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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 amplifier[1]. Therefore, a number of DDCC based voltage-mode (VM) second-order filters[1-25] 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) OAs[26]. However, DDCC based VM second-order filters[1-25] 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 responses[1,3,9,14,16,18,19,23-25].
(iii) Require critical passive component matching condition(s)[4,7,10,12-15,18].
(iv) Each time, provide only one response[3,5,10,13,18,21,24].
(v) Do not use plus-type DDCC (DDCC+) with single Z terminal[7,9-17,20,22-24].
(vi) Do not have the property of high input impedance[4,7,9,11,12,14,17,20].
(vii) Have transfer functions (TFs) with complex combination of input signals[2,5,6,8,10,11,12,13,15,17,20-22].
(viii) Have a capacitor connected in series to X terminal of the DDCC[9]; accordingly, they cannot be operated properly at high frequencies[27].
(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 frequencies[28].

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) fabrication[29-31]. However, LP, BP and HP responses do not provide low output

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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

$$\begin{bmatrix} I_{z+} \\ I_{y1} \\ I_{y2} \\ I_{y3} \\ V_x \end{bmatrix} = \begin{bmatrix} \alpha_k & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & \beta_{k1} & -\beta_{k2} & \beta_{k3} \end{bmatrix} \begin{bmatrix} I_x \\ V_{y1} \\ V_{y2} \\ V_{y3} \end{bmatrix} \quad \ldots (1)$$

In matrix Eq. (1), at sufficiently low frequencies, frequency dependent non-ideal current gain $\alpha_k=1+\varepsilon_k$ (k=1, 2, 3 represents the k^{th} DDCC+) and frequency dependent non-ideal voltage gains $\beta_{kj}=1+\varepsilon_{kj}$ (j=1, 2, 3) are ideally equal to unity. In addition, ε_k and ε_{kj}, ideally equal to zero, are respectively called as current and voltage tracking errors where $|\varepsilon_k|<<1$ and $|\varepsilon_{kj}|<<1$. 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:

$$\frac{V_{o1}}{V_{in}} = -\frac{s^2C_1C_2R_1R_2+1}{D(s)} \quad \ldots (2a)$$

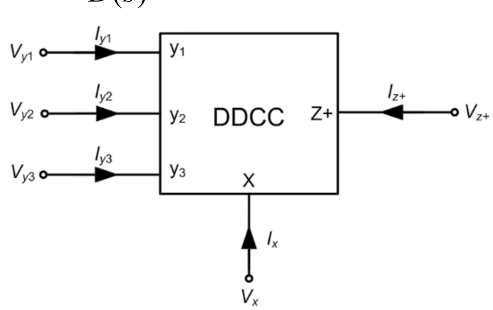

Fig. 1 — Electrical symbol of the DDCC+

$$\frac{V_{o2}}{V_{in}} = \frac{sC_2R_2}{D(s)} \quad \ldots (2b)$$

$$\frac{V_{o3}}{V_{in}} = \frac{s^2C_1C_2R_1R_2}{D(s)} \quad \ldots (2c)$$

$$\frac{V_{o4}}{V_{in}} = -\frac{sC_2R_2}{D(s)} \quad \ldots (2d)$$

$$\frac{V_{o5}}{V_{in}} = -\frac{1}{D(s)} \quad \ldots (2e)$$

Here, $D(s)$ is found as

$$D(s) = s^2C_1C_2R_1R_2 + sC_2R_2 + 1 \quad \ldots (3)$$

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

$$\omega_o = \frac{1}{\sqrt{C_1C_2R_1R_2}} \quad \ldots (4a)$$

$$BW = \frac{\omega_o}{Q} = \frac{1}{C_1R_1} \quad \ldots (4b)$$

$$Q = \sqrt{\frac{C_1R_1}{C_2R_2}} \quad \ldots (4c)$$

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

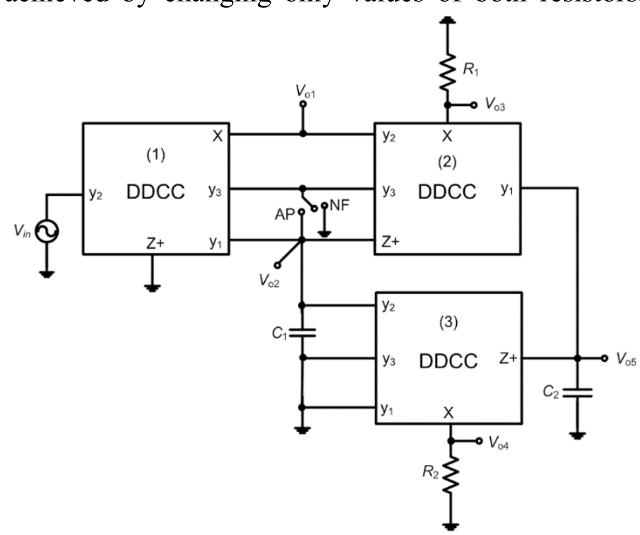

Fig. 2 — The proposed second-order voltage-mode universal filter

For example, if R_1 and R_2 values are changed by keeping $R_1 \times R_2$ fixed, ω_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[35]:

$$\varphi_{\mathrm{NF}}(\omega)=\begin{cases}\pi-\tan^{-1}\left(\dfrac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) & \text{if } \omega<\dfrac{1}{\sqrt{C_1C_2R_1R_2}}\\ \pi & \text{if } \omega=\dfrac{1}{\sqrt{C_1C_2R_1R_2}}\\ 2\pi-\tan^{-1}\left(\dfrac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) & \text{if } \omega>\dfrac{1}{\sqrt{C_1C_2R_1R_2}}\end{cases} \quad \text{… (5)}$$

