTA.

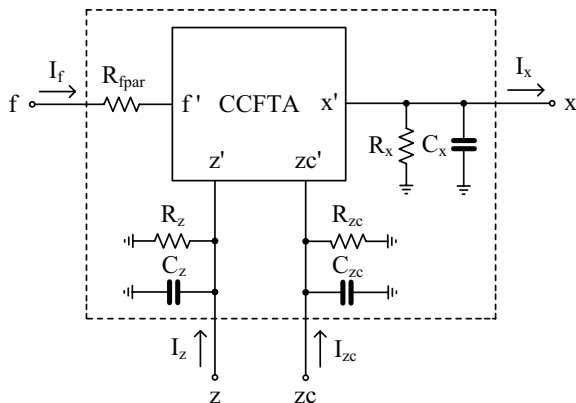


Figure 4: CCFTA with its parasitic components.

The non-ideal CCFTA symbol including various parasitic elements is shown in Fig. 4. The f-terminal exhibits parasitic serial resistance R_{fpar} , the z-terminal exhibits high-value parasitic resistances R_z in parallel with low-value parasitic capacitance C_z , the zc-terminal exhibits high-value parasitic resistance R_{zc} in parallel with low-value parasitic capacitance C_{zc} and the x-terminal exhibits high-value parasitic resistance R_x in parallel with low-value parasitic capacitance C_x .

The non-ideality of transconductance gain g_{mn} of CCFTA can be expressed as

$$g_{mn} = \frac{g_m \omega_g}{s + \omega_g} \tag{12}$$

where ω_g denotes the first-order pole of the transconductance amplifier. In the frequency range of our interest, g_{mn} is modified to [59]

$$g_{mn} \cong g_m (1 - \mu s) \tag{13}$$

where $\mu = 1/\omega_g$.

Equations (11), (13) now result in Fig. 4, the current I_{out} of Fig. 3 can be given by

$$I_{out} = \frac{D(s) I_{in3} - \beta_z R_x g_{mn} I_{in2} - \beta_{z_c} R_x (s C'_2 R_z + 1) I_{in1}}{D(s)} \tag{14}$$

$$\begin{aligned} D(s) &= (s^2 C'_1 C'_2 R_x R_z + s C'_1 R_x + s C'_2 R_z + 1) R_f + \\ &+ (s C'_2 R_z + 1) R_x + g_{mn} R_x R_z \\ &= s^2 C'_1 C'_2 R_f R_x R_z + s C'_1 R_f R_x + s C'_2 R_f R_z + \\ &+ R_f + s C'_2 R_z R_x + R_x + g_{mn} R_x R_z \\ &= s^2 (C'_1 C'_2 R_f R_x R_z) + s (C'_1 R_f R_x + C'_2 R_f R_z + C'_2 R_z R_x) + \\ &+ (g_{mn} R_x R_z + R_f + R_x) \end{aligned} \tag{15}$$

Letting $R_f \ll R_x$ and $R_f \ll R_z$, (15) becomes

$$\left. \begin{aligned} D(s) &= s^2 + s \left(\frac{1}{C'_2 R_z} + \frac{1}{C'_1 R_x} + \frac{1}{C'_1 R_f} \right) + \frac{g_{mn}}{C'_1 C'_2 R_f} \\ &= s^2 + s \frac{1}{C'_1 R_f} + \frac{g_{mn}}{C'_1 C'_2 R_f} \end{aligned} \right\} \tag{16}$$

where $C'_1 = C_1 \parallel C_x$ and $C'_2 = C_2 \parallel C_z$.

From (16) we can see that CCFTA non-idealities affect the circuit characteristics which depart from ideal values. To prevent significant errors, the value of the capacitors C_1 and C_2 should be selected to meet the con-

ditions $C_1 \gg C_x$ and $C_2 \gg C_z$. The non-ideal values of parameters ω_o and Q can be expressed as

$$\omega_o = \sqrt{\frac{g_{mn}}{C_1' C_2' R_f}} \tag{17}$$

$$Q = \sqrt{R_f g_{mn} \left(\frac{C_1'}{C_2} \right)} \tag{18}$$

It should be noted from (17) and (18) that the parameters ω_o and Q are slightly changed by the non-idealities of the CCFTA. However, these effects can be compensated by adjusting the g_m -value. The active and passive sensitivities of the filter parameters are

$$S_{g_{mn}}^{\omega_o} = -S_{C_1}^{\omega_o} = -S_{C_2}^{\omega_o} = -S_{R_f}^{\omega_o} = 0.5 \tag{19}$$

$$S_{R_f}^Q = S_{g_{mn}}^Q = S_{R_z}^Q = S_{C_1}^Q = -S_{C_2}^Q = 0.5 \tag{20}$$

From (19)-(20) we can see that the incremental sensitivities of the active and passive parameters do not exceed 1 in magnitude. Hence, the proposed filter offers low active and passive sensitivities.

