ECCCII-Based Current-Mode Universal Filter with Orthogonal Control of ω_o and Q

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Abstract. *This paper presents a new current-mode current-controlled four-input five-output universal filter employing one current-controlled current conveyor (CCCII), one electronically tunable CCCII and two grounded capacitors. The proposed configuration provides lowpass, bandpass, highpass, bandstop and allpass current responses that taken from the high-output impedance terminals, which enable easy cascadability of the current-mode operation. The filter also offers both orthogonal and electronic controls of the natural frequency and the quality factor through adjusting the bias current of the CCCIIs. For realizing all the filter responses, the proposed filter does not require passive component-matching condition and both active and passive sensitivities are low. In addition, a new current-mode current-controlled single-input five-output universal filter can be achieved by using an additional multiple-output minus-type CCCII. The proposed filter is simulated using PSPICE simulations to confirm the theoretical analysis.*

Keywords

Universal filter, current-mode circuit, electronically tunable current conveyor, analog filter.

1. Introduction

In the present, current-mode signal processing circuits based on second generation current-conveyors (CCIIs) have received considerable attention owing to the fact that their bandwidth, linearity and dynamic range performances are better than op-amp-based voltage-mode signal processing circuits [1], [2]. Especially, current-mode circuit provides simple summation/subtraction signals of currents in node, which results to simple circuitry. Many current-mode universal filters using current conveyors as active elements have been proposed [3]-[43]. A number of CCII-based current-mode universal filters have been presented [3]-[9]. In [3]-[6] single-input multiple-output (SIMO) current-mode universal filters using CCIIs have been reported. Generally, SIMO filter can simultaneously realize three basic filter functions, i.e. lowpass (LP), bandpass (BP) and highpass (HP). However, for the realizations of allpass (AP) and bandstop (BS) functions, some of SIMO filters require passive or signal-matching condition. For more convenience and versatility, the multiple-input multiple-output (MIMO) universal filters could be used. The employment of the MIMO configuration may lead to a reduction of number of active elements for circuit realization. This type of filter, in comparison with the SIMO filter, provides a variety of circuit characteristic with different input and output currents, and usually does not require any parameter matching conditions. Moreover, to realize all the standard biquadratic filter functions, the configuration with multiple inputs seems to be more suitable than the single input configuration [11]. In [7]-[9] CCIIs-based MIMO current-mode universal filter have been presented. Most of the MIMO universal filters based on CCIIs suffer from lack of electronic tuning. By using the second-generation current-controlled current conveyor (CCCII) introduced by Fabre et al. [10], current conveyor applications can be extended to the domain of electronically tunable functions. Many current-mode CCCII-based universal filters have been proposed in the technical literatures [11]-[42]. In [11]-[16] multiple-input universal filters were proposed while single-input universal filters were proposed in [17]-[41]. However, some of these filters do not benefit from orthogonal control of the natural frequency ω_0 and the quality factor Q.

In this paper, we propose a new electronically tunable current-mode four-input five-output universal filter using one CCCII, one electronically tunable CCCII and two grounded capacitors [42], which is advantageous in view of integrated circuit implementation. The proposed circuits can simultaneously realize LP, BP, HP, BS and AP current responses at a high impedance output terminal permitting easy cascadability of the current-mode operation. The circuit parameters ω_0 and Q can be tuned separately and electronically by adjusting the bias current of CCCII. Moreover, high Q-value filters can be obtained. For the realization of all the filter responses no component-matching conditions are required. Both active and passive sensitivities are low. In addition, a new current-mode single-input five-output universal filter can be also achieved by adding a multiple-output minus-type CCCII into the proposed filter. This new circuit has low-input and high-output impedance levels. The comparison between the proposed circuits and some previously CCCII-based filters is summarized in Tab. 1.

2. Proposed Circuit

The well-known schematic for CCCII, implemented with bipolar technology is shown in Fig. 1 [10]. According to Fig. 1, the CCCII has a unity voltage gain between terminal y and x and a unity current gain between terminal x and z. The terminals y and z possess high impedance level and the x terminal has the R_x, which can be given by

$$R_x = \frac{V_T}{2I_o} \quad (1)$$

The R_x is an inner resistance of a translinear mixed loop (Q_1 to Q_4) with grounded resistor equivalent controlled by bias current I_o, where V_T is the thermal voltage (≈ 25 mV). The translinear current conveyor with controlled current gain can be obtained by modifying the original circuit of the CCCII in Fig. 1 and adding additional current mirror with adjustable gain as shown in Fig. 2 [44] to obtain the required current gain at z terminal. Also, the multiple-output translinear current conveyor can be obtained by adding additional current mirrors and cross-coupled current mirrors to obtain the required plus and minus type outputs, respectively [45].

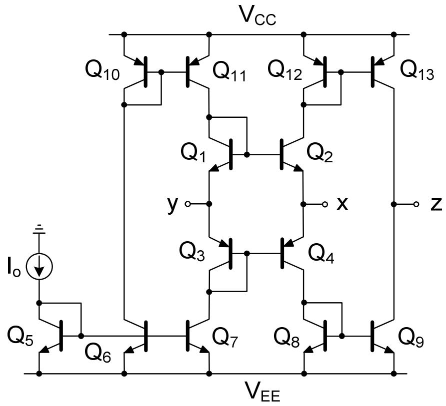


Fig. 1. Schematic implementation for CCCII.