From TF in Eq. (2b), a non-inverting unity gain BP TF with the following phase response is obtained[35]:

$$\varphi_{\mathrm{BP}}(\omega)=\frac{\pi}{2}-\tan^{-1}\left(\frac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) \quad \text{… (6)}$$

From TF in Eq. (2c), a non-inverting unity gain HP TF with the following phase response is obtained[35]:

$$\varphi_{\mathrm{HP}}(\omega)=\pi-\tan^{-1}\left(\frac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) \quad \text{… (7)}$$

From TF in Eq. (2d), an inverting unity gain BP TF with the following phase response is obtained[35]:

$$\varphi_{\mathrm{BP}}(\omega)=-\frac{\pi}{2}-\tan^{-1}\left(\frac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) \quad \text{… (8)}$$

From TF in Eq. (2e), an inverting unity gain LP TF with the following phase response is obtained[35]:

$$\varphi_{\mathrm{LP}}(\omega)=\pi-\tan^{-1}\left(\frac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) \quad \text{… (9)}$$

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

$$\frac{V_{o1}}{V_{in}}=-\frac{s^2C_1C_2R_1R_2\beta_{12}+\alpha_2\alpha_3\beta_{12}\beta_{21}\beta_{32}}{D_{n1}(s)} \quad \text{… (10a)}$$

$$\frac{V_{o2}}{V_{in}}=\frac{sC_2R_2\alpha_2\beta_{12}\beta_{22}}{D_{n1}(s)} \quad \text{… (10b)}$$

$$\frac{V_{o3}}{V_{in}}=\frac{s^2C_1C_2R_1R_2\beta_{12}\beta_{22}}{D_{n1}(s)} \quad \text{… (10c)}$$

$$\frac{V_{o4}}{V_{in}}=-\frac{sC_2R_2\alpha_2\beta_{12}\beta_{22}\beta_{32}}{D_{n1}(s)} \quad \text{… (10d)}$$

$$\frac{V_{o5}}{V_{in}}=-\frac{\alpha_2\alpha_3\beta_{12}\beta_{22}\beta_{32}}{D_{n1}(s)} \quad \text{… (10e)}$$

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

$$D_{n1}(s)=s^2C_1C_2R_1R_2+sC_2R_2\alpha_2\beta_{11}\beta_{22}+\alpha_2\alpha_3\beta_{21}\beta_{32} \quad \text{… (11)}$$

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

$$\omega_o=\sqrt{\frac{\alpha_2\alpha_3\beta_{21}\beta_{32}}{C_1C_2R_1R_2}} \quad \text{… (12a)}$$

$$BW=\frac{\omega_o}{Q}=\frac{\alpha_2\beta_{11}\beta_{22}}{C_1R_1} \quad \text{… (12b)}$$

$$Q=\frac{1}{\beta_{11}\beta_{22}}\sqrt{\frac{C_1R_1\alpha_3\beta_{21}\beta_{32}}{C_2R_2\alpha_2}} \quad \text{… (12c)}$$

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

$$S_{\alpha_2}^{\omega_o}=S_{\alpha_3}^{\omega_o}=S_{\beta_{21}}^{\omega_o}=S_{\beta_{32}}^{\omega_o}=-S_{C_1}^{\omega_o}=-S_{C_2}^{\omega_o}=-S_{R_1}^{\omega_o}=-S_{R_2}^{\omega_o}=\frac{1}{2}$$

$$S_{\alpha_2}^{BW}=S_{\beta_{11}}^{BW}=S_{\beta_{22}}^{BW}=-S_{C_1}^{BW}=-S_{R_1}^{BW}=1$$

$$-S_{\alpha_2}^{Q}=S_{\alpha_3}^{Q}=S_{\beta_{21}}^{Q}=S_{\beta_{32}}^{Q}=S_{C_1}^{Q}=-S_{C_2}^{Q}=S_{R_1}^{Q}=-S_{R_2}^{Q}=\frac{1}{2}$$

$$S_{\beta_{11}}^{Q}=S_{\beta_{22}}^{Q}=-1$$

It is observed from above that all of the active and passive component sensitivities with respect to ω_o, ω_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_{o1}/V_{in} remain the same. Thus, V_{o1}/V_{in} is evaluated as

$$\frac{V_{o1}}{V_{in}}=-\frac{s^2C_1C_2R_1R_2-sC_2R_2+1}{D(s)} \quad \text{… (13)}$$

Phase response for the AP filter in Eq. (13) is computed as follows:

$$\varphi_{\mathrm{AP}}(\omega)=\pi-2\tan^{-1}\left(\frac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) \quad \ldots (14)$$

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

$$\frac{V_{o1}}{V_{in}}=-\frac{s^2C_1C_2R_1R_2\beta_{12}-sC_2R_2\alpha_2\beta_{12}\beta_{23}+\alpha_2\alpha_3\beta_{12}\beta_{21}\beta_{32}}{D_{n2}(s)} \quad \ldots (15a)$$

$$\frac{V_{o2}}{V_{in}}=\frac{sC_2R_2\alpha_2\beta_{12}\beta_{22}}{D_{n2}(s)} \quad \ldots (15b)$$

$$\frac{V_{o3}}{V_{in}}=\frac{s^2C_1C_2R_1R_2\beta_{12}\beta_{22}}{D_{n2}(s)} \quad \ldots (15c)$$

$$\frac{V_{o4}}{V_{in}}=-\frac{sC_2R_2\alpha_2\beta_{12}\beta_{22}\beta_{32}}{D_{n2}(s)} \quad \ldots (15d)$$

$$\frac{V_{o5}}{V_{in}}=-\frac{\alpha_2\alpha_3\beta_{12}\beta_{22}\beta_{32}}{D_{n2}(s)} \quad \ldots (15e)$$

where $D_{n2}(s)$ is found as

$$D_{n2}(s)=s^2C_1C_2R_1R_2+sC_2R_2\alpha_2(\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23})+\alpha_2\alpha_3\beta_{21}\beta_{32} \quad \ldots (16)$$

