# Multiple-Input Universal Filter and Quadrature Oscillator Using Multiple-Input Operational Transconductance Amplifiers

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**ABSTRACT** This paper presents a new multiple-input single −output voltage-mode universal biquad filter based on multiple−input operational transconductance amplifiers (MI-OTA). This work demonstrates that the multiple-input OTA−based universal filter can provide more filtering responses and other benefits, compared to conventional OTA−based one. The filter provides electronic and orthogonal control of the natural frequency and the quality factor. Furthermore, a two-phase quadrature oscillator can be obtained by slightly modifying the proposed universal filter while the condition and frequency of oscillation can be controlled orthogonally and electronically. The performance of the proposed circuit is evaluated in Cadence environment using the TSMC 0.18 $\mu$m CMOS technology. The voltage supply is 1.2 V and the power dissipation of the MI−OTA is 24 $\mu$W. For 1% third intermodulation distortion (IMD3) the dynamic range of the band−pass filter is 78.6 dB. In addition, the proposed filter and oscillator are investigated through experiment tests using LM13700 commercially available OTA.

**INDEX TERMS** Universal filter, quadrature oscillator, operational transconductance amplifier, analog signal processing, voltage−mode circuit.

## I. INTRODUCTION

Analog filters are important signal processing blocks for electronic, communication and control systems applications. For instance, they are used to reject the out−of−band noise in electronic and control systems and to eliminate the carrier signal in communication systems [1]. Universal filters are the circuits that usually provide five standard filtering responses into single topology: low-pass (LP), high−pass (HP), band−pass (BP), band−stop (BS) and all−pass (AP) responses. The filters with orthogonal control of the natural frequency and the quality factor are the most desirable ones. Moreover, the voltage−mode filters with high input impedance and without inverting−type signal inputs are required to avoid additional buffers and inverting amplifiers.

The next important kind of electronic circuits are oscillators. They are used in electronic, telecommunication and control systems to generate waveforms of different shapes, amplitudes and frequencies. Quadrature oscillators are the systems that usually generate two sinusoidal signals with 90° phase shift. They are often used in communication and measurement systems, such as quadrature mixers [1], vector generators, selective voltmeters [2], and many other. The circuits with orthogonal control of the condition of oscillation and the frequency of oscillations are the most desirable ones.

The operational transconductance amplifier (OTA) is the basic active block in OTA−C filter design [3]–[11]. However, more complex filters require relatively large number of OTAs, that increases the silicon area and power dissipation [12]. The multiple-input OTA (MI−OTA) appeared as an attractive alternative technique that reduces the number of single−input OTAs used in filter design [12]–[15]. The multiple−input OTA enables summing and subtracting signals at its inputs. It was claimed that the use of multiple-input OTAs could reduce the number of components, silicon area, and power dissipation by approximately the factor of k, where k is the

The associate editor coordinating the review of this manuscript and approving it for publication was Dušan Grujić.

number of inputs of the OTA [12]. As representative examples, one could provide the third− and seventh−order elliptic and low−pass filters described in [12] and [15]. Nevertheless, the MI−OTAs used in the aforementioned designs were based on parallel connections of differential stages to obtain the multiple-input blocks, resulting not only in increased number of transistors and chip area, but mainly in increased number of current branches, thus leading to a higher power dissipation and a more complex internal structure. A simple structure of MI−OTA, using one differential pair, can be achieved by employing the multiple-input floating−gate transistor (MIFG) [16]. Nevertheless, the MIFG transistor is based on charge conversation making it unpractical with modern CMOS technologies that suffer from gate leakage [17]. Furthermore, the MIFG transistor suffers from a residual charge on its gate, that results in higher voltage offset compared to conventional design.

This paper presents a new MI−OTA based universal filter, with more filtering responses and more versatility, and a quadrature oscillator. Unlike, the aforementioned MI−OTAs, the presented OTA use the multiple-input MOS transistor (MI−MOST) technique that allows simplifying its overall structure and decrease the power dissipation. It is worth noting that the first experimental results of MI-MOST were presented by Khateb et. al in [18]–[20]. Later, this technique has been used in several active blocks and applications [21]–[28].

This paper is organized as follows: Section 2 shows the multiple input OTA. Section 3 presents its applications in the active filter and oscillator circuit. Sections 4 and 5 present the non-ideal analysis and simulation results, respectively. The experimental results are shown in Section 6. Finally, the conclusion is given in Section 7.

