A single-input multiple-output voltage-mode second-order universal filter using only grounded passive components

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A new single-input multi-output voltage-mode universal filter structure with high input impedance yielding easy cascadability is proposed in this paper. The proposed filter can provide all the standard second-order universal filter responses such as low-pass, band-pass, high-pass, notch filter (NF) and all-pass (AP) responses. NF and AP responses possess low output impedances but they can be obtained through a switch. Each time, one of the NF and AP responses can be obtained. Also, other responses which do not have low output impedances can be obtained simultaneously. The proposed filter uses three plus-type differential difference current conveyors and a minimum number of only grounded passive elements, and does not need any critical passive component matching conditions and cancellation constraints; thus, it is suitable for integrated circuit process. By adding extra one DDCC+ and two grounded resistors, a universal filter with a gain is obtained. Several computer simulations using SPICE program are included to verify the theory.

Keywords: Voltage-mode, SIMO, Universal filter, DDCC+

As a current-mode (CM) active device, a differential difference current conveyor (DDCC) enjoys the advantages of both a second-generation current conveyor and a differential difference amplifier1. Therefore, a number of DDCC based voltage-mode (VM) second-order filters2–25 have been reported in related open literature. The use of CM active devices for instance DDCCs has some potential superiority such as a smaller number of components, larger dynamic range, better linearity, wider bandwidth, etc. when compared to that of VM counterparts for example (operational amplifiers) OAs26. However, DDCC based VM second-order filters2–25 have the following drawbacks:

(i) Use of floating passive component(s)3,4,6,9,11-15,17,18,20,22.
(ii) Do not provide all the universal filter responses1,3,9,14,16,18,19,23-25.
(iii) Require critical passive component matching condition(s)4,7,10,12-15,18.
(iv) Each time, provide only one response3,5,10,13,18,21,24.
(v) Do not use plus-type DDCC (DDCC+) with single Z terminal7,9-17,20,22-24.
(vi) Do not have the property of high input impedance4,7,9,11,12,14,17,20.

(vii) Have transfer functions (TFs) with complex combination of input signals2,5,6,8,10,11,12,13,15,17,20-22.
(viii) Have a capacitor connected in series to X terminal of the DDCC9; accordingly, they cannot be operated properly at high frequencies27.
(ix) Have a different active device such as fully differential current conveyor (FDCCII)22.
(x) Consist of operational transconductance amplifiers (OTAs)25; thus, they have limitations at high frequencies28.

In this paper, a new single-input multi-output (SIMO) VM universal filter topology with high input impedance yielding easy cascadability with other VM circuits is proposed. The proposed filter can provide all the standard second-order universal filter responses such as low-pass (LP), band-pass (BP), high-pass (HP), notch filter (NF) and all-pass (AP) responses from the same structure. NF and AP responses have low output impedances whereas they can be obtained via a switch. Each time, one of the NF and AP responses is available. Further, other responses can be obtained simultaneously. The proposed filter configuration employs three standard DDCC+ and a canonical number of only grounded passive components, and does not suffer from any critical passive component matching conditions; accordingly, it is suitable for integrated circuit (IC) fabrication29-31. However, LP, BP and HP responses do not provide low output
impedances; thus, extra buffers are needed if next stages do not have high input impedances. Electronically tunable grounded resistors \(32-34\) can be replaced instead of both resistors of the proposed universal filter to control it externally. By adding extra one DDCC+ and two grounded resistors, a universal filter with a gain is obtained. Some computer simulations based on SPICE program are achieved to exhibit performance, effectiveness and workability of the proposed filter.