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

$$\omega_o=\sqrt{\frac{\alpha_2\alpha_3\beta_{21}\beta_{32}}{C_1C_2R_1R_2}} \quad \ldots (17a)$$

$$BW=\frac{\omega_o}{Q}=\frac{\alpha_2(\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23})}{C_1R_1} \quad \ldots (17b)$$

$$Q=\frac{1}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}\sqrt{\frac{C_1R_1\alpha_3\beta_{21}\beta_{32}}{C_2R_2\alpha_2}} \quad \ldots (17c)$$

From Eq. (17), active and passive component sensitivities with respect to ω_o, ω_o/Q and Q are respectively given as in the following:

$$S_{\alpha_2}^{\omega_o}=S_{\alpha_3}^{\omega_o}=S_{\beta_{21}}^{\omega_o}=S_{\beta_{32}}^{\omega_o}=-S_{C_1}^{\omega_o}=-S_{C_2}^{\omega_o}=-S_{R_1}^{\omega_o}=-S_{R_2}^{\omega_o}=\frac{1}{2}$$

$$S_{\alpha_2}^{BW}=-S_{C_1}^{BW}=-S_{R_1}^{BW}=1$$

$$S_{\beta_{11}}^{BW}=\frac{\beta_{11}\beta_{22}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$S_{\beta_{13}}^{BW}=\frac{\beta_{13}\beta_{22}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$S_{\beta_{23}}^{BW}=-\frac{\beta_{23}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$S_{\beta_{22}}^{BW}=\frac{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$-S_{\alpha_2}^{Q}=S_{\alpha_3}^{Q}=S_{\beta_{21}}^{Q}=S_{\beta_{32}}^{Q}=S_{C_1}^{Q}=-S_{C_2}^{Q}=S_{R_1}^{Q}=-S_{R_2}^{Q}=\frac{1}{2}$$

$$S_{\beta_{11}}^{Q}=-\frac{\beta_{11}\beta_{22}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$S_{\beta_{13}}^{Q}=-\frac{\beta_{13}\beta_{22}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$S_{\beta_{23}}^{Q}=\frac{\beta_{23}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

$$S_{\beta_{22}}^{Q}=-\frac{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}}{\beta_{11}\beta_{22}+\beta_{13}\beta_{22}-\beta_{23}}$$

It is seen from above that all of the passive component sensitivities with respect to ω_o, ω_o/Q and Q are no more than unity in magnitude. The proposed universal filter with a gain is given in Fig. 3. The filter in Fig. 3 is obtained from one in Fig. 2 by adding extra one DDCC+ and two grounded resistors. Hence, the filter in Fig. 3 can provide both inverting and non-inverting second-order VM universal filter responses with gains which increase flexibility in IC technology. If the switch is connected to NF in Fig. 3 and $V_{in1}=V_{in}$ and $V_{in2}=0$ are chosen, the following TFs are simultaneously obtained:

$$\frac{V_{o1}}{V_{in}}=-\frac{R_4}{R_3}\frac{s^2C_1C_2R_1R_2+1}{D(s)} \quad \ldots (18a)$$

$$\frac{V_{o2}}{V_{in}}=\frac{R_4}{R_3}\frac{sC_2R_2}{D(s)} \quad \ldots (18b)$$

$$\frac{V_{o3}}{V_{in}}=\frac{R_4}{R_3}\frac{s^2C_1C_2R_1R_2}{D(s)} \quad \ldots (18c)$$

$$\frac{V_{o4}}{V_{in}} = -\frac{R_4}{R_3}\frac{sC_2R_2}{D(s)} \qquad \text{… (18d)}$$

$$\frac{V_{o5}}{V_{in}} = -\frac{R_4}{R_3}\frac{1}{D(s)} \qquad \text{… (18e)}$$

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:

$$\frac{V_{o1}}{V_{in}} = \frac{R_4}{R_3}\frac{s^2C_1C_2R_1R_2+1}{D(s)} \qquad \text{… (19a)}$$

$$\frac{V_{o2}}{V_{in}} = -\frac{R_4}{R_3}\frac{sC_2R_2}{D(s)} \qquad \text{… (19b)}$$

$$\frac{V_{o3}}{V_{in}} = -\frac{R_4}{R_3}\frac{s^2C_1C_2R_1R_2}{D(s)} \qquad \text{… (19c)}$$

$$\frac{V_{o4}}{V_{in}} = \frac{R_4}{R_3}\frac{sC_2R_2}{D(s)} \qquad \text{… (19d)}$$

$$\frac{V_{o5}}{V_{in}} = \frac{R_4}{R_3}\frac{1}{D(s)} \qquad \text{… (19e)}$$

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

$$\frac{V_{o1}}{V_{in}} = -\frac{R_4}{R_3}\frac{s^2C_1C_2R_1R_2 - sC_2R_2 + 1}{D(s)} \qquad \text{… (20)}$$