Circuits	Type of filter	Number of active elements	Number of capacitors	All-grounded capacitor	Offer five standard filters	Orthogonal control of ω_o and Q	No need matching-condition	Low active & passive sensitivity	High output impedance
Proposed filters									
Fig. 3(a)	MIMO	2 CCCIIs	2	Yes	Yes	Yes	Yes	Yes	Yes
Fig. 3(b)	SIMO	3 CCCIIs	2	Yes	Yes	Yes	Yes	Yes	Yes
Ref. [11] in 2006	MISO	5 CCCIIs	2	Yes	Yes	Yes	No	Yes	Yes
Ref. [12] in 2007	MIMO	4 CCCIIs	2	Yes	Yes	Yes	No	Yes	Yes
Ref.[14] in 2007	MIMO	2 CCCIIs	2	Yes	Yes	No	Yes	Yes	Yes
Ref. [15] in 2010	MIMO	2 CCCIIs	2	Yes	Yes	Yes	No	Yes	Yes
Ref. [16] in 2008 (fig. 3)	MISO	2 CCCII	2	Yes	Yes	No	No	Yes	Yes
Ref. [19] in 1997	SIMO	3 CCCIIs	2	Yes	No	No	No	Yes	Yes
Ref. [20] in 1998	SIMO	6 CCCIIs	2	Yes	No	No	No	Yes	Yes
Ref. [21] in 1998	SIMO	3 CCCIIs	2	Yes	No	No	Yes	Yes	No
Ref. [23] in 2000	SIMO	3 CCCIIs	2	Yes	No	Yes	Yes	Yes	No
Ref. [26] in 2004	SIMO	4 CCCIIs	2	Yes	No	No	Yes	Yes	Yes
Ref. [27] in 2004	SIMO	2 CCCIIs	2	No	No	No	Yes	Yes	No
Ref. [28] in 2005	SIMO	3 CCCIIs	2	Yes	No	No	Yes	Yes	Yes
Ref. [29] in 2006	SIMO	2 CCCIIs	2	No	No	No	Yes	Yes	No
Ref. [30] in 2006	SIMO	4 CCCIIs	2	No	No	No	Yes	Yes	Yes
Ref. [31] in 2007	SIMO	3 CCCIIs	2	Yes	No	No	Yes	Yes	Yes
Ref. [33] in 2008	SIMO	4 CCCIIs	2	Yes	Yes	Yes	Yes	Yes	Yes
Ref. [34] in 2009	SIMO	4 CCCIIs	2	Yes	No	No	No	Yes	Yes
Ref. [35] in 2009	SIMO	5 CCCIIs	2	Yes	No	Yes	Yes	Yes	Yes
Ref. [37] in 2009	SIMO	3 CCCIIs	2	Yes	Yes	No	Yes	Yes	Yes
Ref. [38] in 2010	SIMO	3 UCC	2	Yes	Yes	No	Yes	Yes	Yes
Ref. [39] in 2010	SIMO	3 CCCII	2	Yes	Yes	No	Yes	Yes	Yes
Ref. [40] in 2010	SIMO	3 CCCIIs	2	Yes	Yes	Yes	Yes	Yes	Yes
Ref. [43] in 2004	SIMO	1 CCIII	2	No	No	No	Yes	Yes	No

Note: MISO is multiple-input and single-output.

Tab. 1. Comparison of proposed filters with those of previous circuits.

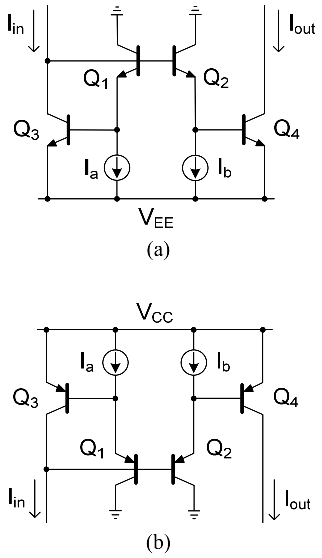


Fig. 2. Current mirrors with adjustable current gain: (a) positive-type, (b) negative-type.

Fig. 3(a) and (b) shows respectively the schematic and the symbol of the electronically tunable CCCII [42]. It has a unity voltage gain between terminals y and x and tunable k current gain between terminals x and z. The latter property makes it different from the current conveyor and hence the name electronically tunable CCCII (ECCCII). A few electronically tunable current conveyors were already described in the literature [46]-[48]. However, these devices are not suitable for electronic-control of ω_o and high Q-value biquadratic filter. The schematic of ECCCII in Fig. 3 is characterized by the relationship:

$$\begin{pmatrix} I_y \\ V_x \\ I_z \\ I_{zk} \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 \\ 1 & R_x & 0 \\ 0 & \pm 1 & 0 \\ 0 & \pm k & 0 \end{pmatrix} \begin{pmatrix} V_y \\ I_x \\ V_z \end{pmatrix} \tag{2}$$

The current gain k of the ECCCII can be expressed as [42]

$$k = \frac{I_a}{I_b} \tag{3}$$

It is evident that the signal is amplified by the factor k and this factor can be varied linearly and controlled by adjusting the I_a/I_b.

