

# Independently tunable high-input impedance voltage-mode universal biquadratic filter using grounded passive components

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Received 16 March 2014; revised 20 November 2014; accepted 25 June 2015

This paper presents an independently tunable high-input impedance voltage-mode universal biquadratic filter using one plus-type second-generation current conveyor (CCII+), one full differential current conveyor (FDCCII), two grounded capacitors and three grounded resistors which can realize voltage-mode universal filtering responses (lowpass, highpass, bandpass, notch and allpass) from the same topology. The proposed circuit not only permits orthogonal tunability of  $\omega_0$  and  $Q$  but also achieves independent tunability of  $\omega_0$  and  $\omega_0/Q$  by using only grounded resistors without any matching conditions. Moreover, the proposed circuit still offers the following advantages: (i) high-input impedance, (ii) the employment of all grounded passive components, (iii) the employment of only two current conveyors, (iv) no need of component matching conditions for realizing any filter responses, (v) no need of extra inverting or non-inverting amplifiers for special input signals, and (vi) low active and passive sensitivities. H-spice simulations with TSMC 0.18  $\mu\text{m}$  1P6M CMOS process technology validate theoretical predictions.

**Keywords:** Active filters, Second-generation current conveyor, Full differential current conveyor, Voltage-mode, High-input impedance, Universal filter

## 1 Introduction

In the analog circuit design, many novel circuits with the good applications and advantages by using different active elements have been proposed<sup>1-40</sup>. For example, the applications and advantages in the designing high-input impedance voltage-mode active filters have received considerable attention<sup>11-40</sup>. Voltage-mode active filters with high-input impedance are of great interest because it can be easily cascaded to synthesize high-order filters. Therefore, many high-input impedance voltage-mode biquadratic filters have already been studied<sup>11-40</sup>.

The circuits<sup>32-40</sup> cannot realize all five universal filtering functions (lowpass, highpass, bandpass, notch and allpass). The filters consist of more filtering functions, meaning more applications for which they can be used. Although the voltage-mode universal biquad filters using single active element have been proposed<sup>41-43</sup>, these biquad filters have low-input impedance, floating passive elements, and need the component matching conditions for realizing all five filtering responses. Therefore, several high-input impedance voltage-mode universal biquadratic filters which can realize all five universal filtering responses have been presented<sup>11-31</sup>.

The structures<sup>11,13-15,16(main structure),17,18,19(1st structure),20,21</sup> employ three active elements in addition to two<sup>11,13,14(1st structure),15,16(main structure),17-21</sup>/three<sup>14(2nd structure)</sup> capacitors and one<sup>20</sup>/two<sup>13,18</sup>/three<sup>11,14,17/</sup>four<sup>15,16(main structure),19</sup> resistors, respectively. However, these structures<sup>11,13-15,16(main structure),17,18,19(1st structure),20,21</sup> can not offer either of the following two attractive advantages: (i) independent tunability of  $\omega_0$  and  $\omega_0/Q$  without any matching conditions and (ii) the use of only grounded passive elements. The structures<sup>11,13-15,16(main structure),17,18,19(1st structure)</sup> need to use floating resistors which are not suitable for the variations of filter parameters. The structures<sup>13,14,20,21</sup> need to use floating capacitors which are not attractive for monolithic IC implementation.

The circuits<sup>12,16(sub-structure),19(2nd structure),22-31</sup> employ three<sup>16(sub-structure),22,24</sup> DDCCs/three<sup>26</sup> DVCC/two DVCC and one<sup>19(2nd structure)</sup> DDCC/two<sup>27</sup> FDCCII/one FDCCII and one<sup>23,25</sup> DDCC/two<sup>28,31</sup> DDCCTAs/three<sup>12</sup> DDCCTAs/two<sup>30</sup> VDTAs/six<sup>29</sup> OTAs as active elements, respectively, together with two<sup>22-25,27,28,30,31/</sup>three<sup>16(sub-structure),26</sup>/four<sup>19(2nd structure)</sup> grounded resistors and two grounded capacitors<sup>12,16(sub-structure),19(2nd structure),22-31</sup> as passive elements, respectively. These circuits<sup>12,16(sub-structure),19(2nd structure),22-31</sup> have feature (ii).

However, all of the circuits<sup>11-31</sup> do not meet feature (i) and the circuits described in Refs (12-14, 18, 19<sup>1st structure</sup>, 21-25, 26<sup>1st structure</sup>, 27-31) also do not permit orthogonal control of the  $\omega_0$  and  $Q$  since the denominators of the filtering transfer functions of these circuits show that the parameters  $\omega_0$  and  $Q$  are interactive. Hence, the filters<sup>12,28,31</sup> need some matching conditions of bias currents<sup>12,28,31</sup> and resistor<sup>28,31</sup> to obtain the non-interactive filter parameter control. Similarly, the filters<sup>29,30</sup> need matching conditions of bias currents and set the parameter  $Q$  by capacitors to obtain the non-interactive filter parameter control. However, these may not be convenient for the users. The structures<sup>11,15-17,19(2nd structure),20,26(2nd structure)</sup> permit orthogonal control of the  $\omega_0$  and  $Q$ , but only few<sup>19(2nd structure),20,26(2nd structure)</sup> can achieve independent tunability of  $\omega_0$  and  $\omega_0/Q$ . However, it should be noted that all of them<sup>14-17,19(2nd structure),20,26(2nd structure),28,30,31</sup> (except for Ref. (11)) need the component matching condition for realizing allpass filter function and therefore, they lose the advantages of orthogonal and independent tunabilities in the allpass response.

Although the recently reported good high-input impedance structure<sup>11</sup> permits orthogonal control of the  $\omega_0$  and  $Q$  without any matching conditions, the resistor used for orthogonal tunability is not grounded. The use of grounded resistors can be replaced by electronic resistors to obtain electronic tunability<sup>44,45</sup>. Moreover, the further advantage, namely independent tunability of  $\omega_0$  and  $\omega_0/Q$ , cannot be achieved<sup>11</sup>.