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

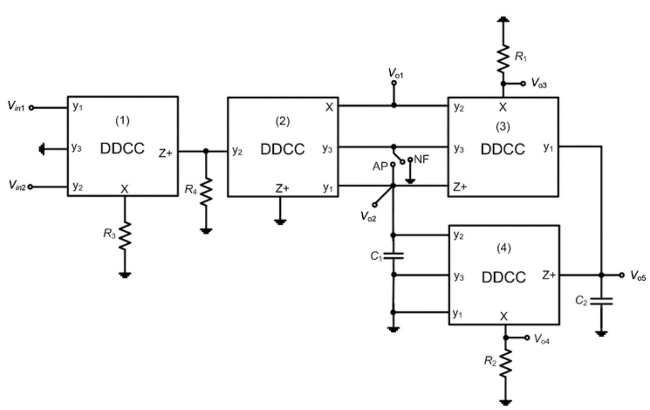

Fig. 3 — The proposed second-order universal filter with gain

$$\frac{V_{o1}}{V_{in}} = \frac{R_4}{R_3}\frac{s^2C_1C_2R_1R_2 - sC_2R_2 + 1}{D(s)} \qquad \text{… (21)}$$

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

$$\varphi_{\text{AP}}(\omega) = -2\tan^{-1}\left(\frac{\omega C_2R_2}{1-\omega^2C_1C_2R_1R_2}\right) \qquad \text{… (22)}$$

It is seen from Eq. (22) that as the frequency goes from zero to infinity, $\varphi_{\text{AP}}(\omega)$ 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:

$$\begin{bmatrix} I_{z+} \\ I_{y1} \\ I_{y2} \\ I_{y3} \\ V_x \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 & sC_{zk+}+1/R_{zk+} \\ 0 & sC_{yk1} & 0 & 0 & 0 \\ 0 & 0 & sC_{yk2} & 0 & 0 \\ 0 & 0 & 0 & sC_{yk3} & 0 \\ R_{xk} & 1 & -1 & 1 & 0 \end{bmatrix} \begin{bmatrix} I_x \\ V_{y1} \\ V_{y2} \\ V_{y3} \\ V_{z+} \end{bmatrix} \qquad \text{… (23)}$$

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

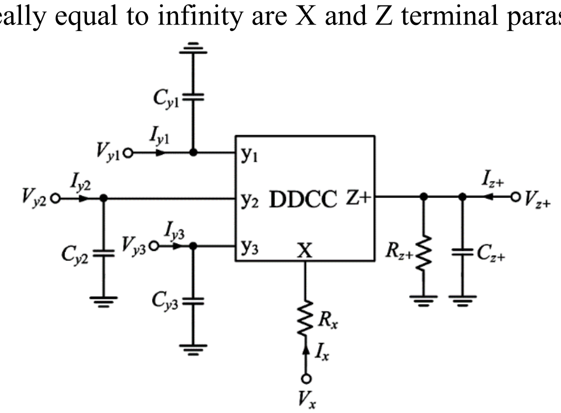

Fig. 4 — DDCC+ with its parasitic impedances

resistors of the k^{th} DDCC+, respectively. C_{zk+} is Z terminal parasitic capacitor of the k^{th} DDCC+. Also, C_{ykj} (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:

$$\frac{V_{o1}}{V_{in}} = -1 + \frac{sC_2(R_2 + R_{x3})}{D_{px}(s)} \quad \text{… (24a)}$$

$$\frac{V_{o2}}{V_{in}} = \frac{sC_2(R_2 + R_{x3})}{D_{px}(s)} \quad \text{… (24b)}$$

$$\frac{V_{o3}}{V_{in}} = \frac{s^2 C_1 C_2 R_1 (R_2 + R_{x3})}{D_{px}(s)} \quad \text{… (24c)}$$

$$\frac{V_{o4}}{V_{in}} = -\frac{sC_2 R_2}{D_{px}(s)} \quad \text{… (24d)}$$

$$\frac{V_{o5}}{V_{in}} = -\frac{1}{D_{px}(s)} \quad \text{… (24e)}$$

where $D_{px}(s)$ is computed as follows:

$$D_{px}(s) = s^2 C_1 C_2 (R_1R_2 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3}) + sC_2(R_2 + R_{x3}) + 1 \quad \text{… (25)}$$

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

$$\omega_o = \sqrt{\frac{1}{C_1C_2(R_1R_2 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3})}} \quad \text{… (26a)}$$

$$BW = \frac{\omega_o}{Q} = \frac{R_2 + R_{x3}}{C_1(R_1R_2 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3})} \quad \text{… (26b)}$$

$$Q = \frac{1}{R_2 + R_{x2}}\sqrt{\frac{C_1(R_1R_2 + R_2R_{x2} + R_1R_{x3} + R_{x2}R_{x3})}{C_2}} \quad \text{… (26c)}$$