Fig. 4 shows the block diagram of the proposed universal filter. It comprises of two integrators and one amplifier. Based on Fig. 4 the proposed current-mode universal filter is shown in Fig. 5. From Fig. 5(a), CCCII_1 and C_1 operate as the first integrator ($1/s\tau_1$) and current amplifier (k) while CCCII_2 and C_2 operate as the second integrator ($1/s\tau_2$). The filter in Fig. 5(a) is consisted of one CCCII, one ECCCII and two grounded capacitors and hence the name ECCCII-based universal filter. The use of grounded capacitor makes the proposed filter ideal for integration point of view [49], [50]. Using (2) and nodal analysis, the transfer functions of the proposed biquadratic filter in Fig. 5(a) can be expressed as

$$\frac{I_{LP}}{I_{in}} = -\frac{1}{s^2 R_{x1} R_{x2} C_1 C_2 + s R_{x2} C_2 k_1 + 1}, \tag{4}$$

$$\frac{I_{BP}}{I_{in}} = -\frac{s R_{x2} C_2 k_1}{s^2 R_{x1} R_{x2} C_1 C_1 + s R_{x2} C_2 k_1 + 1}, \tag{5}$$

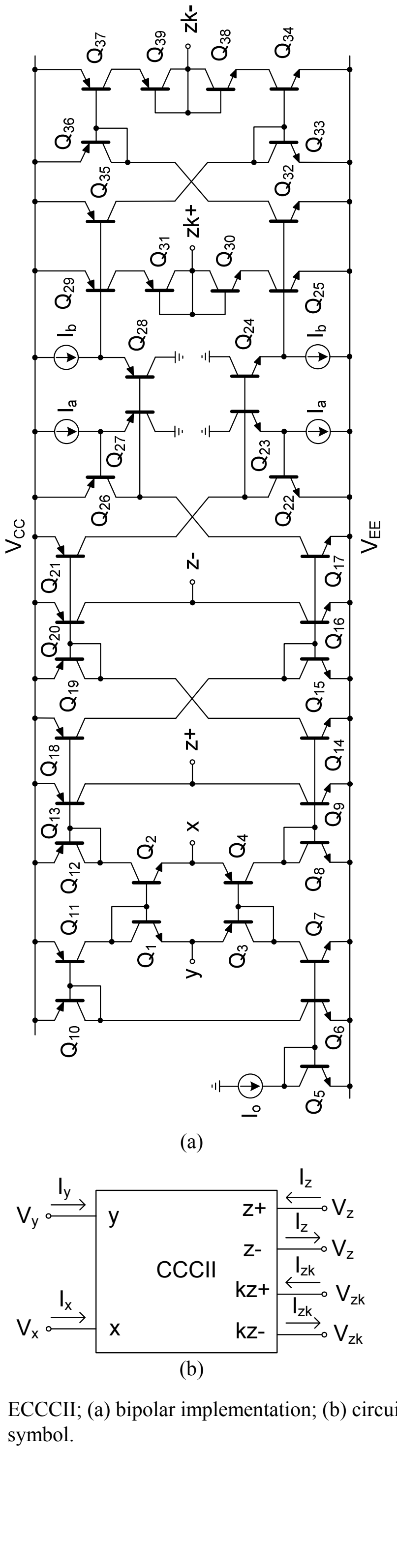


Fig. 3. ECCCII; (a) bipolar implementation; (b) circuit symbol.

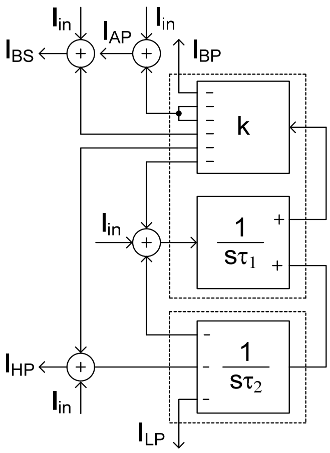


Fig. 4. Block diagram of proposed universal filter.

$$\frac{I_{HP}}{I_{in}}=\frac{s^2R_{x1}R_{x2}C_1C_2}{s^2R_{x1}R_{x2}C_1C_2+sR_{x2}C_2k_1+1}, \tag{6}$$

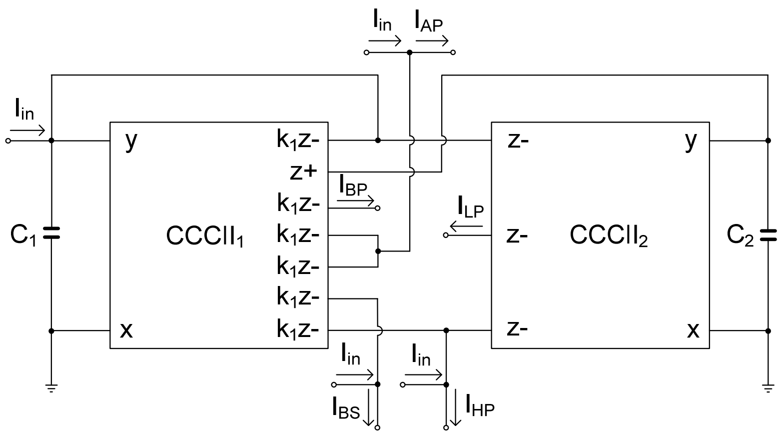
$$\frac{I_{BS}}{I_{in}}=\frac{s^2R_{x1}R_{x2}C_1C_2+1}{s^2R_{x1}R_{x2}C_1C_2+sR_{x2}C_2k_1+1}, \tag{7}$$

$$\frac{I_{AP}}{I_{in}}=\frac{s^2R_{x1}R_{x2}C_1C_2-sR_{x2}C_2k_1+1}{s^2R_{x1}R_{x2}C_1C_2+sR_{x2}C_2k_1+1}. \tag{8}$$