In the present paper, the new proposed high-input impedance voltage-mode universal filter biquad using one CCII+ (with simpler implementation configuration than of the CCII-), one FDCCII, two grounded capacitors and three grounded resistors (minimum components for independent and orthogonal tunabilities) achieve the above attractive advantage and still offer many important advantages as presented in Table 1. A comparison between the proposed biquad filter and the previous reported high-input impedance voltage-mode universal biquad filters as following advantages in all five universal filtering functions: (i) the use of only two active components, (ii) independent tunability of the

Table 1 — Comparison of the previous reported high-input impedance voltage-mode universal biquad filters

Previous reported filters	Advantages						
	(i)	(ii)	(iii)	(iv)	(v)	(vi)	(vii)
Ref [11] in 2012	DDCC = 3	No	Yes	No	Yes	Yes	No
Ref [12] in 2013	DDCCTA = 3	No	No	Yes	Yes	Yes	Yes
Ref [13] in 2001	CCII = 3	No	No	No	Yes	No	No
Ref [14] in 2004	CCII = 3	No	No	No	No	No	No
Ref [15] in 2006	DVCC = 3	No	No	No	No	Yes	Yes
Ref [16 <sup>main structure</sup> ] in 2012	DDCC = 3	No	No	No	No	Yes	Yes
Ref [16 <sup>sub-structure</sup> ] in 2012	DDCC = 3	No	Yes	Yes	Yes	Yes	Yes
Ref [17] in 2010	DVCC = 3	No	No	No	No	Yes	No
Ref [18] in 2011	DDCC = 3	No	No	No	Yes	Yes	Yes
Ref [19 <sup>1st structure</sup> ] in 2012	DVCC = 3	No	No	No	No	Yes	Yes
Ref [19 <sup>2nd structure</sup> ] in 2012	DVCC = 2 and DDCC = 1	No	No	Yes	No	Yes	Yes
Ref [20] in 2010	OTA = 2 and CFA = 1	No	No	No	No	Yes	Yes
Ref [21] in 2003	OTA = 2 and CCII = 1	No	No	No	Yes	No	No
Ref [22] in 2008	DDCC = 3	No	No	Yes	Yes	Yes	Yes
Ref [23] in 2010	FDCCII = 1 and DDCC = 1	No	No	Yes	Yes	Yes	Yes
Ref [24] in 2007	DDCC = 3	No	No	Yes	Yes	Yes	Yes
Ref [25] in 2008	FDCCII = 1 and DDCC = 1	No	No	Yes	Yes	Yes	Yes
Ref [26 <sup>1st structure</sup> ] in 2010	DVCC = 3	No	No	Yes	No	Yes	Yes
Ref [26 <sup>2nd structure</sup> ] in 2010	DVCC = 3	No	No	Yes	No	Yes	Yes
Ref [27 <sup>1st structure</sup> ] in 2008	FDCCII = 2	No	No	Yes	Yes	Yes	Yes
Ref [27 <sup>2nd structure</sup> ] in 2008	FDCCII = 2	No	No	Yes	Yes	Yes	Yes
Ref [28] in 2012	DDCCTA = 2	No	No	Yes	No	Yes	Yes
Ref [29] in 2010	OTA = 6	No	No	Yes	Yes	Yes	Yes
Ref [30] in 2013	VDTA = 2	No	No	Yes	No	Yes	Yes
Ref [31] in 2013	DDCCTA = 2	No	No	Yes	No	Yes	Yes
Proposed circuit	FDCCII = 1 and CCII = 1	Yes	Yes	Yes	Yes	Yes	Yes

parameters  $\omega_0$  and  $\omega_0/Q$  without any matching conditions, (iii) orthogonal tunability of  $\omega_0$  and  $Q$  without any matching conditions, (iv) the employment of all grounded passive components, (v) no need to impose component choice, (vi) no need of inverting or non-inverting amplifiers, and (vii) low active and passive sensitivity performances (Table 1). As can be seen, none of the previous reported high-input impedance voltage-mode universal biquad filters is capable of achieving all five universal filtering functions with the advantage (ii): independent tunability of the parameters  $\omega_0$  and  $\omega_0/Q$  without any matching conditions.

## 2 Proposed Circuit

The proposed high-input impedance voltage-mode universal biquadratic filter is shown in Fig. 1 using only one FDCCII, one CCII+, two grounded capacitors and three grounded resistors. It is noticed that all capacitors and all resistors are grounded and, as such, are attractive for integrated circuit implementation and suitable for the adjustment of filter parameters, respectively. Moreover, since the input signals ( $V_{in1}$ ,  $V_{in2}$  and  $V_{in3}$ ) are connected to the high-input impedance input nodes of the FDCCII (i.e. the  $Y_2$ ,  $Y_3$  and  $Y_4$  of the FDCCII), the circuit enjoys the advantage of high input impedance.

By using standard notation, the port relations of a CCII+ can be characterized by  $I_Y = 0$ ,  $V_X = V_Y$  and

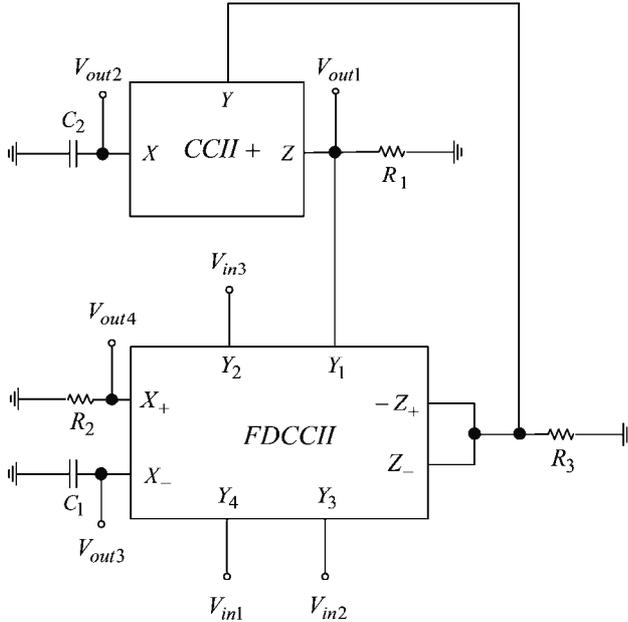


Fig. 1 — Proposed independently tunable high-input impedance voltage-mode universal filter structure

$I_Z = +I_X$ , the port relations<sup>46</sup> of a FDCCII can be characterized by  $I_{Y1} = I_{Y2} = I_{Y3} = I_{Y4} = 0$ ,  $V_{X+} = V_{Y1} - V_{Y2} + V_{Y3}$ ,  $V_{X-} = -V_{Y1} + V_{Y2} + V_{Y4}$ ,  $I_{-Z+} = -I_{X+}$  and  $I_{-Z-} = I_{X-}$ . Derived by each nodal equation of the proposed circuit that the input–output relationship matrix form of Fig. 1 can be expressed as:

$$\begin{bmatrix} G_1 & -sC_2 & 0 & 0 \\ 0 & -G_3 & sC_1 & -G_2 \\ 1 & 0 & 1 & 0 \\ 1 & 0 & 0 & -1 \end{bmatrix} \begin{bmatrix} V_{out1} \\ V_{out2} \\ V_{out3} \\ V_{out4} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ V_{in1} + V_{in3} \\ V_{in3} - V_{in2} \end{bmatrix} \quad \dots (1)$$

From Fig. 1 and the above matrix form, it should be noted that the  $-Z+$  and  $Z-$  are connected to single node which is a useful design method. The useful design can be achieved the second row of Eq. (1) which is a mathematical base of independent tunability and using grounded passive components.