4 Application to sixth-order filters

It is well-known that biquadratic filters can be used to realize high-order filters [2]. To confirm the applicability of the proposed universal biquadratic filter, high-order filters using the proposed biquadratic filter are designed. The structures of high-order filters such as sixth-order Butterworth LP, HP and BP filters have been designed. Sixth-order Butterworth LP filter can be designed by cascading tree second-order LP filters. In case of second-order HP filter, additional multiple-input current follower (CF) is required. Fig. 5(a) shows the block diagram of a second-order HP filter. The bipolar implementation of multiple-output CF is shown in Fig. 5(b). Sixth-order Butterworth BP filter can be obtained by cascading a sixth-order Butterworth HP filter and a sixth-order Butterworth LP filter. The block diagram is depicted in Fig. 6. To obtain a sixth-order Butterworth LP and HP characteristics, the filters have been designed using Tables 2 and 3. From Table 2, HP filter is designed for the cut-off frequency of 1 MHz while the LP filter is designed for the cut-off frequency of 3 MHz. Using the bias currents and capacitor-values as shown in Tables 2 and 3, sixth-order Butterworth BP filter can be obtained with the bandwidth of 2 MHz.

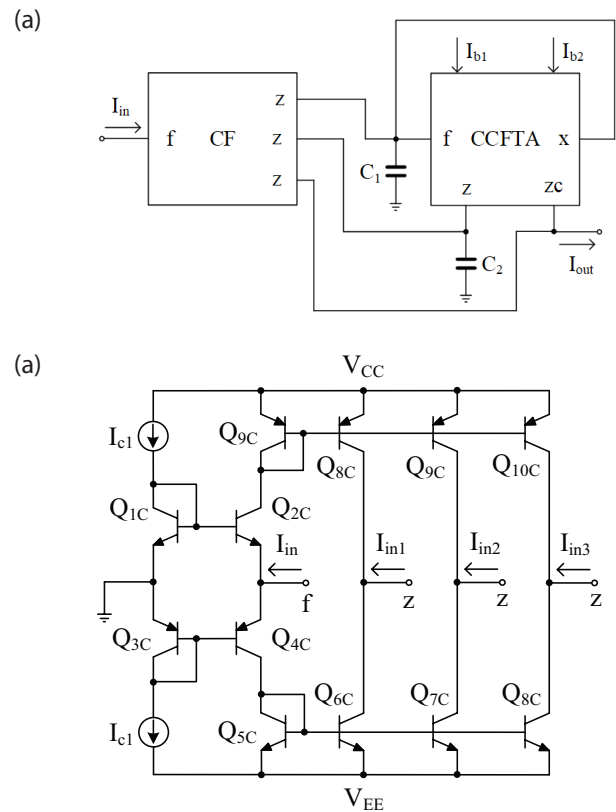


Figure 5: (a) second-order high-pass filter, (b) multiple-output current follower.

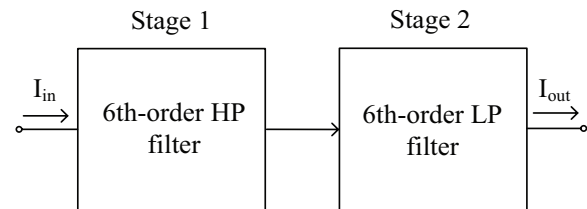


Figure 6: Sixth-order band-pass Butterworth filter.

Table 2: Parameters for sixth-order Butterworth LP filter.