It is seen from Eq. (26) that ω_o, ω_o/Q and Q a bit change due to X terminal parasitic resistors of the DDCC+s. 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:

$$\frac{V_{o1}}{V_{in}} = -1 + \frac{1}{R_1R_{z3+}\left(\frac{1}{R_1R_2} + \left(\frac{1}{R_1} + \frac{1}{R_{z2+}} + s(C_1 + C_{y32} + C_{y11} + C_{z2+})\right)\left(\frac{1}{R_{z3+}} + s(C_2 + C_{z3+} + C_{y21})\right)\right)} + \frac{s(C_2 + C_{z3+} + C_{y21})}{R_1\left(\frac{1}{R_1R_2} + \left(\frac{1}{R_1} + \frac{1}{R_{z2+}} + s(C_1 + C_{y32} + C_{y11} + C_{z2+})\right)\left(\frac{1}{R_{z3+}} + s(C_2 + C_{z3+} + C_{y21})\right)\right)} \quad \text{… (27a)}$$

$$\frac{V_{o2}}{V_{in}} = \frac{1}{R_1R_{z3+}\left(\frac{1}{R_1R_2} + \left(\frac{1}{R_1} + \frac{1}{R_{z2+}} + s(C_1 + C_{y32} + C_{y11} + C_{z2+})\right)\left(\frac{1}{R_{z3+}} + s(C_2 + C_{z3+} + C_{y21})\right)\right)} + \frac{s(C_2 + C_{z3+} + C_{y21})}{R_1\left(\frac{1}{R_1R_2} + \left(\frac{1}{R_1} + \frac{1}{R_{z2+}} + s(C_1 + C_{y32} + C_{y11} + C_{z2+})\right)\left(\frac{1}{R_{z3+}} + s(C_2 + C_{z3+} + C_{y21})\right)\right)} \quad \text{… (27b)}$$

$$\frac{V_{o3}}{V_{in}} = \frac{s^2(C_1 + C_{y32} + C_{y11} + C_{z2+})(C_2 + C_{z3+} + C_{y21})R_1R_2R_{z2+}R_{z3+}}{D_{pzy}(s)} + \frac{+s\left((C_1 + C_{y32} + C_{y11} + C_{z2+})R_1R_2R_{z2+} + (C_2 + C_{z3+} + C_{y21})R_1R_2R_{z3+}\right) + R_1R_2}{D_{pzy}(s)} \quad \text{… (27c)}$$

$$\frac{V_{o4}}{V_{in}} = -\frac{1}{R_1R_{z3+}\left(\frac{1}{R_1R_2} + \left(\frac{1}{R_1} + \frac{1}{R_{z2+}} + s(C_1 + C_{y32} + C_{y11} + C_{z2+})\right)\left(\frac{1}{R_{z3+}} + s(C_2 + C_{z3+} + C_{y21})\right)\right)} - \frac{s(C_2 + C_{z3+} + C_{y21})}{R_1\left(\frac{1}{R_1R_2} + \left(\frac{1}{R_1} + \frac{1}{R_{z2+}} + s(C_1 + C_{y32} + C_{y11} + C_{z2+})\right)\left(\frac{1}{R_{z3+}} + s(C_2 + C_{z3+} + C_{y21})\right)\right)} \quad \text{… (27d)}$$

$$\frac{V_{o5}}{V_{in}} = -\frac{R_{z2+}R_{z3+}}{D_{pzy}(s)} \quad \text{… (27e)}$$

Here, $D_{pzy}(s)$ is calculated as

$$\begin{aligned} D_{pzy}(s) = {} & s^2(C_1 + C_{y32} + C_{y11} + C_{z2+})(C_2 + C_{z3+} + C_{y21})R_1R_2R_{z2+}R_{z3+} \\ & + s\left((C_1 + C_{y32} + C_{y11} + C_{z2+})R_1R_2R_{z2+} + (C_2 + C_{z3+} + C_{y21})R_2(R_1 + R_{z2+})R_{z3+}\right) \\ & + R_1R_2 + R_2R_{z2+} + R_{z2+}R_{z3+} \end{aligned} \quad \text{… (28)}$$

Results and Discussion

Simulations of the proposed universal filter are achieved by SPICE program using 0.13 μm 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 $R_1 = R_2 = 1$ kΩ to obtain the SIMO VM second-order universal filter