From the above matrix form, the following four output voltages can be obtained as:

$$V_{out1} = \frac{s^2 C_1 C_2 (V_{in1} + V_{in3}) - s C_2 G_2 (V_{in2} - V_{in3})}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad \dots (2)$$

$$V_{out2} = \frac{s C_1 G_1 (V_{in1} + V_{in3}) - G_1 G_2 (V_{in2} - V_{in3})}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad \dots (3)$$

$$V_{out3} = \frac{s C_2 G_2 (V_{in1} + V_{in2}) + G_1 G_3 (V_{in1} + V_{in3})}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad \dots (4)$$

$$V_{out4} = \frac{s^2 C_1 C_2 (V_{in1} + V_{in2}) + G_1 G_3 (V_{in2} - V_{in3})}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad \dots (5)$$

From Eqs (2)-(5), the voltage-mode five standard filtering transfer functions (lowpass, highpass, bandpass, notch and allpass) can be obtained according to input and output conditions as follows:

**Case I:** If  $V_{in2} = V_{in3} = 0$  (grounded) and  $V_{in1} =$  input voltage signal ( $V_{in}$ ), then three output voltage signals are obtained as shown in Eqs. (6-8):

$$\frac{V_{out1}}{V_{in}} = \frac{s^2 C_1 C_2}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{highpass}) \quad \dots (6)$$

$$\frac{V_{out2}}{V_{in}} = \frac{s C_1 G_1}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{bandpass}) \quad \dots (7)$$

$$\frac{V_{out4}}{V_{in}} = \frac{s^2 C_1 C_2}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{highpass}) \quad \dots (8)$$

**Case II:** If  $V_{in1} = V_{in3} = 0$  (grounded) and  $V_{in2} =$  input voltage signal ( $V_{in}$ ), then four output voltage signals are obtained as below: [shown in Eqs (9)-(12)]

$$\frac{V_{out1}}{V_{in}} = \frac{-s C_2 G_2}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{inverting bandpass}) \quad \dots (9)$$

$$\frac{V_{out2}}{V_{in}} = \frac{-G_1 G_2}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{inverting lowpass}) \quad \dots (10)$$

$$\frac{V_{out3}}{V_{in}} = \frac{s C_2 G_2}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{bandpass}) \quad \dots (11)$$

$$\frac{V_{out4}}{V_{in}} = \frac{s^2 C_1 C_2 + G_1 G_3}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{notch}) \quad \dots (12)$$

Examining Eqs (11) and (12) show that the voltage difference between  $V_{out4}$  and  $V_{out3}$  yields an allpass filter response.

$$\frac{V_{out4} - V_{out3}}{V_{in}} = \frac{s^2 C_1 C_2 - s C_2 G_2 + G_1 G_3}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{allpass}) \quad \dots (13)$$

To affect the subtraction, the output of the allpass filter can be directly obtained between the output terminal of  $V_{out4}$  and that of  $V_{out3}$  without need of extra inverting or non-inverting amplifiers. Moreover, since the structure does not need to use inverting-type input signals or double-type input signals for the use of special input signals, the structure has no need of extra inverting or non-inverting amplifiers for realizing any filter transfer functions.

**Case III:** If  $V_{in1} = V_{in2} = 0$  (grounded) and  $V_{in3} =$  input voltage signal ( $V_{in}$ ), then two output voltage signals are obtained as shown in Eqs (14-15).

$$\frac{V_{out3}}{V_{in}} = \frac{G_1 G_3}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{lowpass}) \quad \dots (14)$$

$$\frac{V_{out4}}{V_{in}} = \frac{-G_1 G_3}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{inverting lowpass}) \quad \dots (15)$$

Note that by making the resistor  $R_1$  floating that does not affect tunable functions, and then connecting  $V_{in1}$ ,  $V_{in2}$  with the floating terminal of the  $R_1$  as a new voltage input signal  $V_{in}$ , the proposed circuit can also realize a single-terminal-output allpass response from the node of  $V_{out1}$  as below:

$$\frac{V_{out1}}{V_{in}} = \frac{s^2 C_1 C_2 - s C_2 G_2 + G_1 G_3}{s^2 C_1 C_2 + s C_2 G_2 + G_1 G_3} \quad (\text{allpass}) \quad \dots (16)$$

In all cases, there are no critical component-matching conditions in the design. Moreover, the structure does not need extra inverting or non-inverting amplifier for realizing any filter transfer functions. Inspection of Eqs (2-16) shows that, in all cases, the parameters  $\omega_0$  and  $\omega_0/Q$ , and  $Q$  are given by:

$$\omega_0 = \sqrt{\frac{G_1 G_3}{C_1 C_2}} \quad \dots (17)$$

$$\frac{\omega_0}{Q} = \frac{G_2}{C_1} \quad \dots (18)$$

$$Q = \frac{1}{G_2} \sqrt{\frac{C_1 G_1 G_3}{C_2}} \quad \dots (19)$$

In all cases, the parameters  $\omega_0$  and  $Q$  can be orthogonally adjustable by tuning the grounded resistor  $R_3$  for  $\omega_0$  first and then grounded resistor  $R_2$  for  $Q$  without disturbing parameter  $\omega_0$ . Moreover, the parameters  $\omega_0$  can be independently controlled by adjusting the grounded resistor  $R_3$  without disturbing the parameter  $\omega_0/Q$ , and the parameter  $\omega_0/Q$  can be independently controlled by adjusting the grounded resistor  $R_2$  without disturbing the parameter  $\omega_0$ . The use of grounded resistors can be replaced by electronic resistors to obtain electronic tunability<sup>44,45</sup>. It is important to note that the different values of the  $Q$  and  $\omega_0/Q$  can be easily obtained by only varying one grounded resistor  $R_2$  (no need of two resistors matching conditions) without disturbing parameter  $\omega_0$ . Moreover, the different values of the  $\omega_0$  also can be easily obtained by only varying one grounded resistor  $R_3$  (no need of two resistors matching conditions) without disturbing parameter  $\omega_0/Q$ . Therefore, the proposed circuit enjoys important advantages: independent tunability of the parameters  $\omega_0$  and  $\omega_0/Q$ , and orthogonal control of the parameters

$\omega_0$  and  $Q$  by using only grounded resistors without any matching conditions.