Second-order LP filter	Value ($f_o = 3$ MHz)
Stage 1 (FSF = 1.000, Q = 0.5177)	$I_{b1} = I_{b2} = 50.7 \mu A$, $C_1 = 107$ pF, $C_2 = 100$ pF
Stage 2 (FSF = 1.000, Q = 0.7071)	$I_{b1} = I_{b2} = 69.3 \mu A$, $C_1 = 200$ pF, $C_2 = 100$ pF
Stage 3 (FSF = 1.000, Q = 1.9320)	$I_{b1} = I_{b2} = 94.68 \mu A$, $C_1 = 746.5$ pF, $C_2 = 50$ pF

Table 3: Parameters for sixth-order Butterworth HP filter.

Second-order HP filter	Value ($f_o = 1$ MHz)
Stage 1 (FSF = 1.000, Q = 0.5177)	$I_{b1} = I_{b2} = 33.79 \mu\text{A}$, $C_1 = 214 \text{ pF}$, $C_2 = 200 \text{ pF}$
Stage 2 (FSF = 1.000, Q = 0.7071)	$I_{b1} = I_{b2} = 46.2 \mu\text{A}$, $C_1 = 400 \text{ pF}$, $C_2 = 200 \text{ pF}$
Stage 3 (FSF = 1.000, Q = 1.9320)	$I_{b1} = I_{b2} = 126.2 \mu\text{A}$, $C_1 = 2.98 \text{ nF}$, $C_2 = 200 \text{ pF}$

5 Simulation results

The proposed filters are verified with PSPICE simulations. The CCFTA in Fig. 2 was implemented with bipolar transistor array HFA3096 [60]. The supply voltages were $V_{CC} = -V_{EE} = 3$ V. Simulated performance of CCFTA is given in Table 4.

Fig. 7 depicts simulated frequency responses of LP, BP, HP and BS filters with $I_{b1} = I_{b2} = 90 \mu\text{A}$, $C_1 = 3000 \text{ pF}$ and $C_2 = 100 \text{ pF}$, resulting in a natural angular frequency $f_o \cong 1$ MHz and $Q \cong 2.73$. Fig. 8 shows simulated frequency responses of the magnitude and phase characteristics of the AP filter at $f_o \cong 1$ MHz. It is clear from Figs. 7 and 8 that the proposed filter performs five standard filtering functions such as LP, BP, HP BS and AP filters with a single topology. Fig. 9 shows the simulated frequency response of a BP filter when the bias currents I_b (i.e., $I_b = I_{b1} = I_{b2}$) were simultaneously adjusted to 20, 50, 100 and 200 μA , respectively, while keeping $C_1 = 3000 \text{ pF}$ and $C_2 = 100 \text{ pF}$ for a constant $Q \cong 2.73$. This simulation result confirms (9).

Table 4: Simulated parameters of CCFTA.

Parameters	Value
Technology	Bipolar
Supply voltage	± 3 V
R_f ($I_{b1} = 1\text{-}250 \mu\text{A}$)	12k to 52 [Ω]
g_m ($I_{b2} = 1\text{-}250 \mu\text{A}$)	21 μ to 51m [A/V]
Current gain (I_z/I_f)	0.991
Current gain (I_{zc}/I_f)	0.991
-3dB bandwidth: I_z/I_f	
I_{zc}/I_f	191 MHz
I_x/I_f ($I_{b1} = I_{b2} = 100 \mu\text{A}$, $R_z = 495 \Omega$)	191 MHz 66 MHz
R_z/C_z @ $I_{b1} = 100 \mu\text{A}$	63 k Ω // 0.792 pF
R_{zc}/C_{zc} @ $I_{b1} = 100 \mu\text{A}$	184 k Ω // 0.304 pF
R_f/L_f @ $I_{b1} = 100 \mu\text{A}$	128 Ω // 0.127 nH
R_x/C_x @ $I_{b2} = 100 \mu\text{A}$	363 k Ω // 0.208 pF
Static power dissipation	
@ $I_{b1} = I_{b2} = 1 \mu\text{A}$	43.5 μW
@ $I_{b1} = I_{b2} = 250 \mu\text{A}$	108 mW

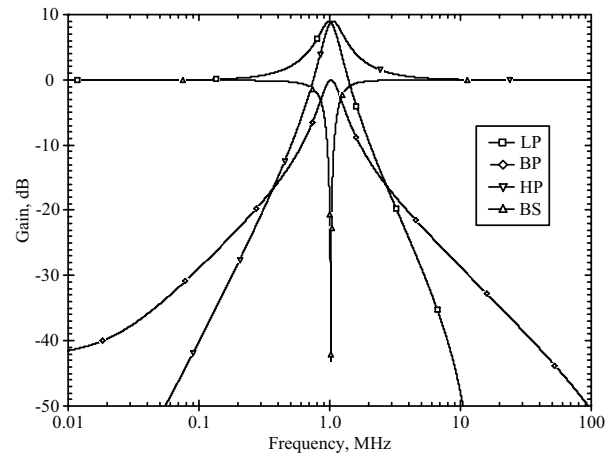


Figure 7: Simulated frequency responses of LP, BP, HP and BS filters.