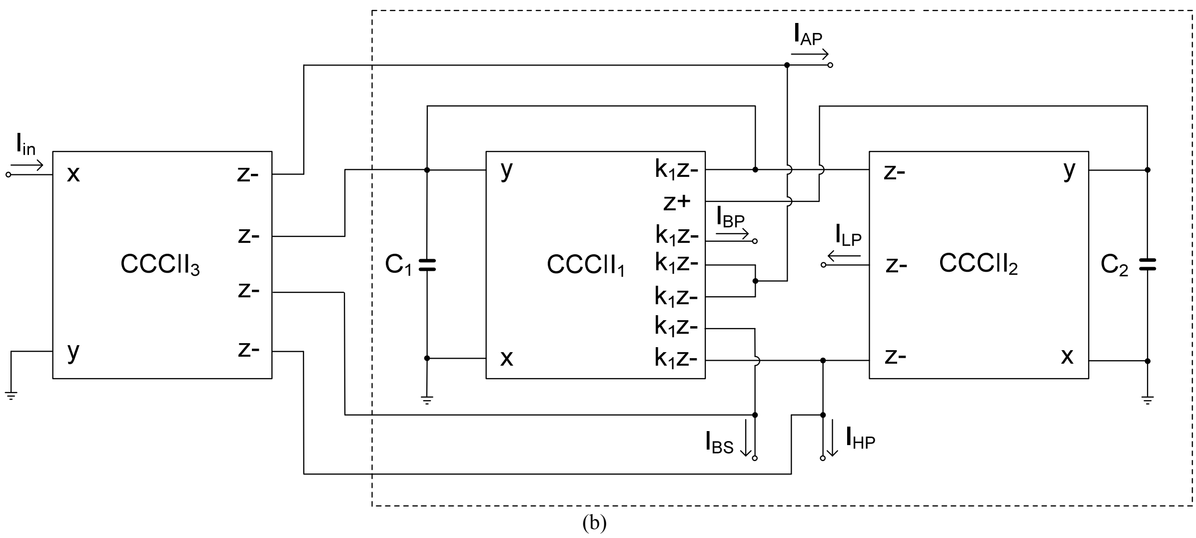
Therefore, five filter functions can be achieved. It should be mentioned that there is no component-matching condition for realizing all the standard type biquadratic filtering functions. The natural frequency ω_o and the quality factor Q are given by

$$\omega_o=\sqrt{\frac{1}{R_{x1}R_{x2}C_1C_2}}, \tag{9}$$

$$Q=\frac{1}{k_1}\sqrt{\frac{R_{x1}C_1}{R_{x2}C_2}} \tag{10}$$



(a)



(b)

Fig. 5. Proposed current-mode universal filters: (a) MIMO-type filter, (b) realization for SIMO-type filter.

with $$k_1 = \frac{I_{a1}}{I_{b1}} \qquad (11)$$

where I_{a1} and I_{b1} are the bias currents of $CCCII_1$. If we set $R_{x1} = R_{x2}$, $C_1 = C_2$ and substituting (11) into (10), equation (10) becomes

$$Q = \frac{I_{b1}}{I_{a1}}. \qquad (12)$$

From (1) and (9), the parameter ω_o for all filter responses can be electronically tuned by varying I_{o1} and/or I_{o2} without affecting the parameter Q. For the Q-value, it can be controlled linearly and separately by adjusting the ratios of the bias currents I_{b1} and I_{a1}, where the high-Q biquad can be realized when the appropriate current gain is chosen. Moreover, the Q-value is also temperature independent. This means that the proposed circuit can work as a current-tunable filter, and its parameters ω_o and Q can be independently tuned over a wide range.

Note that the requirement of the four input currents for realizing all the filter responses is not a major disadvantage. In practice, the multiple input currents can easily be obtained by using an additional multiple-output CCCII with grounded y-terminal. From Fig. 1, an input current I_{in} is then injected to the terminal x, while the four input currents can be taken from terminals z-. Fig. 5(b) shows the complete realization of universal filter which gives a transfer function of the form given by (4) to (8). From Fig. 5(b), it can be seen that the proposed universal filter employed only three CCCIIs and two grounded capacitors. Moreover, the input signal I_{in} is connected to the low input impedance terminal of the CCCII. Therefore, the second proposed circuit provides the advantage of having low-input and high-output impedance.

3. Non-ideal Effects

To consider the non-ideal effect of a CCCII by taking the non-idealities of the CCCIIs into account, the relationship of the terminal voltages and currents can be rewritten as

$$\begin{pmatrix} I_Y \\ V_X \\ I_Z \\ I_{Zk} \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 \\ \alpha_j & R_x & 0 \\ 0 & \pm\beta_j & 0 \\ 0 & \pm\beta_{kj}k_j & 0 \end{pmatrix} \begin{pmatrix} V_Y \\ I_X \\ V_Z \end{pmatrix} \qquad (13)$$

where $\alpha_j = 1 - \varepsilon_{vj}$ and $\varepsilon_{vj}(\varepsilon_{vj} \ll 1)$ is the voltage tracking error from V_y terminal to V_x terminal of the j-th CCCII, $\beta_j = 1 - \varepsilon_{ij}$ and $\varepsilon_{ij}(\varepsilon_{ij} \ll 1)$ is the output current tracking error of the j-th CCCII and $\beta_{kj} = 1 - \varepsilon_{ij}$ and $\varepsilon_{ij}(\varepsilon_{ij} \ll 1)$ is the output current tracking error of the j-th CCCII.