**SIMO Universal Filter Topology**

Defining matrix equation of the DDCC+ whose electrical symbol is depicted in Fig. 1 can be given by

\[
\begin{bmatrix}
I_{z1} \\
I_{y1} \\
I_{y2} \\
I_{y3} \\
V_x \\
V_y
\end{bmatrix} =
\begin{bmatrix}
\alpha_k & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
0 & \beta_{y1} & -\beta_{y2} & \beta_{y3} \\
0 & 0 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
I_z \\
V_y1 \\
V_y2 \\
V_y3 \\
V_x1 \\
V_x2
\end{bmatrix}
\]

... (1)

In matrix Eq. (1), at sufficiently low frequencies, frequency dependent non-ideal current gain \(\alpha_k=1+\epsilon_k\) \((k=1, 2, 3\) represents the \(k^{th}\) DDCC+) and frequency dependent non-ideal voltage gains \(\beta_{yj}=1+\epsilon_{yj}\) \((j=1, 2, 3\) are ideally equal to unity. In addition, \(\epsilon_k\) and \(\epsilon_{yj}\), ideally equal to zero, are respectively called as current and voltage tracking errors where \(|\epsilon_k|<<1\) and \(|\epsilon_{yj}|<<1\). The proposed SIMO VM second-order universal filter with high input impedance is given in Fig. 2. The proposed filter has two cases, the first case is a switch connected to NF and the second one is the switch connected to AP.

If the switch is connected to NF, the following TFs are simultaneously obtained:

\[
\frac{V_{o1}}{V_{in}} = -\frac{s^2C_1C_2R_2R_1 + 1}{D(s)} \quad \text{... (2a)}
\]

\[
\frac{V_{o2}}{V_{in}} = \frac{sC_2R_2}{D(s)} \quad \text{... (2b)}
\]

\[
\frac{V_{o3}}{V_{in}} = \frac{s^2C_1C_2R_1R_2}{D(s)} \quad \text{... (2c)}
\]

\[
\frac{V_{o4}}{V_{in}} = -\frac{sC_2R_2}{D(s)} \quad \text{... (2d)}
\]

\[
\frac{V_{o5}}{V_{in}} = -\frac{1}{D(s)} \quad \text{... (2e)}
\]

Here, \(D(s)\) is found as

\[
D(s) = s^2C_1C_2R_1R_2 + sC_2R_2 + 1 \quad \text{... (3)}
\]

Angular resonance frequency \((\omega_o)\), bandwidth \((\omega_o/Q)\) and quality factor \((Q)\) derived from Eq. (3) can be respectively expressed as

\[
\omega_o = \frac{1}{\sqrt{C_1C_2R_1R_2}} \quad \text{... (4a)}
\]

\[
BW = \frac{\omega_o}{Q} = \frac{1}{C_1R_1} \quad \text{... (4b)}
\]

\[
Q = \frac{C_2R_2}{\omega_o} \quad \text{... (4c)}
\]

It is seen from Eq. (4(a-c)) that \(Q\) can be changed by keeping \(\omega_o\) constant or vice versa, which can be achieved by changing only values of both resistors.

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**Fig. 1 — Electrical symbol of the DDCC+**

**Fig. 2 — The proposed second-order voltage-mode universal filter**
For example, if $R_1$ and $R_2$ values are changed by keeping $R_1 \times R_2$ fixed, $\omega_o$ remains constant and $Q$ becomes variable. However, $Q$ of the proposed filter in Fig. 2 cannot be controlled orthogonally. Also, it has unity gains.

From TF in Eq. (2a), an inverting unity gain NF TF with the following phase response is obtained:

$$
\varphi_{nf}(\omega) = \begin{cases} 
\pi - \tan^{-1} \left( \frac{\omega C_2 R_2}{1 - \omega^2 C_2 R_2} \right) & \text{if } \omega < \frac{1}{\sqrt{C_1 C_2 R_2}} \\
\pi & \text{if } \omega = \frac{1}{\sqrt{C_1 C_2 R_2}} \\
2\pi - \tan^{-1} \left( \frac{\omega C_2 R_2}{1 - \omega^2 C_2 R_2} \right) & \text{if } \omega > \frac{1}{\sqrt{C_1 C_2 R_2}}
\end{cases}
$$

... (5)

From TF in Eq. (2b), a non-inverting unity gain BP TF with the following phase response is obtained:

$$
\varphi_{in}(\omega) = \pi - \tan^{-1} \left( \frac{\omega C_2 R_2}{1 - \omega^2 C_2 R_2} \right)
$$

... (6)