### 3 Non-ideal Analysis

Taking the tracking errors of the FDCCII into account, the relationship of the terminal voltages and currents can be written as:  $I_{Y1} = I_{Y2} = I_{Y3} = I_{Y4} = 0$ ,  $V_{X+} = \beta_{11}(s)V_{Y1} - \beta_{12}(s)V_{Y2} + \beta_{13}(s)V_{Y3}$ ,  $V_{X-} = -\beta_{21}(s)V_{Y1} + \beta_{22}(s)V_{Y2} + \beta_{24}(s)V_{Y4}$ ,  $I_{-Z+} = -\alpha_1(s)I_{X+}$ ,  $I_{Z-} = \alpha_2(s)I_{X-}$ , and taking the tracking errors of the CCII+ into account, the relationship of the terminal voltages and currents can be written as:  $I_Y = 0$ ,  $V_X = \beta_{33}(s)V_Y$ ,  $I_Z = \alpha_3(s)I_X$ , where  $\alpha_m(s)$  and  $\beta_{mn}(s)$  represent the frequency transfers of the internal current and voltage followers of the FDCCII and CCII+. They can be approximated by the first order lowpass functions<sup>46-48</sup>. For frequencies much less than the corner frequencies of the FDCCII and CCII+, all  $\alpha_m(s)$  and  $\beta_{mn}(s)$  are real quantities of magnitudes slightly less than one<sup>46-48</sup>. Assuming the circuit works at frequencies much less than the corner frequencies of  $\alpha_m(s)$  and  $\beta_{mn}(s)$ , namely,  $\alpha_m(s) = \alpha = 1 - \varepsilon_i$  and  $\varepsilon_i$  ( $\varepsilon_i \ll 1$ ) denotes the current tracking error of the FDCCII and CCII+, and  $\beta_{mn}(s) = \beta = 1 - \varepsilon_v$  and  $\varepsilon_v$  ( $\varepsilon_v \ll 1$ ) denotes the voltage tracking error of the FDCCII and CCII+. Incorporating the two sources of tracking errors into the proposed circuit in Fig. 1 that the denominator of transfer functions becomes:

$$D(s) = s^2 C_1 C_2 \alpha_1 \alpha_3 \beta_{12} \beta_{33} + s C_2 G_2 \alpha_2 \alpha_3 \beta_{22} \beta_{33} + G_1 G_3 \quad \dots (20)$$

The non-ideal  $\omega_0$ ,  $\omega_0/Q$ , and  $Q$  are given by:

$$\omega_0 = \sqrt{\frac{G_1 G_3}{C_1 C_2 \alpha_1 \alpha_3 \beta_{12} \beta_{33}}} \quad \dots (21)$$

$$\frac{\omega_0}{Q} = \frac{G_2 \alpha_2 \beta_{22}}{C_1 \alpha_1 \beta_{12}} \quad \dots (22)$$

$$Q = \frac{1}{G_2 \alpha_2 \beta_{22}} \sqrt{\frac{C_1 G_1 G_3 \alpha_1 \beta_{12}}{C_2 \alpha_3 \beta_{33}}} \quad \dots (23)$$

fb

This shows that, considering non-ideal  $\omega_0$ ,  $\omega_0/Q$ , and  $Q$ , the proposed biquad still enjoys important advantages: independent tunability of the parameters  $\omega_0$  and  $\omega_0/Q$  and orthogonal control of the parameters  $\omega_0$  and  $Q$  by using only grounded resistors without any matching conditions. The active and passive sensitivities of  $\omega_0$ ,  $\omega_0/Q$ , and  $Q$  are shown as follows:

$$\begin{aligned} S_{\alpha_2}^{a_0} &= S_{\beta_{11}}^{a_0} = S_{\beta_{13}}^{a_0} = S_{\beta_{21}}^{a_0} = S_{\beta_{22}}^{a_0} = S_{\beta_{24}}^{a_0} = S_{G_2}^{a_0} = 0 \\ S_{\alpha_1}^{a_0} &= S_{\alpha_3}^{a_0} = S_{\beta_{12}}^{a_0} = S_{\beta_{33}}^{a_0} = S_{C_1}^{a_0} = S_{C_2}^{a_0} = -S_{G_1}^{a_0} \\ &= -S_{G_3}^{a_0} = -\frac{1}{2} \\ S_{\alpha_3}^{(a_0/Q)} &= S_{\beta_{11}}^{(a_0/Q)} = S_{\beta_{13}}^{(a_0/Q)} = S_{\beta_{21}}^{(a_0/Q)} = S_{\beta_{24}}^{(a_0/Q)} = S_{\beta_{33}}^{(a_0/Q)} \\ &= S_{G_1}^{(a_0/Q)} = S_{G_3}^{(a_0/Q)} = S_{C_2}^{(a_0/Q)} = 0 \\ S_{\alpha_2}^{(a_0/Q)} &= S_{\beta_{22}}^{(a_0/Q)} = -S_{\beta_{12}}^{(a_0/Q)} = -S_{\alpha_1}^{(a_0/Q)} = S_{G_2}^{(a_0/Q)} \\ &= -S_{C_1}^{(a_0/Q)} = 1 \\ S_{\beta_{11}}^Q &= S_{\beta_{13}}^Q = S_{\beta_{21}}^Q = S_{\beta_{23}}^Q = 0, \quad S_{\alpha_2}^Q = S_{\beta_{22}}^Q = S_{G_2}^Q = -1 \\ S_{\alpha_1}^Q &= -S_{\alpha_3}^Q = S_{\beta_{12}}^Q = -S_{\beta_{33}}^Q = S_{C_1}^Q = -S_{C_2}^Q \\ &= -S_{G_1}^Q = S_{G_3}^Q = \frac{1}{2} \quad \dots (24) \end{aligned}$$

(The above line is:  $= S_{G_1}^Q = S_{G_3}^Q = \frac{1}{2}$ )

The proposed biquad filter parameter sensitivities are low and not larger than unity in absolute value.