With regards to the effect of parasitic parameters, it can be evaluated using the non-ideal CCCII symbol as shown in Fig. 6. Typically, R_y, $R_{z\pm}$ and $R_{zk\pm}$ possess high value parasitic resistance while C_y, $C_{z\pm}$ and $C_{zk\pm}$ possess low parasitic capacitance. Using (13) and taking into account the non-ideal CCCII characteristics as shown in Fig. 6, the modified demonstrate $D(s)$ of (4)-(8) can be expressed as

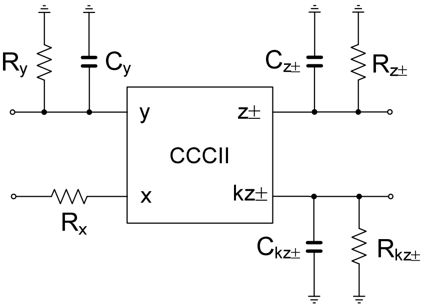


Fig. 6. Simplified equivalent circuit of the non-ideal CCCII.

X	$S_x^{\omega_o}$	S_x^Q
R_{x1}	-0.5	0.5
R_{x2}	-0.5	-0.5
C_1'	-0.5	0.5
C_2'	-0.5	-0.5
α_1	0.5	-0.5
α_2	0.5	0.5
β_1	0.5	0.5
β_2	0.5	0.5
k_1	0.0	-1.0
β_{k1}	0.0	-1.0

Tab. 2. Sensitivities of circuit components.

$$D(s) = \frac{s^2}{n_1 n_2} + \alpha_1\beta_{k1}k_1\frac{s}{n_2} + \alpha_1\alpha_2\beta_1\beta_2 \qquad (14)$$

where $$n_1 = \left(\frac{1}{R_{x1}C_1'}\right)\left(\frac{s}{s+\omega_1}\right), \qquad (15)$$

$$n_2 = \left(\frac{1}{R_{x2}C_2'}\right)\left(\frac{s}{s+\omega_2}\right), \qquad (16)$$

$\omega_1 = \frac{1}{R_{n1}C_1'}$, $\omega_2 = \frac{1}{R_{n2}C_2'}$, $C_1' = C_1 + C_{y1} + C_{k1z1-} + C_{z2-}$, $C_2' = C_2 + C_{y2} + C_{z1+}$, $R_{n1} = R_{y1} // R_{k1z1-} // R_{z2-}$, $R_{n2} = R_{y2} // R_{z1+}$. Form (14)-(16), the effect of the parasitic parameters of CCCII are dependent on two parasitic poles. The frequency of operation should be larger than ω_1 and ω_2 for close to ideal operation at high frequency. Using PSPICE simulations, the parasitic elements of the CCCII for $I_o = 32$ μA can be calculated as $C_{y1} = C_{y2} = 2.2$ pF, $C_{z1+} = C_{z2-} \approx 1.67$ pF, $C_{k1z1-} = 1.1$ pF, $R_{y1} = R_{y2} = 358$ kΩ, $R_{z2-} = 350$ kΩ, $R_{z1+} = 366$ kΩ, and $R_{k1z1-} = 151$ kΩ. For example, if the proposed filter was designed for $f_o = 1$ MHz by choosing $C_1 = C_2 = 0.4$ nF and $R_{x1} = R_{x2} \approx 394$ Ω, then the parasitic pole f_1 will be located at approximately

7.5 kHz and f_2 will be located at approximately 2.1 kHz. Thus, the ideal operation of the filter will be valid for frequencies higher than approximately $10f_1$ [9], [10]. Assume that ω_1 and $\omega_2 \gg 1$, the parameters ω_o and Q can be rewritten by

$$\omega_o = \sqrt{\frac{\alpha_1\alpha_2\beta_1\beta_2}{R_{x1}R_{x2}C_1'C_2'}}\,, \tag{17}$$

$$Q = \frac{1}{k_1\beta_{k1}}\sqrt{\frac{R_{x1}C_1'}{R_{x2}C_2'}\frac{\alpha_2\beta_1\beta_2}{\alpha_1}}\,. \tag{18}$$

From (17) and (18), the tracking errors slightly change the natural frequency and the quality factor. However, the parameters ω_o and Q can still be orthogonally controllable. The incremental sensitivities of the parameters ω_o and Q are calculated in Tab. 2. It can be indicated that the sensitivities of the proposed filter are low.