From TF in Eq. (2c), a non-inverting unity gain HP TF with the following phase response is obtained:

$$
\varphi_{in}(\omega) = \pi - \tan^{-1} \left( \frac{\omega C_2 R_2}{1 - \omega^2 C_2 R_2} \right)
$$

... (7)

From TF in Eq. (2d), an inverting unity gain BP TF with the following phase response is obtained:

$$
\varphi_{in}(\omega) = -\pi - \tan^{-1} \left( \frac{\omega C_2 R_2}{1 - \omega^2 C_2 R_2} \right)
$$

... (8)

From TF in Eq. (2e), an inverting unity gain LP TF with the following phase response is obtained:

$$
\varphi_{in}(\omega) = \pi - \tan^{-1} \left( \frac{\omega C_2 R_2}{1 - \omega^2 C_2 R_2} \right)
$$

... (9)

If only non-ideal gains are taken into account, TFs in Eq. (2) turn to

$$
\frac{V_{o1}}{V_{in}} = -s^2 C_2 C_1 R_1 R_2 \beta_{22} + s C_2 R_2 \alpha_3 \beta_{21} \beta_{32} \frac{D_{n1}(s)}{D_{n1}(s)} \quad \ldots (10a)
$$

$$
\frac{V_{o2}}{V_{in}} = \frac{s C_2 R_2 \alpha_3 \beta_{21} \beta_{32}}{D_{n1}(s)} \quad \ldots (10b)
$$

$$
\frac{V_{o3}}{V_{in}} = \frac{s^2 C_2 C_1 R_1 R_2 \beta_{22}}{D_{n1}(s)} \quad \ldots (10c)
$$

$$
\frac{V_{o4}}{V_{in}} = -\frac{s C_2 R_2 \alpha_3 \beta_{21} \beta_{32}}{D_{n1}(s)} \quad \ldots (10d)
$$

$$
\frac{V_{o5}}{V_{in}} = \frac{\alpha_3 \beta_{21} \beta_{32}}{D_{n1}(s)} \quad \ldots (10e)
$$

Here, $D_{n1}(s)$ is found as

$$
D_{n1}(s) = s^2 C_2 C_1 R_1 R_2 + s C_2 R_2 \alpha_3 \beta_{21} \beta_{32} + \alpha_3 \beta_{21} \beta_{32}
$$

... (11)

If non-ideal gains are considered, $\omega_o$, $\omega_o/Q$ and $Q$ derived from Eq. (11) can be respectively given as

$$
\omega_o = \sqrt{\frac{\alpha_3 \beta_{21} \beta_{32}}{C_1 C_2 R_2}} \quad \ldots (12a)
$$

$$
BW = \frac{\omega_o}{Q} = \frac{\alpha_3 \beta_{21} \beta_{32}}{C_1 R_1} \quad \ldots (12b)
$$

$$
Q = \frac{1}{\beta_{21} \beta_{32}} \sqrt{\frac{C_1 R_1}{C_2 R_2}} \quad \ldots (12c)
$$

From Eq. (12), active and passive component sensitivities with respect to $\omega_o$, $\omega_o/Q$ and $Q$ are respectively given below

$$
\begin{align*}
S &= S = S = \frac{1}{2} \\
\frac{a_2}{a_3} &= \frac{a_3}{a_2} \quad \ldots (13b) \\
\frac{a_2}{a_3} &= \frac{a_3}{a_2} \quad \ldots (13c) \\
\frac{a_2}{a_3} &= \frac{a_3}{a_2} \quad \ldots (13d)
\end{align*}
$$

It is observed from above that all of the active and passive component sensitivities with respect to $\omega_o$, $\omega_o/Q$ and $Q$ are no more than unity in magnitude. If the switch is connected to AP, all the TFs in Eq. (2) except $V_{o4}/V_{in}$ remain the same. Thus, $V_{o4}/V_{in}$ is evaluated as

$$
\frac{V_{o4}}{V_{in}} = -\frac{s^2 C_2 C_1 R_1 R_2 + s C_2 R_2 + 1}{D(s)} \quad \ldots (13)
$$