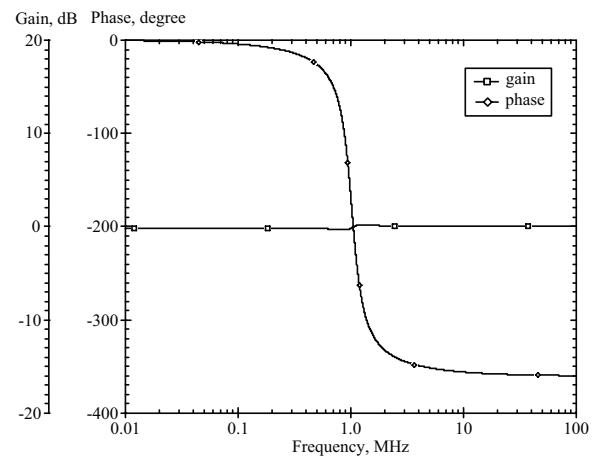


Figure 8: Simulated magnitude and phase response of an AP filter.

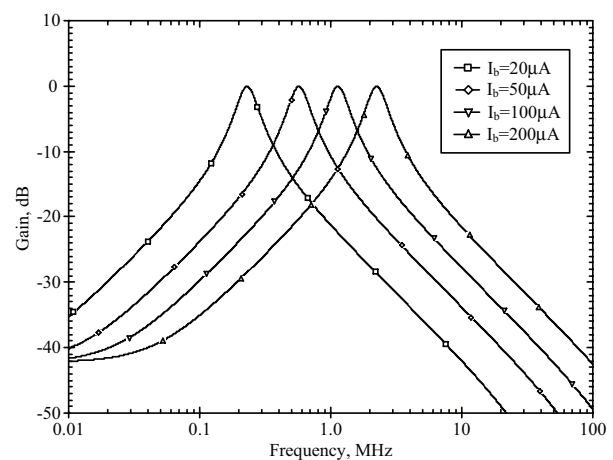


Figure 9: Frequency responses of a BP filter when I_b is varied.

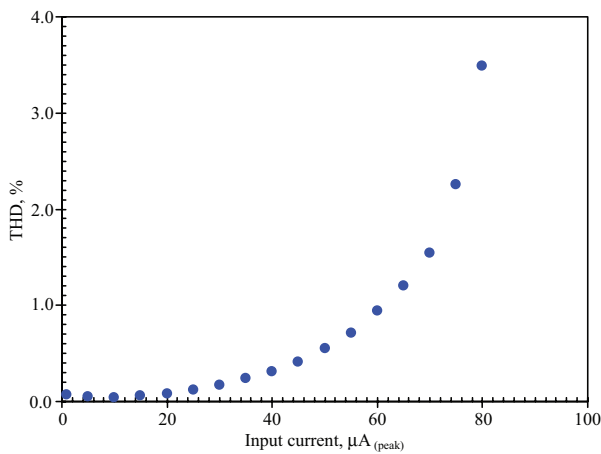


Figure 10: Dependence of the output harmonic distortion of LP filter on the input current amplitude for a 100 kHz input signal.

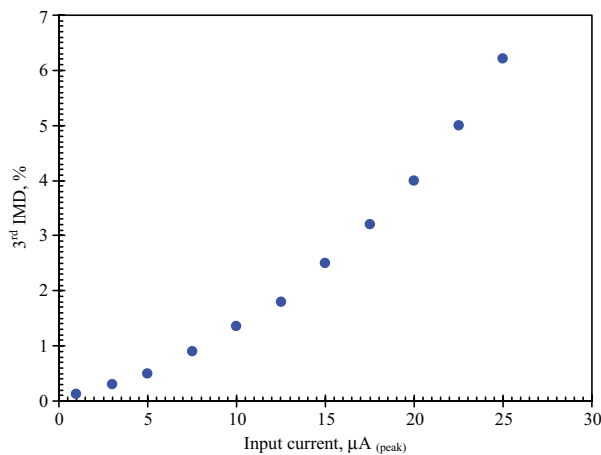


Figure 11: Dependence of the 3rd IMD of BP filter on input current amplitudes.