4. Simulation Results

In order to verify the characteristics of the proposed filter in Fig. 5, PSPICE simulators were carried out. The CCCIIs were performed with the transistor model of HFA3046 as listed in Tab. 3 and the DC supply voltage $V_{CC} = -V_{EE} = 3$ V. When the ECCCII in Fig. 3 was simulated by PSPICE simulators, the parameters $\alpha = 0.998$, $\beta_+ = 0.995$ and $\beta_- = 1.065$ were obtained and other simulated results were presented in Tab. 4. The frequency response of the ECCCII with different bias current I_a of 50, 250, 500, 700 to 1000 μA was investigated and shown in Fig. 7. As an example design, $C_1 = C_2 = 0.4$ nF was given. Fig. 8 shows the simulated frequency responses of the filter with $I_{o1} = I_{o2} = 32$ μA and $I_{a1} = I_{b1} = 50$ μA. This setting was designed to obtain the LP, BP, HP and BS filter responses with $f_o \cong 1.02$ MHz and $Q = 1$. Fig. 9 shows the simulated frequency responses of the gain and phase characteristics of the AP filter at $f_o \cong 1$ MHz. It was clear from both figures that the proposed filter performs five standard biquadratic filtering functions well. In Figs. 8 and 9, the pole frequency of 1 MHz was obtained. The pole frequency was 1 MHz instead of 1.02 MHz owing to the effect of non-ideal CCCIIs. Fig. 10 shows the simulated BP filter response when the DC bias currents I_o (i.e. $I_o = I_{o1} = I_{o2}$) were simultaneously adjusted for the values 15, 25, 40 and 60 μA, respectively, while keeping $I_{a1} = I_{b1} = 50$ μA for $Q = 1$. This result was confirmed by (9). To demonstrate the current gain of current conveyor tuning of Q, the bias currents were set to be constant at $I_{o1} = I_{o2} = 32$ μA and $I_{a1} = 50$ μA. Fig. 11 shows the corresponding current characteristics of the BP filter when I_{b1} was varied. It is obvious that high values of the Q can be easily obtained from high ratios of current gain as was confirmed by (12). From Figs. 8 to 11 it is evident that the simulation results agree quite well with the theoretical analysis. Therefore, the proposed filter easily obtains separate electronic control of parameter ω_o and Q and high-Q value filter by using ECCCII. Recently, a new active device, so-called double current controlled current feedback amplifier (DCC-CFA) was proposed [51], [52]. This device provides the current gain that can be used to realize the filter with separate control of parameter ω_o and Q and high-Q value. However, this DCC-CFA was realized using bipolar and CMOS technologies and thus BiCMOS technology was needed for integrated circuit implementation which is difficult as a fabrication process.

.model NUHFARRY NPN (IS=1.840E-16 XTI=3.000E+00 EG=1.110E+00 VAF=7.200E+01 VAR= 4.500E+00 BF=1.036E+02 ISE=1.686E-19 NE=1.400E+00 IKF= 5.400E-02 XTB=0.000E+00 BR= 1.000E+01 ISC=1.605E-14 NC=1.800E+00 IKR=5.400E-02 RC=1.140E+01 CJC=3.980E-13 MJC= 2.400E-01 VJC=9.700E-01 FC=5.000E-01 CJE=2.400E-13 MJE=5.100E-01 VJE=8.690E-01 TR=4.000E-09 TF=10.51E-12 ITF=3.500E-02 XTF=2.300E+00 VTF=3.500E+00 PTF=0.000E+00 XCJC=9.000E-01 CJS=1.150E-13 VJS=7.500E-01 MJS=0.000E+00 RE=1.848E+00 RB=5.007E+01 RBM=1.974E+00 KF=0.000E+00 AF=1.000E+00)
.model PUHFARRY PNP (IS=1.027E-16 XTI=3.000E+00 EG=1.110E+00 VAF=3.000E+01 VAR=4.500E+00 BF=5.228E+01 ISE=9.398E-20 NE=1.400E+0 IKF=5.412E-02 XTB=0.000E+00 BR= 7.000E+00 ISC=1.027E-14 NC=1.800E+00 IKR=5.412E-02 RC=3.420E+01 CJC=4.951E-13 MJC=3.000E-01 VJC=1.230E+00 FC=5.000E-01 CJE=2.927E-13 MJE=5.700E-01 VJE=8.800E-01 TR=4.000E-09 TF=20.05E-12 ITF=2.001E-02 XTF=1.534E+00 VTF= 1.800E+00 PTF=0.000E+00 XCJC=9.000E-01 CJS=1.150E-13 VJS=7.500E-01 MJS=0.000E+00 RE=1.848E+00 RB=3.271E+01 RBM=9.902E-01 KF=0.000E+00 AF=1.000E+00)

Tab. 3. Parameter of HFA3046 transistor array.

Parameters	Value
Power supply	± 3 V
DC voltage range	-2 to 2 V
Voltage error range	<7 mV
DC current range	-25 to 25 mA
Bandwidth (-3dB) @ [I_o= 50 μA, I_a= 50 μA, I_b= 50 μA]	
Voltage follower (V_x/V_y)	370 MHz
Current follower (I_z/I_x)	177 MHz
Current follower (I_{zk}/I_x)	112 MHz
R_x, L_x @ [I_o =50 μA]	253 Ω, 0.73 μH
R_y, C_y	358 kΩ, 2.2 pF
R_{z+}, C_{z+}	366 kΩ, 1.67 pF
R_{zk}, C_{zk} @ [I_a=I_b=50 μA]	150 kΩ, 1.19 pF
R_{zk}, C_{zk} @ [I_a=1000 μA, I_b=50 μA]	13.59 kΩ, 0.39 pF
R_{zk}, C_{zk} @ [I_a=50 μA, I_b=1000 μA]	3.6 MΩ, 0.23 pF

Tab. 4. Simulated specifications of CCCII used.

In order to test the time-domain of the proposed filter, a LP filter at 1 MHz of cut-off frequency was simulated. In this case, a 100 kHz in-band of LP filter was selected. Fig. 12 shows the time domain responses for LP filter when a 100 kHz sinusoidal input current with 100 μA peak was applied to the filter. It was observed that 100 μA peak

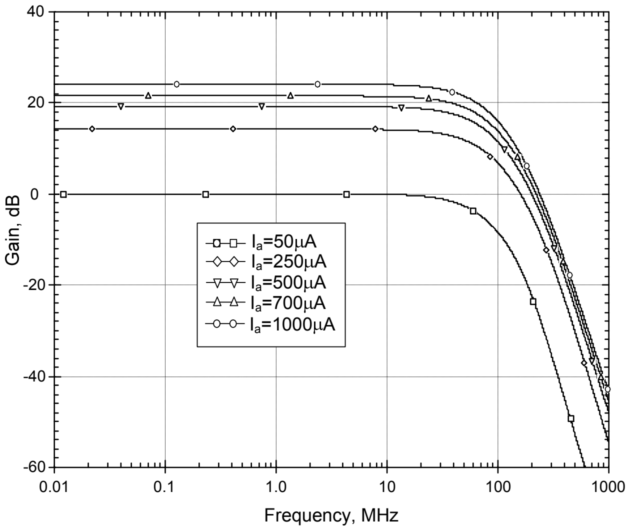


Fig. 7. Simulated response of current gain k.