Phase response for the AP filter in Eq. (13) is computed as follows:
It is observed from Eq. (14) that as the frequency goes from zero to infinity, \( \varphi_{AP}(\omega) \) changes from 180° to -180°. If only non-ideal gains are considered, the following TFs for the second case can be obtained as:

\[
V_{a1} = \frac{s^2C_2C_2R_2\beta_{12}\beta_{22} - sC_2R_2\alpha_2\beta_{12}\beta_{22} + \alpha_2\alpha_2\beta_{12}\beta_{22}}{D_{a2}(s)}
\]

\[
V_{a2} = \frac{sC_2R_2\alpha_2\beta_{12}\beta_{22}}{D_{a2}(s)}
\]

\[
V_{a3} = \frac{s^2C_2C_2R_2\beta_{12}\beta_{22}}{D_{a2}(s)}
\]

\[
V_{a4} = \frac{-sC_2R_2\alpha_2\beta_{12}\beta_{22}}{D_{a2}(s)}
\]

\[
V_{a5} = \frac{-\alpha_2\alpha_2\beta_{12}\beta_{22}}{D_{a2}(s)}
\]

where \( D_{a2}(s) \) is found as

\[
D_{a2}(s) = s^2C_2C_2R_2 + sC_2R_2\alpha_2(\beta_{12}\beta_{22} + \beta_{12}\beta_{22} - \beta_{22}) + \alpha_2\alpha_2\beta_{12}\beta_{22}
\]

If non-ideal gains are taken into account, \( \omega_o, \omega_o/Q \) and \( Q \) derived from Eq. (16) can be respectively given as follows:

\[
\omega_o = \frac{\alpha_2\alpha_2\beta_{12}\beta_{22}}{C_2C_2R_2}
\]

\[
BW = \frac{\omega_o}{Q} = \frac{\alpha_2(\beta_{12}\beta_{22} + \beta_{12}\beta_{22} - \beta_{22})}{C_2R_2}
\]

\[
Q = \frac{1}{\beta_{12}\beta_{22} + \beta_{12}\beta_{22} - \beta_{22}}
\]

From Eq. (17), active and passive component sensitivities with respect to \( \omega_o, \omega_o/Q \) and \( Q \) are respectively given as in the following:
\[
\frac{V_{o1}}{V_{in}} = \frac{R_1}{R_3} \frac{sC_z R_z + 1}{D(s)} \quad ... (18d)
\]
\[
\frac{V_{o5}}{V_{in}} = \frac{R_4}{R_3} \frac{1}{D(s)} \quad ... (18e)
\]

If the switch is connected to NF in Fig. 3 and \( V_{in2} = V_{in} \) and \( V_{in1} = 0 \) are chosen, the following TFs are simultaneously obtained:

\[
\frac{V_{o1}}{V_{in}} = \frac{R_1}{R_3} \frac{s^2C_zR_zR_2 + 1}{D(s)} \quad ... (19a)
\]
\[
\frac{V_{o2}}{V_{in}} = - \frac{R_4}{R_3} \frac{sC_z R_z}{D(s)} \quad ... (19b)
\]
\[
\frac{V_{o3}}{V_{in}} = \frac{R_1}{R_3} \frac{s^2C_zR_zR_2}{D(s)} \quad ... (19c)
\]
\[
\frac{V_{o4}}{V_{in}} = \frac{R_4}{R_3} \frac{sC_z R_z}{D(s)} \quad ... (19d)
\]
\[
\frac{V_{o5}}{V_{in}} = \frac{R_4}{R_3} \frac{1}{D(s)} \quad ... (19e)
\]

If the switch is connected to AP in Fig. 3 and \( V_{in1} = V_{in} \) and \( V_{in2} = 0 \) are chosen, all the TFs in Eq. (18) except \( V_{o1}/V_{in} \) remain the same. Thus, \( V_{o1}/V_{in} \) is evaluated as

\[
\frac{V_{o1}}{V_{in}} = \frac{R_1}{R_3} \frac{s^2C_zR_zR_2 - sC_z R_z + 1}{D(s)} \quad ... (20)
\]