Table 1 — 0.13 µm IBM CMOS technology parameters

```
.MODEL CMOSN NMOS ( LEVEL = 7
+VERSION = 3.1 TNOM = 27 TOX = 3.2E-9
+XJ = 1E-7 NCH = 2.3549E17 VTH0 = 0.0408721
+K1 = 0.325863 K2 = -0.0303381 K3 = 1E-3
+K3B = 7.9752313 W0 = 1.005139E-7 NLX = 9.892661E-7
+DVT0W = 0 DVT1W = 0 DVT2W = 0
+DVT0 = 1.2297627 DVT1 = 0.1473877 DVT2 = 0.295815
+U0 = 451.7567843 UA = -1.42062E-10 UB = 3.125058E-18
+UC = 4.349531E-10 VSAT = 1.104974E5 A0 = 0.1756127
+AGS = 0.0121649 B0 = 5.453993E-6 B1 = 5E-6
+KETA = 0.05 A1 = 4.699783E-4 A2 = 0.476527
+RDSW = 150 PRWG = 0.3491049 PRWB = 0.1116032
+WR = 1 WINT = 1.273353E-8 LINT = 1.040852E-8
+DWG = -2.333272E-9 DWB = 2.870557E-8 VOFF = -5.88255E-3
+NFACTOR = 2.5 CIT = 0 CDSC = 2.4E-4
+CDSCD = 0 CDSCB = 0 ETA0 = 2.748809E-6
+ETAB = -0.0153583 DSUB = 4.054516E-6 PCLM = 1.9787164
+PDIBLC1 = 0.9653375 PDIBLC2 = 0.01 PDIBLCB = 0.1
+DROUT = 0.9990938 PSCBE1 = 7.952366E10 PSCBE2 = 5.012991E-10
+PVAG = 0.5350786 DELTA = 0.01 RSH = 6.8
+MOBMOD = 1 PRT = 0 UTE = -1.5
+KT1 = -0.11 KT1L = 0 KT2 = 0.022
+UA1 = 4.31E-9 UB1 = -7.61E-18 UC1 = -5.6E-11
+AT = 3.3E4 WL = 0 WLN = 1
+WW = 0 WWN = 1 WWL = 0
+LL = 0 LLN = 1 LW = 0
+LWN = 1 LWL = 0 CAPMOD = 2
+XPART = 0.5 CGDO = 4E-10 CGSO = 4E-10
+CGBO = 1E-12 CJ = 8.406526E-4 PB = 0.8
+MJ = 0.4923081 CJSW = 1.939781E-10 PBSW = 0.99
+MJSW = 0.2751883 CJSWG = 3.3E-10 PBSWG = 0.99
+MJSWG = 0.2751883 CF = 0 PVTH0 = -1.031224E-3
+PRDSW = 0 PK2 = 1.629017E-3 WKETA = 0.0106762
+LKETA = 8.760864E-3 PU0 = -3.5021185 PUA = -3.13657E-11
+PUB = 0 PVSAT = 653.2294237 PETA0 = 1E-4 PKETA = -0.0140591)
.MODEL CMOSP PMOS ( LEVEL = 7
+VERSION = 3.1 TNOM = 27 TOX = 3.2E-9
+XJ = 1E-7 NCH = 4.1589E17 VTH0 = -0.2178731
+K1 = 0.3055794 K2 = -1.881877E-4 K3 = 0.0955725
+K3B = 6.5385817 W0 = 1E-6 NLX = 3.118875E-7
+DVT0W = 0 DVT1W = 0 DVT2W = 0
+DVT0 = 0.2602151 DVT1 = 0.1593124 DVT2 = 0.1
+U0 = 100 UA = 1.043597E-9 UB = 1E-21
+UC = -4.36034E-11 VSAT = 2E5 A0 = 1.844554
+AGS = 0.2915063 B0 = -4.189558E-6 B1 = 5E-6
+KETA = 0.0414839 A1 = 0.0228958 A2 = 1
+RDSW = 105.3697072 PRWG = -0.1019642 PRWB = 0.5
+WR = 1 WINT = 0 LINT = 9.95995E-9
+DWG = 1.093168E-9 DWB = -2.857077E-8 VOFF = -0.1022829
+NFACTOR = 1.5332272 CIT = 0 CDSC = 2.4E-4
+CDSCD = 0 CDSCB = 0 ETA0 = 0.011015
+ETAB = -0.0285373 DSUB = 2.460721E-3 PCLM = 1.6249923
+PDIBLC1 = 0 PDIBLC2 = -4.302895E-9 PDIBLCB = -1E-3
+DROUT = 1.282078E-3 PSCBE1 = 2.169291E9 PSCBE2 = 6.594654E-10
+PVAG = 1.5395235 DELTA = 0.01 RSH = 6.5
+MOBMOD = 1 PRT = 0 UTE = -1.5
+KT1 = -0.11 KT1L = 0 KT2 = 0.022
+UA1 = 4.31E-9 UB1 = -7.61E-18 UC1 = -5.6E-11
+AT = 3.3E4 WL = 0 WLN = 1
+WW = 0 WWN = 1 WWL = 0
+LL = 0 LLN = 1 LW = 0
+LWN = 1 LWL = 0 CAPMOD = 2
+XPART = 0.5 CGDO = 3E-10 CGSO = 3E-10
+CGBO = 1E-12 CJ = 1.174275E-3 PB = 0.8310047
+MJ = 0.4126286 CJSW = 1.312194E-10 PBSW = 0.99
+MJSW = 0.1 CJSWG = 4.22E-10 PBSWG = 0.99
+MJSWG = 0.1 CF = 0 PVTH0 = 5.166851E-4
+PRDSW = 42.1520552 PK2 = 1.857124E-3 WKETA = 0.0358202
+LKETA = 0.0271244 PU0 = -1.0381257 PUA = -4.75151E-11
+PUB = 4.084847E-22 PVSAT = -50 PETA0 = -2E-4 PKETA = -3.142785E-3)
```

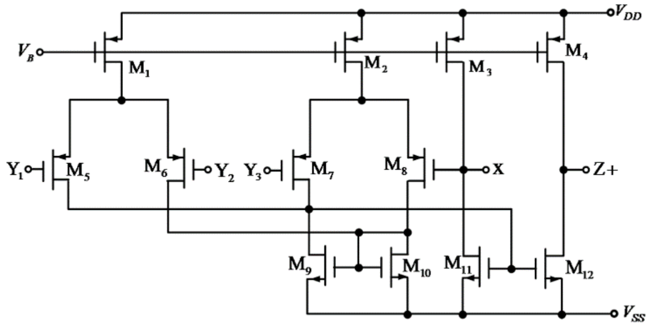

Fig. 5 — Internal structure of the DDCC+ derived from one in Ref.[1]

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

PMOS Transistors	W(µm)/L(µm)
M_1 - M_8	39/1.04
NMOS Transistors	W(µm)/L(µm)
M_9 - M_{12}	13/1.04

responses with Q=1 at resonance frequency $f_o \cong 3.18$ MHz. Power dissipation of the proposed biquadratic universal filter in Fig. 2 is found as 2.62 mW. Non-ideal gains, α, β_1, β_2 and β_3 at sufficiently low frequencies are respectively found as 0.997, 0.999, 0.997 and 0.999 in SPICE simulations. Furthermore, $R_x \cong 55.6\ \Omega$, $R_{z+} \cong 23$ kΩ, $C_{z+} \cong 11.85$ fF, $C_{y1} \cong 193.6$ fF, $C_{y2} \cong 191.2$ fF and $C_{y3} \cong 44.4$ fF are evaluated in SPICE simulations.