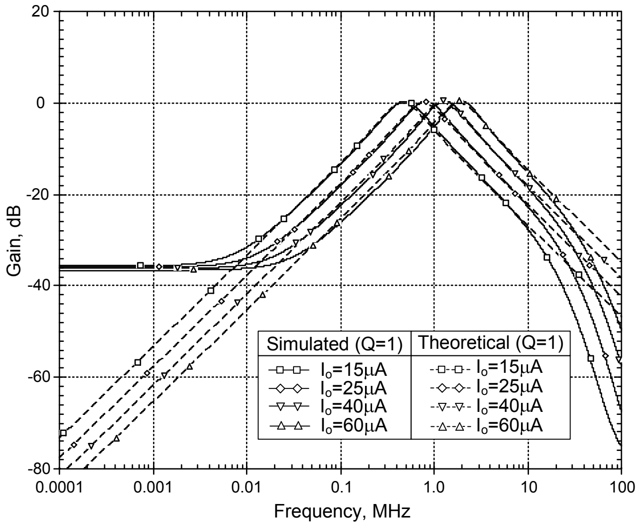


Fig. 10. Simulated frequency responses of the BP filter when I_o is varied.

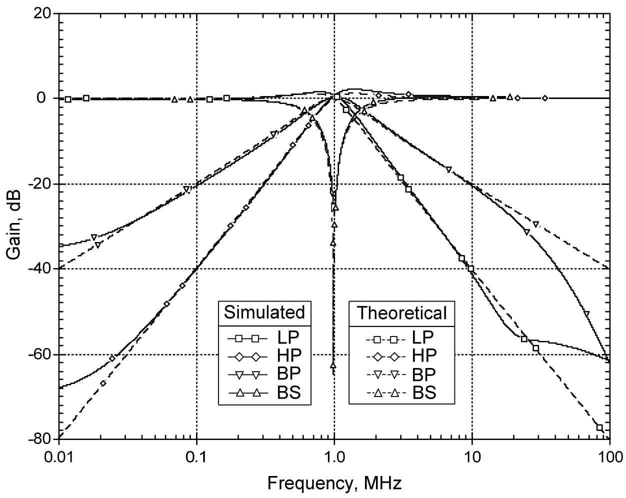


Fig. 8. Simulated LP, BP, HP and BS of the proposed filter.

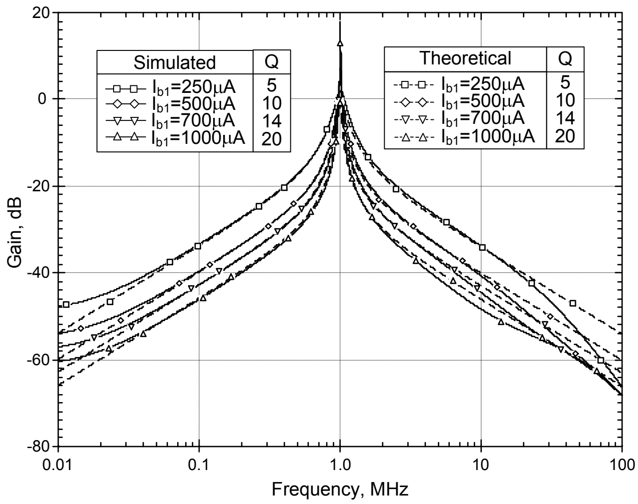


Fig. 11. Simulated frequency responses of the BP filter when I_{b1} is varied.

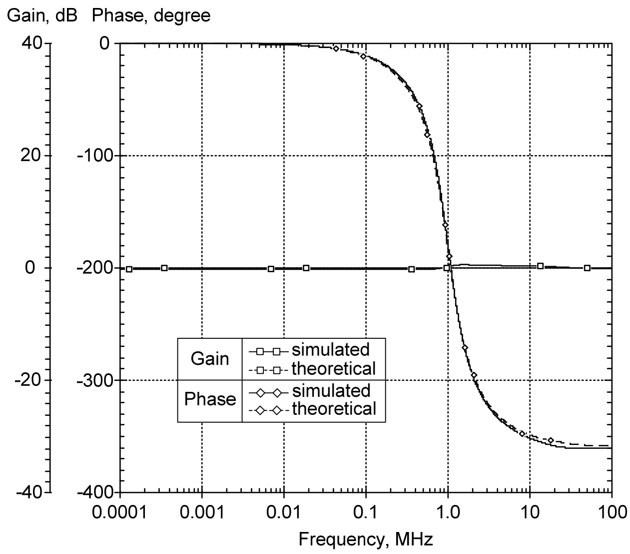


Fig. 9. Simulated AP response of the proposed filter.