If the switch is connected to AP in Fig. 3 and \( V_{in2} = V_{in} \) and \( V_{in1} = 0 \) are chosen, all the TFs in Eq. (19) except \( V_{o1}/V_{in} \) remain the same. Thus, \( V_{o1}/V_{in} \) is evaluated as

\[
\frac{V_{o1}}{V_{in}} = \frac{R_1}{R_3} \frac{s^2C_zC_zR_zR_2 - sC_zR_z + 1}{D(s)} \quad ... (21)
\]

Phase response for the AP filter in Eq. (21) is computed as follows:

\[
\varphi_{AP}(\omega) = -2\tan^{-1}\left(\frac{\omega C_z R_z}{1 - \omega^2 C_z C_z R_z R_2}\right) \quad ... (22)
\]

It is seen from Eq. (22) that as the frequency goes from zero to infinity, \( \varphi_{AP}(\omega) \) changes from 0° to -360°. Apart from this, the proposed filters in Figs 2 and 3 have the property of high input impedances; thus, they can be easily cascaded with previous stages without needing any voltage buffers. NF and AP responses have the feature of low output impedances; accordingly, a load can be directly connected without requiring any voltage buffers. However, LP, BP and HP responses need voltage buffers if loads are connected.

**Parasitic Impedance Effects on the Filter Performances**

DDCC+ with its parasitic impedances which is demonstrated in Fig. 4 is defined by the following matrix equation:

\[
\begin{bmatrix}
I_{x1} \\
I_{x2} \\
I_{x3} \\
I_{x4}
\end{bmatrix}
= \begin{bmatrix}
1 & 0 & 0 & 0 \\
0 & sC_{x1} & 1/R_{x1} & 0 \\
0 & 0 & sC_{x2} & 0 \\
0 & 0 & 0 & sC_{x3}
\end{bmatrix}
\begin{bmatrix}
I_{x} \\
V_{y1} \\
V_{y2} \\
V_{y3}
\end{bmatrix}
\]

... (23)

Here, \( R_{xk} (k=1, 2, 3) \) ideally equal to zero and \( R_{zk+} \) ideally equal to infinity are X and Z terminal parasitic

---

**Fig. 3** — The proposed second-order universal filter with gain

**Fig. 4** — DDCC+ with its parasitic impedances
resistors of the \( k^{th} \) DDCC+, respectively. \( C_{2k} \) is \( Z \) terminal parasitic capacitor of the \( k^{th} \) DDCC+. Also, \( C_{yj} \) (\( j=1, 2, 3 \)) is \( Y \) terminal parasitic capacitor of the \( k^{th} \) DDCC+ where \( j \) represents the \( j^{th} \) \( Y \) terminal of each DDCC+.

If the switch is connected to NF as an example, the following TFs by only considering X terminal parasitic resistors for simplicity are simultaneously obtained:

\[
\frac{V_{o1}}{V_{in}} = -1 + \frac{sC_2(R_2 + R_{x3})}{D_{p1}(s)} \quad \cdots (24a)
\]

\[
\frac{V_{o2}}{V_{in}} = \frac{sC_2(R_2 + R_{x3})}{D_{p2}(s)} \quad \cdots (24b)
\]

\[
\frac{V_{o3}}{V_{in}} = \frac{s^2C_2C_1R_2(R_2 + R_{x3})}{D_{p3}(s)} \quad \cdots (24c)
\]

\[
\frac{V_{o4}}{V_{in}} = \frac{sC_2R_2}{D_{p4}(s)} \quad \cdots (24d)
\]

\[
\frac{V_{o5}}{V_{in}} = -\frac{1}{D_{p5}(s)} \quad \cdots (24e)
\]

where \( D_{p1}(s) \) is computed as follows:

\[
D_{p1}(s) = s^2C_1C_2(R_2R_1 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3}) + sC_2(R_2 + R_{x3}) + 1
\]

\[
\cdots (25)
\]