### 4 H-spice Simulations

The CMOS implementations of the simplified<sup>47</sup> FDCCII and CCII+ (Ref.49) are shown in Figs 2 and 3, respectively, with the NMOS transistor aspect ratios ( $W/L=5 \mu\text{m}/0.8 \mu\text{m}$ ) and PMOS transistor aspect ratios<sup>9,47,50</sup> ( $W/L=10 \mu\text{m}/0.8 \mu\text{m}$ ). To verify the theoretical analysis of the proposed voltage-mode universal filter as shown in Fig. 1, the H-SPICE simulations, using the TSMC (Taiwan Semiconductor Manufacturing Company, Ltd.) 0.18 $\mu\text{m}$  1P6M CMOS process technology with the parameters<sup>9,50</sup> of level 49, have been performed. All simulated filter component values are  $C_1 = C_2 = 10 \text{ pF}$ , and the other component values are: (i)  $R_1 = R_2 = R_3 = 10 \text{ k}\Omega$  for bandpass, lowpass, highpass, notch, and allpass responses with  $Q = 1$ , (ii)  $R_2 = 40 \text{ k}\Omega$ ,  $R_1 = R_3 = 10 \text{ k}\Omega$  for bandpass, lowpass, highpass, notch, and allpass responses with  $Q = 4$ , (iii)  $R_2 = 80 \text{ k}\Omega$ ,  $R_1 = R_3 = 10 \text{ k}\Omega$  for bandpass, lowpass, highpass, notch, and allpass responses with  $Q = 8$ . The theoretical resonant frequency of all simulated filters is located at  $f_0 = 1.5915 \text{ MHz}$ . In Fig. 2, the supply voltages are  $V_{DD} = -V_{SS} = 0.9 \text{ V}$  and the biasing currents are  $I_b = I_{sb} = 28 \mu\text{A}$ . In Fig. 3, the supply voltages are  $V_{DD} = -V_{SS} = 0.9 \text{ V}$  and the biasing voltages are  $V_{b1} = 0.17 \text{ V}$  and  $V_{b2} = -0.38 \text{ V}$ . Figures 4-9 show that the operative frequencies of proposed universal filters are from 10 kHz to

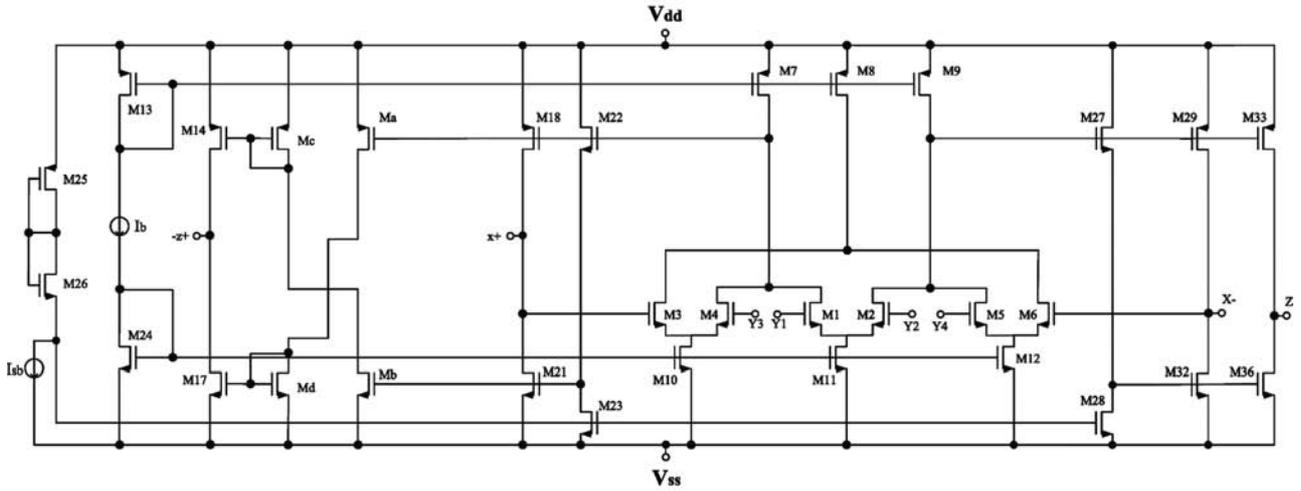


Fig. 2 — CMOS implementation of the simplified FDCCII

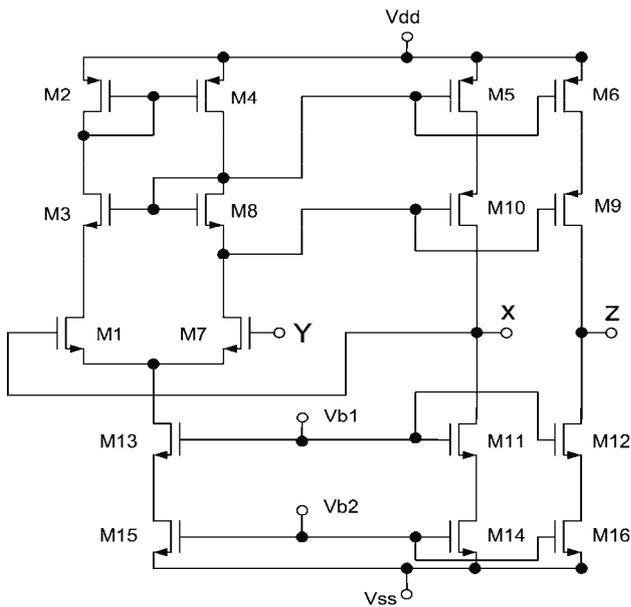


Fig. 3 — CMOS implementation of the CCII+

100 MHz. Moreover, Figs 4-9 also include the tuning possibilities ( $Q = 1, 4,$  and  $8$ ) of all five filtering responses (lowpass, highpass, bandpass, notch, and allpass) which have been verified with theoretical results. Figures 4-6 show the simulated lowpass, bandpass, highpass, and notch amplitude frequency responses with  $Q = 1, Q = 4,$  and  $Q = 8$ . Figures 7-9 show the simulated allpass phase and amplitude-frequency responses with  $Q = 1, Q = 4,$  and  $Q = 8$ . The simulated resonant frequencies and percentage errors are summarized in Table 2. As can be seen, there is a close agreement between theory and