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 $C_1 = C_2 = 50$ pf and $f_o \cong 3.18$ MHz are chosen. Thus, resistors are chosen as $R_1 = 0.5$

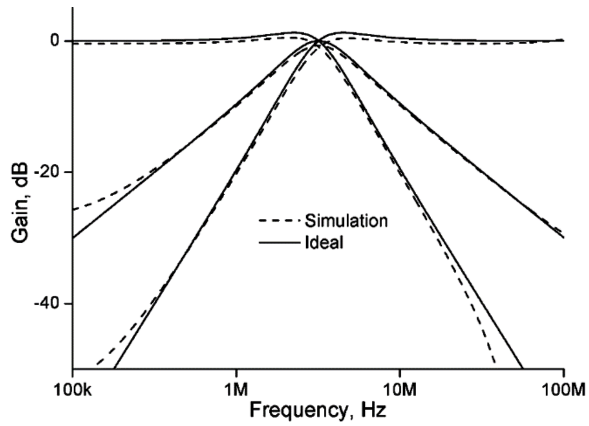

Fig. 6 — Simulation and ideal low-pass, band-pass and high-pass responses versus frequency

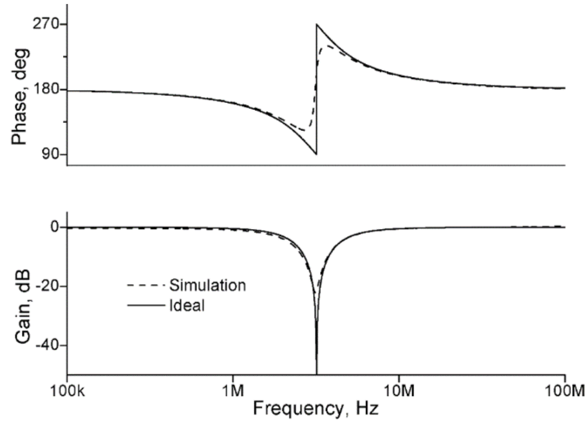

Fig. 7 — Simulation and ideal phase and gain responses for the notch filter

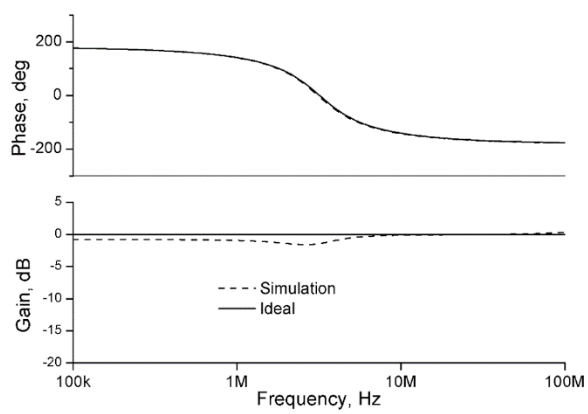

Fig. 8 — Simulation and ideal phase and gain responses for the all-pass filter

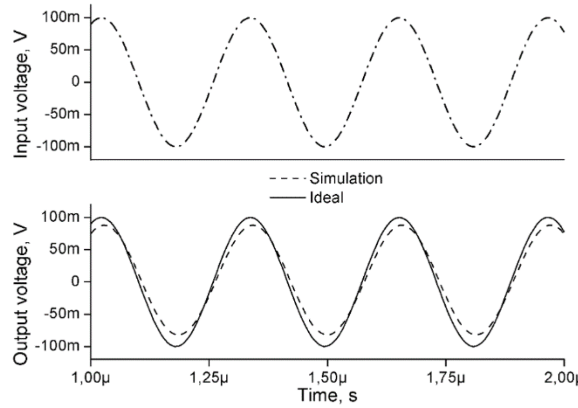

Fig. 9 — Simulation and ideal time domain responses for the all-pass filter

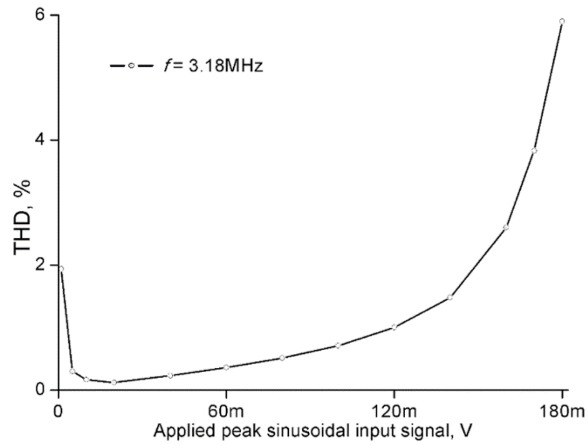

Fig. 10 — THD variations versus applied peak sinusoidal input voltage signal

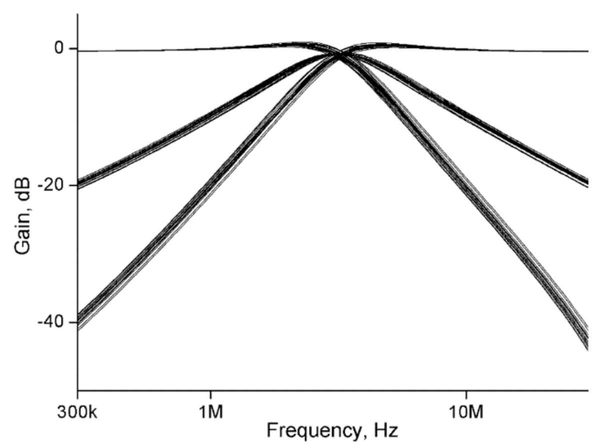

Fig. 11 — A Monte Carlo analysis for the proposed filter in Fig. 2

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 $f_o \cong 3.18$ 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[1-25] and the proposed one