If only X terminal parasitic resistors are taken into account, \( \omega_o, \omega_o/Q \) and \( Q \) derived from Eq. (25) can be respectively given as

\[
\omega_o = \frac{1}{\sqrt{C_1C_2(R_2R_1 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3})}} \quad \cdots (26a)
\]

\[
BW = \frac{\omega_o}{Q} = \frac{R_2 + R_{x3}}{C_1(R_2R_1 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3})} \quad \cdots (26b)
\]

\[
Q = \frac{1}{R_2 + R_{x3}} \sqrt{\frac{C_1(R_2R_1 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3})}{C_2}} \quad \cdots (26c)
\]

It is seen from Eq. (26) that \( \omega_o, \omega_o/Q \) and \( Q \) a bit change due to X terminal parasitic resistors of the DDCC+. If the switch is connected to NF as an example, the following TFS by only considering Z and Y terminal parasitic impedances for simplicity are simultaneously obtained:

\[
V_{o1} = -1 + \frac{1}{r_{R_{x1}R_{x1}} \left( \frac{1}{R_2} + \frac{1}{R_{x1}} + s(C_1 + C_{x2} + C_{x3} + C_{x4}) \right) \left( \frac{1}{R_2} + s(C_1 + C_{x2} + C_{x3} + C_{x4}) \right)} \quad \cdots (27a)
\]

\[
V_{o2} = \frac{1}{r_{R_{x1}R_{x1}} \left( \frac{1}{R_2} + \frac{1}{R_{x1}} + s(C_1 + C_{x2} + C_{x3} + C_{x4}) \right) \left( \frac{1}{R_2} + s(C_1 + C_{x2} + C_{x3} + C_{x4}) \right)} \quad \cdots (27b)
\]

\[
V_{o3} = \frac{s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3}}{D_{p1}(s)} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} \quad \cdots (27c)
\]

\[
V_{o4} = \frac{1}{r_{R_{x1}R_{x1}} \left( \frac{1}{R_2} + \frac{1}{R_{x1}} + s(C_1 + C_{x2} + C_{x3} + C_{x4}) \right) \left( \frac{1}{R_2} + s(C_1 + C_{x2} + C_{x3} + C_{x4}) \right)} \quad \cdots (27d)
\]

\[
V_{o5} = -\frac{R_{x2}R_{x3}}{D_{p5}(s)} \quad \cdots (27e)
\]

Here, \( D_{p5}(s) \) is calculated as

\[
D_{p5}(s) = s^2(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} + s(C_1 + C_{x2} + C_{x3} + C_{x4})R_2R_{x2}R_{x3} \quad \cdots (28)
\]

Results and Discussion

Simulations of the proposed universal filter are achieved by SPICE program using 0.13 \( \mu \)m IBM CMOS technology parameters given in Table 1. Internal structure of the DDCC+ derived from one in Ref.1 is given in Fig. 5. Symmetrical DC power supply voltages are selected as \( V_{DD} = +V_{SS} = 0.75 \) V. Bias voltage \( V_B \) is taken as 0.2 V. Dimensions of the MOS transistors are given in Table 2. Passive components of the proposed filter of Fig. 2 in the simulations of Figs 6-11 are chosen as \( C_1=C_2=50 \) pF and \( R_I=R_2=1 \) k\( \Omega \) to obtain the SIMO VM second-order universal filter.
Table 1 — 0.13 μm IBM CMOS technology parameters

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<td>1.2297627 DVT1 = 0.1473877 DVT2 = 0.295815</td>
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<td>U0</td>
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MODEL CMOS SIMPLE MOS (LEVEL = 7)

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<td>0 LLN = 1 LWW = 0</td>
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<td>LWN</td>
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Simulation and ideal LP, BP and HP responses are given in Fig. 6. Simulation and ideal phase and gain responses for the NF are shown in Fig. 7. Simulation and ideal phase and gain responses for the AP are shown in Fig. 8. Time domain responses for the AP filter at resonance frequency are demonstrated in Fig. 9 where a 100 mV peak sinusoidal input voltage signal is applied. Also, total harmonic distortion (THD) variations for the AP filter at resonance frequency with respect to applied peak sinusoidal input voltage signal are given in Fig. 10. One observes from Fig. 10 that the proposed VM second-order universal filter can be operated properly up to approximately 150 mV applied peak input voltage signal. Input and corresponding output noises at resonance frequency for the AP filter are respectively calculated as 5.25×10^8 V/Hz and 4.46×10^8 V/Hz. Also, a Monte Carlo analysis with twenty runs is performed in Fig. 11 where both resistor values are changed 10% uniformly.