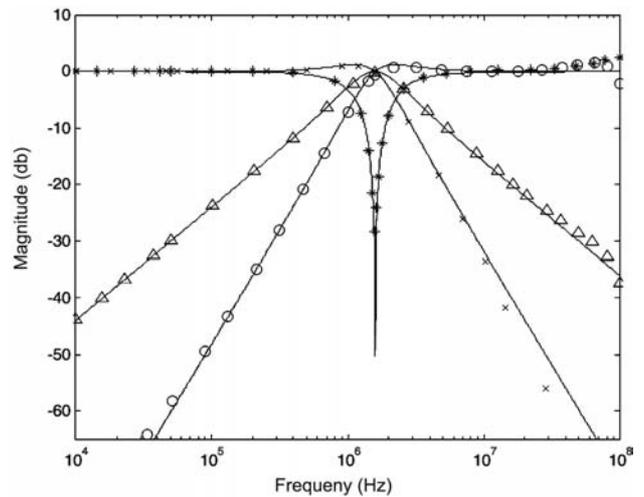


Fig. 4—Amplitude-frequency responses of the proposed highpass, lowpass, bandpass, and notch signals with  $f_0 = 1.5915$  MHz and  $Q = 1$  ( $\circ$ , simulated highpass response;  $\times$ , simulated lowpass response;  $\Delta$  simulated bandpass response;  $*$ , simulated notch response; and —, theoretical curve)

simulation. By varying only grounded resistor  $R_2$ , the  $Q$  (and  $\omega_0/Q$ ) can be controlled without disturbing parameter  $\omega_0$ . From Figs 4-9 and Table 2, the proposed circuit can realize all five filtering responses with the different  $Q$  (and  $\omega_0/Q$ ) without disturbing parameter  $\omega_0$ . Figures 10-12 show the  $dc$  characteristic of simulated lowpass, bandpass, highpass, notch and allpass signals with  $Q = 1, Q = 4,$  and  $Q = 8$ . The theoretical  $dc$  gains of simulated lowpass, bandpass, highpass, notch, and allpass signals are  $-G_2/G_3, 0, 0, 1,$  and  $1$ , respectively. The  $dc$  input and output voltages of all five simulated filtering signals

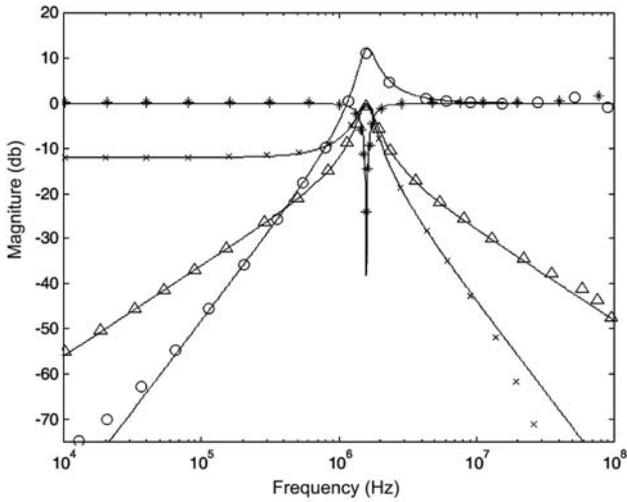


Fig. 5 — Amplitude-frequency responses of the proposed highpass, lowpass, bandpass, and notch signals with  $f_0 = 1.5915$  MHz and  $Q = 4$  (○, simulated highpass response; ×, simulated lowpass response; △ simulated bandpass response; \*, simulated notch response; and —, theoretical curve)

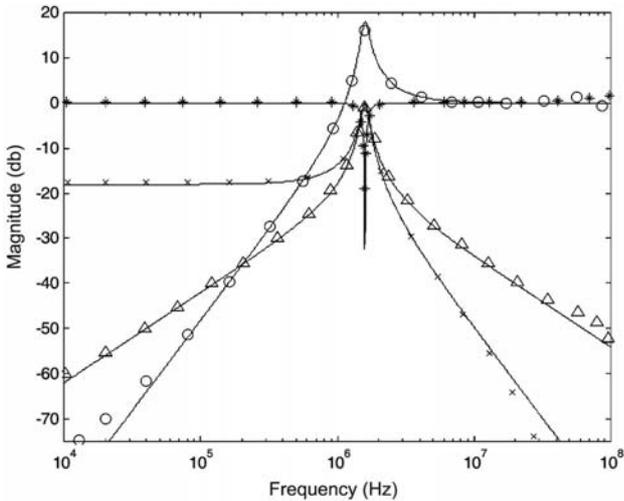


Fig. 6 — Amplitude-frequency responses of the proposed highpass, lowpass, bandpass, and notch signals with  $f_0 = 1.5915$  MHz and  $Q = 8$  (○, simulated highpass response; ×, simulated lowpass response; △ simulated bandpass response; \*, simulated notch response; and —, theoretical curve)

are shown in horizontal coordinate and vertical coordinate of Figs. 10-12, respectively, and therefore, the simulated  $dc$  gains can be obtained from the slope of the straight line shown in Figs 10-12. The simulation results agree well with the theoretical values. By setting input voltages to zero, it also can be observed from Figs 10-12 that the bandpass, highpass, and allpass responses almost have no the  $dc$  offset

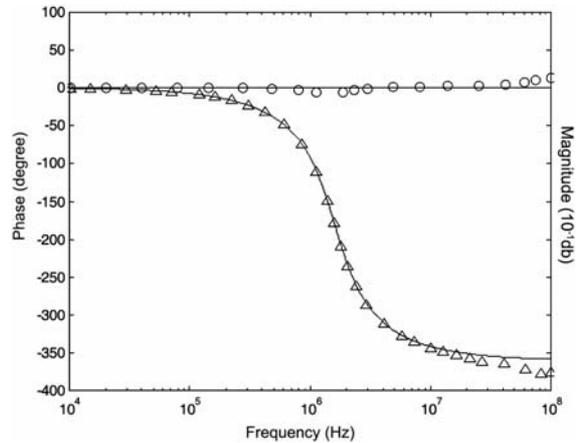


Fig. 7 — Phase-frequency and amplitude-frequency responses of the proposed allpass signal with  $f_0 = 1.5915$  MHz and  $Q = 1$  (△ simulated phase; ○, simulated amplitude; and —, theoretical curve)