In order to test the linearity of the proposed filter, two methods were used; single-tone and two-tone tests. A single-tone test was performed by applying a sinusoidal signal of $f_o = 100$ kHz at the input of a LP filter. The dependence of the output harmonic distortion on the input amplitude is shown in Fig. 10. From this result, the THD was about 1.2 % when the input signal was 65 μA (peak) and it increases to 3.49 % when input signal increases to 80 μA (peak). A two-tone test was performed on the BP filter by applying two closely spaced tones with equal input signal amplitudes simultaneously at the input of BP filter. Fig. 11 shows the dependence of the 3rd IMD (intermodulation distortion) of BP filter on the input signals amplitudes. The two closely spaced tones with $f_1 = 0.8$ MHz and $f_2 = 1.2$ MHz had the same amplitude. It shows that the 3rd IMD is 6.2 % for the input signals amplitude of 25 μA (peak). The proposed filter was investigated using a Monte-Carlo analysis. The simulation test was the fluctuation of f_o changes caused

by the deviation of the capacitors. In this test, the BP filter was simulated for 5 % tolerances of capacitors C_1 and C_2 at $f_o = 1$ MHz, $Q \cong 2.73$ and 200 Gaussian distribution runs. Fig. 12 shows the derived histogram of f_o . The standard deviation (σ) of f_o was 31.96 kHz and the minimal and maximal values of f_o were 0.928 MHz and 1.1 MHz, respectively.

From (9) we can see that ω_o depends on V_T which in turn depends on the absolute temperature. Thus temperature stability of the proposed filter's ω_o of the proposed filter on parameter ω_o was investigated by varying temperature from 0° to 75°. The simulated frequency responses of the BP filter corresponding to different temperatures are depicted in Fig. 13.

When temperature varied between 0 and 75°, the corresponding f_o varied between 1.1 MHz and 0.879 MHz. This effect is expressed by (7). This problem can be solved by using a bias current source with the current proportional to the absolute temperature [61].

Three sixth-order Butterworth filters were also tested using parameters given in Tables 2 and 3. Simulated frequency responses of sixth-order Butterworth LP and HP filters are depicted in Figs. 14 and 15, respectively. The cut-off frequencies of 3 MHz and 1 MHz were obtained. The sixth-order Butterworth BP filter was simulated and the result is depicted in Fig. 16. The bandwidth (BW) of 2 MHz was expressed. The power consumptions for sixth-order Butterworth LP, HP and BP filters were 9.05 mW, 18.5 mW and 82.4 mW, respectively.

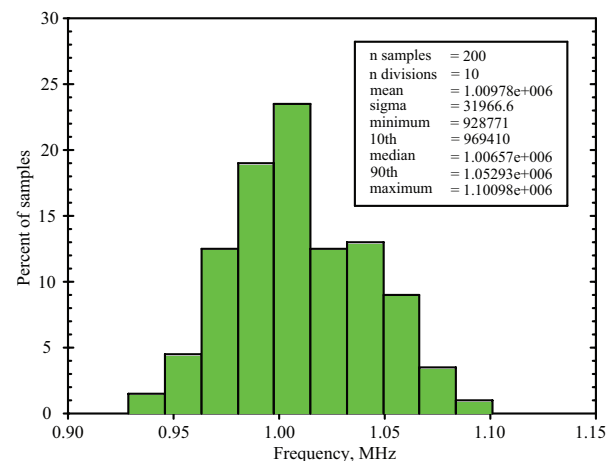


Figure 12: Histogram of Monte-Carlo analysis for BP filter with a 5% variation of capacitors C_1 and C_2 .