References	# of DDCC(s)	# of Resistor(s)		# of Capacitor(s)		Universal	High Input Impedance	Matching Condition	Power Supply Voltages	Technology	Power dissipation (mW)
		Floating	Grounded	Floating	Grounded						
[1]	2	0	3	0	2	No	Yes	No	±2.5V	CD4007	NA
[2]	3	0	2	0	2	Yes	Yes	No	±1.25V	0.18 µm	NA
[3] in Fig. 2	1	2	0	0	2	No	Yes	No	±2.5V	0.5 µm	NA
[4]	2	2	1	0	2	Yes	No	Yes	±1.25V	0.25 µm	NA
[5]	3	0	2	0	2	Yes	Yes	No	±1.65V	0.35 µm	NA
[6]	3	1	2	0	2	Yes	Yes	No	±1.25V	0.25 µm	NA
[7] in Fig. 1	2 [a]	2	1	0	2	Yes	No	Yes	±1.25V	0.25 µm	NA
[8]	3	1	1	0	2	Yes	Yes	No	±1.25V	0.18 µm	NA
[9] in Fig. 1	1 [a]	2	1	0	2	No	No	No	±1.25V	0.25 µm	NA
[10]	3 [a]	0	3	0	2	Yes	Yes	Yes	±1.25V	0.25 µm	NA
[11]	2 [a]	1	2	0	2	Yes	No	No	±1.25V	0.25 µm	NA
[12]	2 [a]	3	0	0	2	Yes	No	Yes	±1.25V	0.25 µm	NA
[13] in Fig. 1	3 [a]	1	3	0	2	Yes	Yes	Yes	±1.25V	0.18 µm	4.266
[14]	1 [a]	2	1	0	2	No	No	Yes	±1.25V	0.18 µm	NA
[15]	3 [a]	1	3	0	2	Yes	Yes	Yes	±1.25V	0.18 µm	NA
[16]	2 [a]	0	2	0	2	No	Yes	No	±3.3V	1.2 µm	NA
[17]	2 [a]	2	0	0	2	No	No	No	±1.65V	0.35 µm	30.95
[18]	2	1	1	0	2	No	Yes	Yes	±2V	0.5 µm	NA
[19]	2	0	2	0	2	No	Yes	No	±1.65V	0.35 µm	0.52
[20]	2 [a]	1	1	0	2	Yes	No	No	±3.3V	0.5 µm	NA
[21] in Fig. 2	3	0	2	0	2	Yes	Yes	No	±0.5V	0.18 µm	0.0297
[22]	1 [b]	2	4	0	2	Yes	No	No	±0.9V	0.18 µm	NA
[23]	3 [a]	0	5	0	2	No	Yes	No	±2.5V	0.5 µm	NA
[24]	3 [a]	0	2	0	2	No	Yes	No	NA	0.35 µm	NA
[25]	1 [c]	0	0	0	2	No	Yes	No	±1.65V	0.35 µm	83
Present work	3	0	2	0	2	Yes	Yes	No	±0.75V	0.13 µm	2.62

NA: not available [a]do not use DDCC with single Z+ terminal , [b]use extra one FDCCII[c]employ additional two OTAs

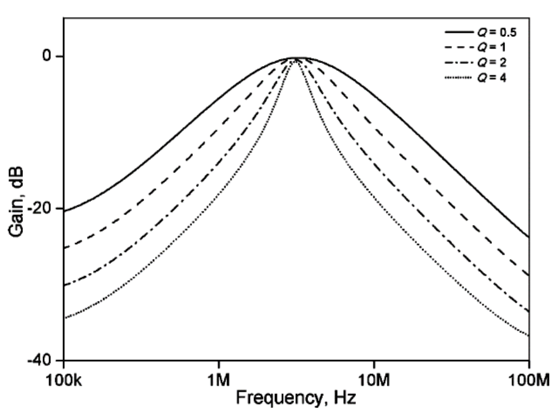

Fig. 12 — Orthogonality of the proposed filter in Fig. 2

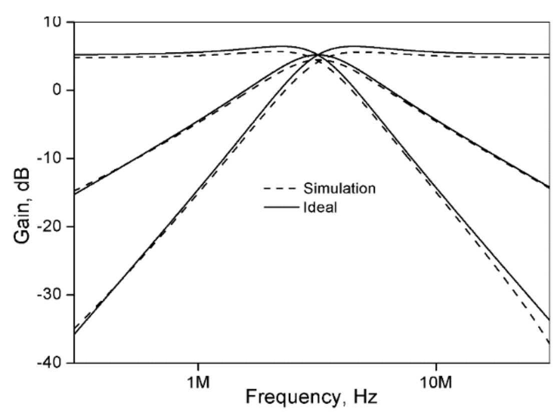

Fig. 13 — Low-pass, band-pass and high-pass responses with gain two

between them arises from parasitic impedances and frequency dependent non-ideal current and voltage gains. On the other hand, the proposed second-order VM universal filter is compared with previously published DDCC based filters[1-25] in Table 3.

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.

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