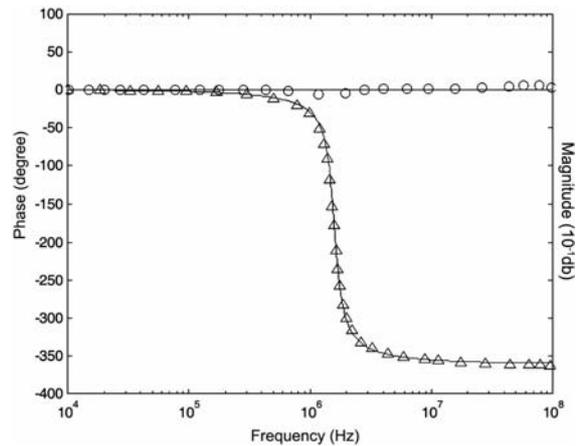


Fig. 8 — Phase-frequency and amplitude-frequency responses of the proposed allpass signal with  $f_0 = 1.5915$  MHz and  $Q = 4$  (△ simulated phase; ○, simulated amplitude; and —, theoretical curve)

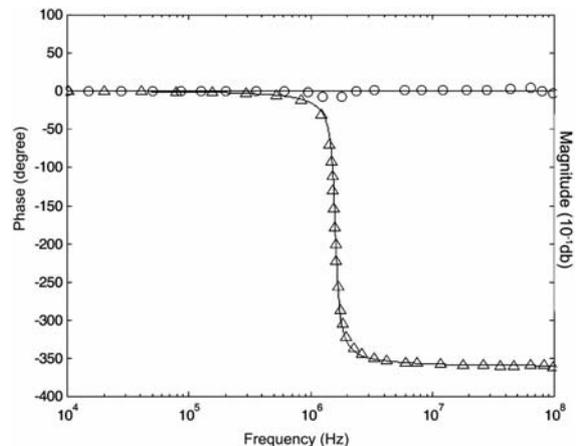


Fig. 9 — Phase-frequency and amplitude-frequency responses of the proposed allpass signal with  $f_0 = 1.5915$  MHz and  $Q = 8$  (△ simulated phase; ○, simulated amplitude; and —, theoretical curve)

Table 2 — Simulated resonant frequencies  $f_0$ , its percentage errors, and  $dc$  offset voltage of the proposed biquad filter with theoretical  $f_0 = 1.5915$  MHz

Simulated $f_0$ (MHz) Percentage errors $dc$ offset voltage (mV)	Filter type				
	Lowpass	Highpass	Bandpass	Notch	Allpass
Simulated $f_0$ (MHz) ( $Q = 1$ )	1.5917	1.5722	1.5958	1.5884	1.5812
Percentage error ( $Q = 1$ )	0.0126%	1.213%	0.270%	0.195%	0.647%
$dc$ offset voltage (mV) ( $Q = 1$ )	7.8770	0.0005322	0.0005322	-10.6946	0.0005322
Simulated $f_0$ (MHz) ( $Q = 4$ )	1.5894	1.5815	1.5885	1.5885	1.5804
Percentage error ( $Q = 4$ )	0.132%	0.628%	0.189%	0.189%	0.697%
$dc$ offset voltage (mV) ( $Q = 4$ )	0.1509	0	0	-10.7435	0
Simulated $f_0$ (MHz) ( $Q = 8$ )	1.5892	1.5834	1.5879	1.5895	1.5794
Percentage error ( $Q = 8$ )	0.145%	0.509%	0.226%	0.126%	0.760%
$dc$ offset voltage (mV) ( $Q = 8$ )	-1.1439	0	0	-10.7517	0

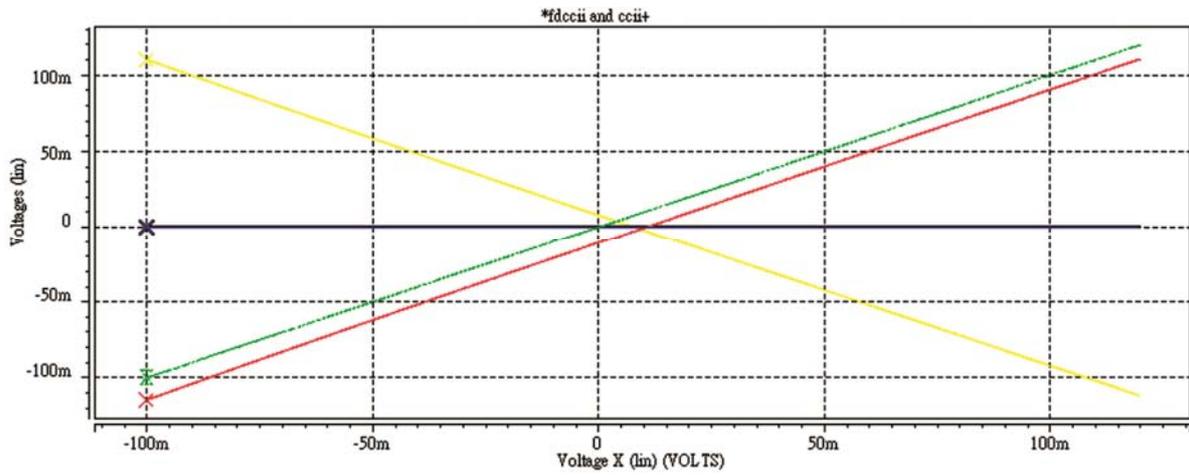


Fig. 10 —  $dc$  characteristic of proposed lowpass, bandpass, highpass, notch, and allpass signals with  $Q = 1$  (— (blue), simulated highpass  $dc$  characteristic; — (yellow), simulated lowpass  $dc$  characteristic; — (black), simulated bandpass  $dc$  characteristic; — (red), simulated notch  $dc$  characteristic; and — (green), simulated allpass  $dc$  characteristic)