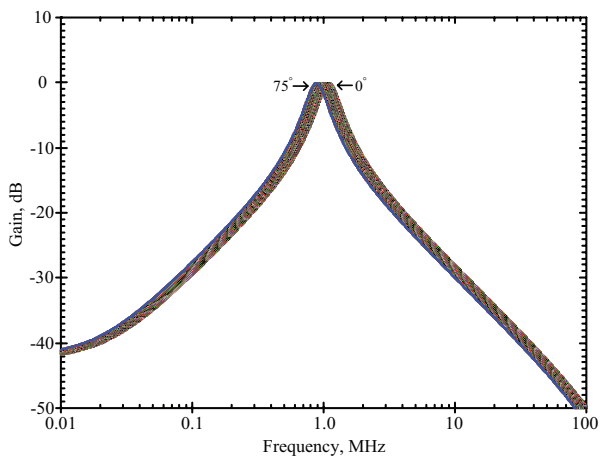


Figure 13: Simulated frequency responses of the BP filter for different temperatures.

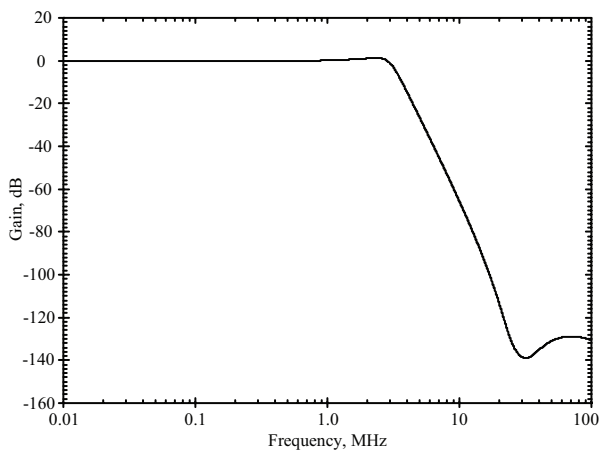


Figure 14: Simulated magnitude response of the sixth-order Butterworth LP filter.

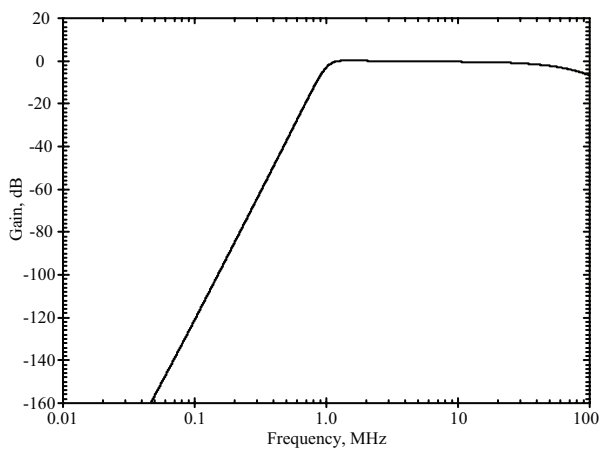


Figure 15: Simulated magnitude response of the sixth-order Butterworth HP filter.

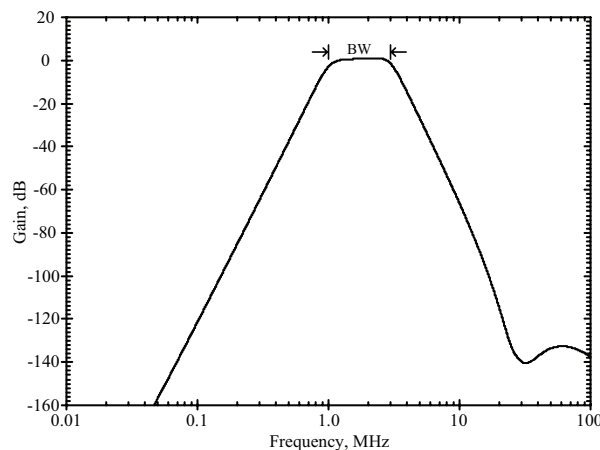


Figure 16: Simulated magnitude response of the sixth-order Butterworth BP filter.

6 Conclusions

In this paper, a new electronically tunable current-mode multifunction biquadratic filter employing one CCFTA and two grounded capacitors is presented. The proposed filter offers the following properties: (i) employment of grounded capacitors which is ideal for IC implementation; (ii) ability to implement LP, BP, HP, BS and AP filter responses without inverted input signals and passive component-matching conditions; (iii) orthogonal control of parameters ω_o and Q; (iv) current-controlled of parameter ω_o ; (v) high output impedance output which can be directly connected to next the stage and (vi) low active and passive sensitivities. The proposed biquadratic filter has been used to implement high-order filters such as sixth-order Butterworth LP, HP and BP filters to confirm the applicability of the presented structure. Simulation results confirm the performance of the proposed filters.

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