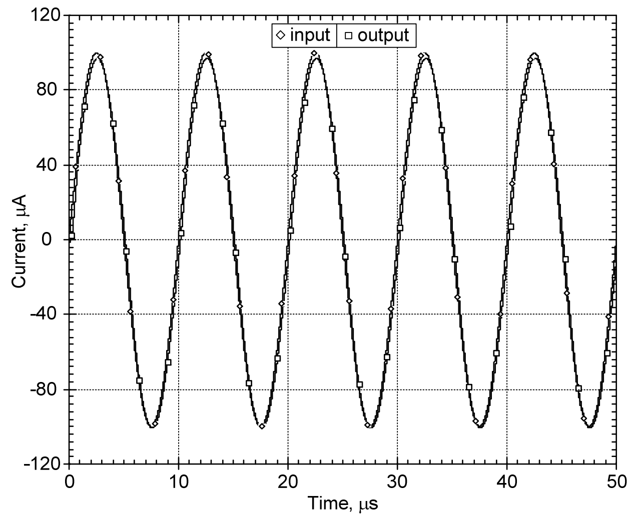


Fig. 12. Time-domain input and output signal waveforms to demonstrate the dynamic range of the proposed filter.

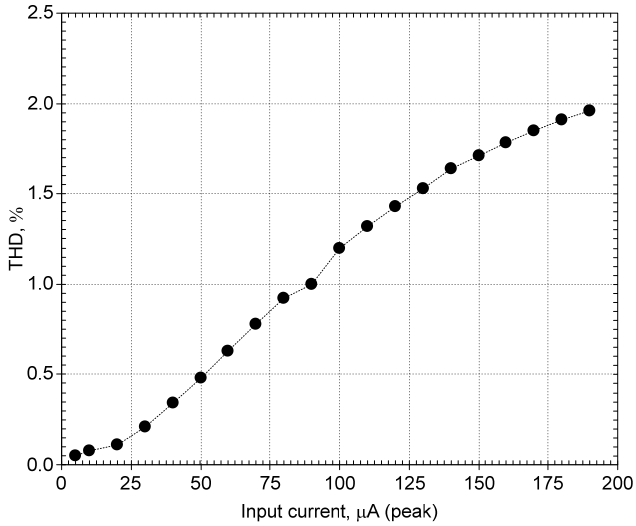


Fig. 13. Dependence of the output harmonic distortion of LP filter on input current amplitude.

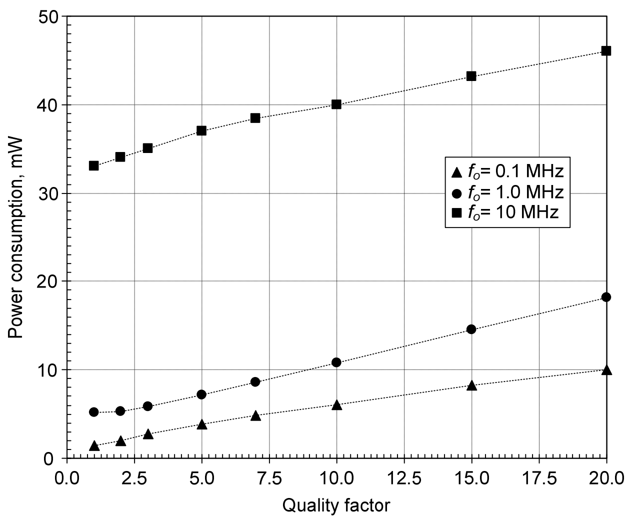


Fig. 14. Simulated power consumption when the quality factor is varied using the bias current.

input signal current level was possible without significant distortion. The result of the I_{LP} total harmonic distortion (THD) analysis was approximately 1.2 %. The harmonic distortion increases rapidly if the input signal was increased beyond 200 µA peak. The dependence of the THD of LP filter on input current amplitude was shown in Fig. 13. The THD was about 1.96 % when the input signal was increased to 190 µA (peak). The power consumption of the proposed filter with different natural frequencies f_o (0.1 MHz, 1 MHz and 10 MHz) when the quality factor was varied using the bias current, was investigated. Fig. 14 shows the power consumption of the circuit with the bias current I_o of 3.1 µA (f_o = 0.1 MHz), 32 µA (f_o = 1 MHz) and 340 µA (f_o = 100 MHz) when the quality factor was varied from 1 to 20 using the bias current I_{b2} from 50 (Q = 1) to 1000 µA (Q = 20) while constant I_{b1} = 50 µA. The THD for different values of Q was also investigated. The simulated result was reported that for Q of 1, 3, 5 and 10, the THDs were 1.28, 2.4, 2.4, and 3.2 % for the input signals of 50, 30, 20 and 10 µA peak, respectively.

5. Conclusions

A new four inputs and five outputs electronically tunable current-mode universal biquad filter employing only one CCCII and one ECCCII and two grounded capacitors was proposed. The use of grounded capacitor and absence from any resistors makes the proposed filter ideal for integrated circuit implementation. In addition, a new single input and five outputs can be achieved by using an additional CCCII to obtain four plus input currents. The proposed circuit possesses the following properties: (1) employment of three translinear current conveyors; (2) ability of realizing all the five standard biquadratic filtering functions, i.e. LP, BP, HP, BS and AP filters with one single topology, (3) separate electronic control of the parameters ω_o and Q, (4) employment of all grounded capacitors, (5) no need to impose critical-matching condition for realizing all the filter responses; (6) provision of low-input impedance, high-output impedance and low active and passive sensitivities and especially (7) offer of electronically tunable high Q-value filter. Simulation results are also given to demonstrate the effectiveness of our schemes. The simulation results obtained were found to be in good agreement with the theory.

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