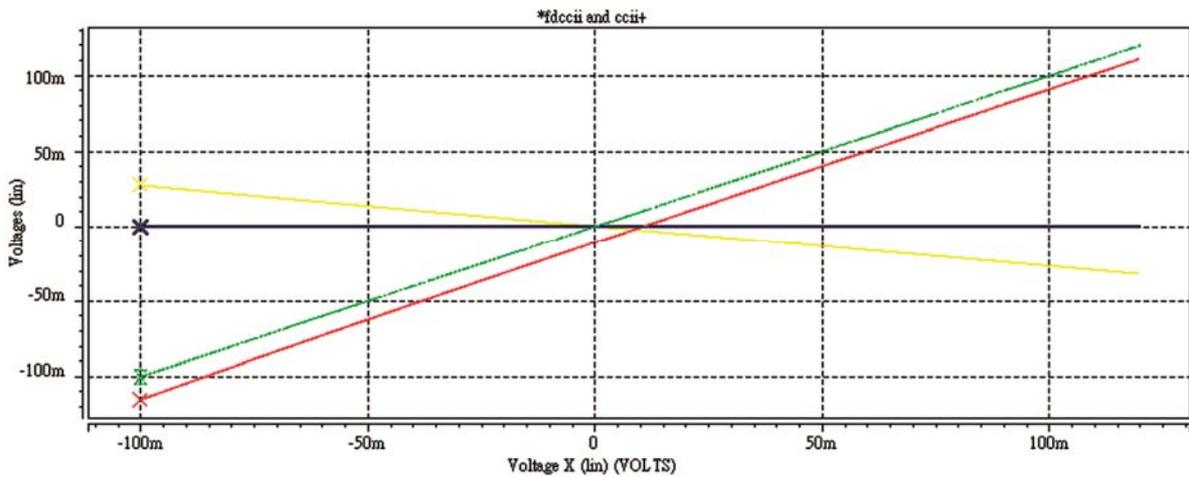


Fig. 11 —  $dc$  characteristic of proposed lowpass, bandpass, highpass, notch, and allpass signals with  $Q = 4$  (— (blue), simulated highpass  $dc$  characteristic; — (yellow), simulated lowpass  $dc$  characteristic; — (black), simulated bandpass  $dc$  characteristic; — (red), simulated notch  $dc$  characteristic; and — (green), simulated allpass  $dc$  characteristic)

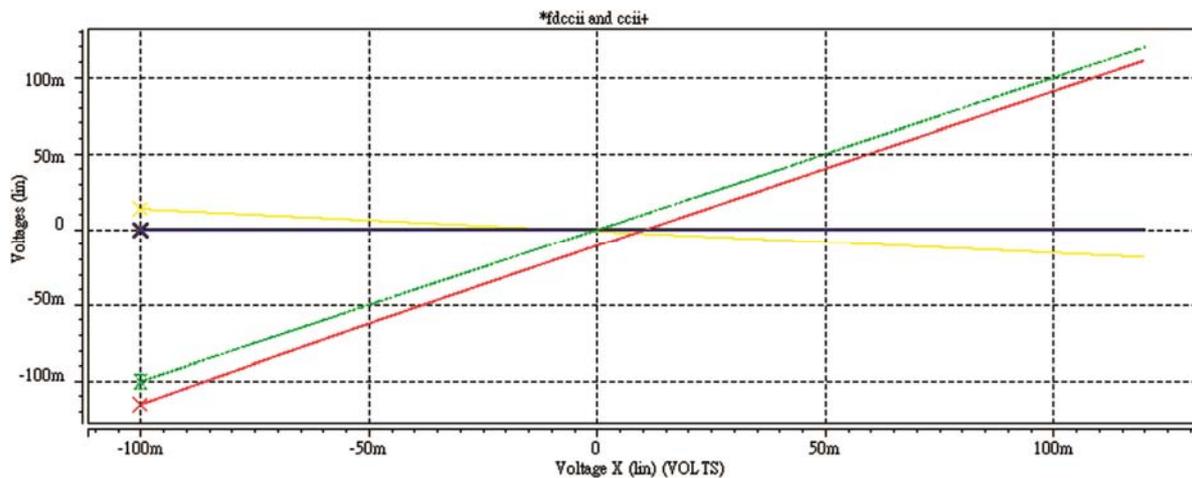


Fig. 12— $dc$  characteristic of proposed lowpass, bandpass, highpass, notch, and allpass signals with  $Q = 8$  (— (blue), simulated highpass  $dc$  characteristic; — (yellow), simulated lowpass  $dc$  characteristic; — (black), simulated bandpass  $dc$  characteristic; — (red), simulated notch  $dc$  characteristic; and — (green), simulated allpass  $dc$  characteristic)

voltages, and notch response has about the  $dc$  offset voltage of  $-10$  mV. Table 2 also presents the  $dc$  offset voltages of all five universal filters with  $Q = 1$ ,  $Q = 4$ , and  $Q = 8$ .

## 5 Conclusions

Although many high-input impedance voltage-mode biquadratic filters have been proposed<sup>11-40</sup>, none can offer all five standard filter functions with the very attractive advantage, namely independent tunability of  $\omega_0$  and  $\omega_0/Q$  without any matching conditions. In the present paper, a new high-input impedance voltage-mode universal biquadratic filter using only one CCII+ (with simpler implementation configuration than of the CCII-), one FDCCII, two grounded capacitors and three grounded resistors is presented which can realize all five standard filter functions (lowpass, highpass, bandpass, notch, and allpass) with the following three attractive advantages: (i) independent tunability of the parameters  $\omega_0$  and  $\omega_0/Q$  without any matching conditions, (ii) orthogonal tunability of the parameters  $\omega_0$  and  $Q$  without any matching conditions, and (iii) using only three grounded resistors (minimum components for independent and orthogonal tunabilities) suitable for the variations of filter parameters. Moreover, the proposed high-input impedance voltage-mode biquad filter still offers the following advantages: (i) the employment of only two grounded capacitors, (ii) the employment of only two current conveyors, (iii) no need of component matching conditions for realizing any filter responses,

(iv) no need of extra inverting or non-inverting amplifiers for special input signals, (v) simultaneous realization of many different filter functions, (vi) no need to change the filter topology, and (vii) low active and passive sensitivities. H-Spice simulations, using TSMC 0.18  $\mu\text{m}$  1P6M CMOS process technology and supply voltages  $\pm 0.9$  V, confirm the theoretical predictions.

## Acknowledgement

The author thanks the Ministry of Science and Technology (formerly National Science Council) of Taiwan, R. O. C. The Ministry of Science and Technology, Taiwan, R O C supported this work under grant number NSC 102-2221-E-149-004-.

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