Orthogonality of the proposed filter in Fig. 2 is shown in Fig. 12 where C1=C2=50 pf and f0≈3.18 MHz are chosen. However, resistors are chosen as R1=0.5

**Table 2 — Dimensions of the transistors of the DDCC+ in Fig. 5**

PMOS Transistors $W_L(\mu m)/L(\mu m)$

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<th>M1</th>
<th>M2</th>
</tr>
</thead>
<tbody>
<tr>
<td>39</td>
<td>1.04</td>
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</tbody>
</table>

NMOS Transistors $W_L(\mu m)/L(\mu m)$

<table>
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<th>M12</th>
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<tr>
<td>15</td>
<td>1.04</td>
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</table>
kΩ and \( R_2 = 2 \) kΩ yielding \( Q = 0.5 \), \( R_1 = 1 \) kΩ and \( R_2 = 1 \) kΩ yielding \( Q = 1 \), \( R_1 = 2 \) kΩ and \( R_2 = 0.5 \) kΩ yielding \( Q = 2 \) and \( R_1 = 4 \) kΩ and \( R_2 = 0.25 \) kΩ resulting in \( Q = 4 \). Apart from this, LP, BP and HP responses of the filter in Fig. 3 are given in Fig. 13 in which \( C_1 = C_2 = 50 \) pF, \( R_1 = R_2 = R_3 = 1 \) kΩ and \( R_4 = 2 \) kΩ resulting in \( f_o \approx 3.18 \) MHz, \( Q = 1 \) and gain=2. Also, power consumption of the proposed biquadratic universal filter in Fig. 3 is found as 4.36 mW.

It is seen from Figs 6-13 that simulation and ideal results are close to each other whereas a bit discrepancy...
Table 3 — Comparison of the previously published DDCC based filters\textsuperscript{1-25} and the proposed one

<table>
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<tr>
<th>References</th>
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<th># of Resistor(s)</th>
<th># of Capacitor(s)</th>
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<th>High Input</th>
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Present work | 3 | 2 | 0 | 2 | Yes | Yes | Yes | No | ±0.75V | 0.13 \(\mu\)m | 2.62 |

NA: not available\textsuperscript{a} do not use DDCC with single \(Z^+\) terminal, \textsuperscript{b} use extra one FDCCII\textsuperscript{b} employ additional two OTAs

Fig. 12 — Orthogonality of the proposed filter in Fig. 2

Fig. 13 — Low-pass, band-pass and high-pass responses with gain two

Conclusions

In this paper, a new SIMO VM universal filter configuration with the property of high input impedance is proposed. The proposed filter can realize all the standard second-order universal filter responses...
such as LP, BP, HP, NF and AP. NF and AP responses have low output impedances but LP, BP and HP responses do not have the feature of low output impedances. Each time, one of NF and AP responses is obtained via a switch. Moreover, other responses can be obtained simultaneously. The proposed filter consists of three standard DDCC+s and a minimum number of only grounded passive components, and does not require any critical passive component matching conditions and cancellation constraints; thus, it is suitable for IC process. It has only resistors but no capacitors connected in series to X terminals of the DDCC+s; accordingly, it can be operated at high frequencies. A universal filter with gain is obtained by adding extra one DDCC+ and two grounded resistors. Further, the universal filter with gain can provide both inverting and non-inverting second-order VM LP, BP, HP, NF and AP TFs. Some computer simulations based on SPICE program confirm the theory well as desired. It is expected that the proposed second-order VM universal filter will be beneficial in a number of areas such as signal processing, control and communication systems.

References
26 Ferri G & Guerrini N C, Low voltage, low power CMOS current conveyors, (Springer), 2003.