## II. MULTIPLE-INPUT OTA

Fig. 1 (a) shows the circuit symbol of OTA. Its ideal characteristic can be described by

$$I_o = g_m (V_{1+} - V_{1-}) \tag{1}$$

where $I_o$ is the output current, $g_m$ is the transconductance gain, $V_{1+}$ and $V_{1-}$ denote, respectively, the voltage of the non−inverting and inverting input terminals.

The symbol of multiple−input OTA is shown in Fig. 1 (b). Its ideal characteristic can be described by

$$I_o = g_m \left( \sum_{i=1}^{n} V_{i+} - \sum_{i=1}^{n} V_{i-} \right) \tag{2}$$

where n is the number of required inputs.

In order to realize the multiple−input OTA, one could use n single-input OTAs with their outputs connected together, as shown in Fig. 1 (c). However, this realization results in increased chip area and power consumption that limits its usefulness in low−voltage and low−power applications. Alternative way to realize the multiple−input OTA is to use one OTA with n parallel-connected differential stages [12]–[14]. However, this method still suffers from increased chip area, and power consumption.
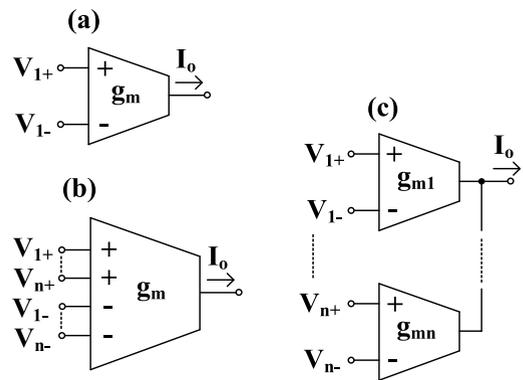


**FIGURE 1.** Circuit symbol of OTA, (a) single−ended OTA, (b) multiple−input OTA and (c) possible realization of multiple−input OTA.
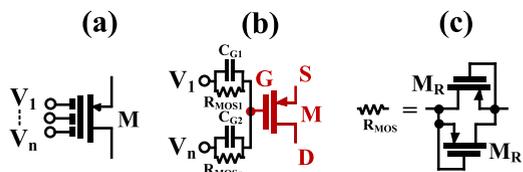


**FIGURE 2.** Multiple−input MOST: (a) symbol, (b) realization and, (c) realization of the large resistance value.
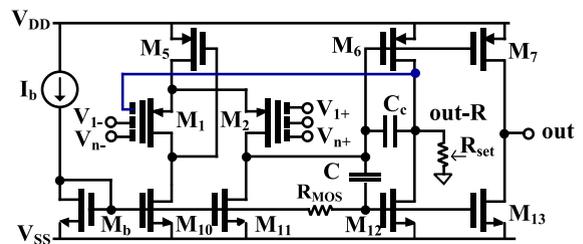


**FIGURE 3.** Multiple−input OTA.

The MI-OTA can be simply realized, using the MI−MOST [18]–[20]. The symbol of the MI−MOST with n inputs is depicted in Fig. 2 (a). The arbitrary number of inputs can be obtained by coupling the input terminals ($V_1,\ldots, V_n$) to the gate terminal (G) of the conventional MOST by n input capacitors ($C_{G1},..,C_{Gn}$). To ensure the DC signal path, the high resistances ($R_{MOS1},..,R_{MOSn}$), created by two MOSTs ($M_R$) operating in cut-off region, are connected in parallel to each input capacitor, as shown in Fig. 2 (b) and (c). Due to the employment of the transistor $M_R$ a high resistance value of the order of several GΩ is simply obtained with minimum chip area. In this work, to realize the MI−OTA, a two-stage OTA with two outputs denoted as out-R and out is used, as shown in Fig. 3. The differential stage consists of one MI−MOST differential pair $M_1$, $M_2$, a flipped voltage follower $M_5$, and two current sources $M_{10}$, $M_{11}$. The second stage operates in so−called super class AB and consists of $M_6$ and $M_{12}$. In this stage, $R_{MOS}$ ensures the DC biasing of the gate of $M_{12}$ while the capacitor C ensures the AC signal path to this gate [28].
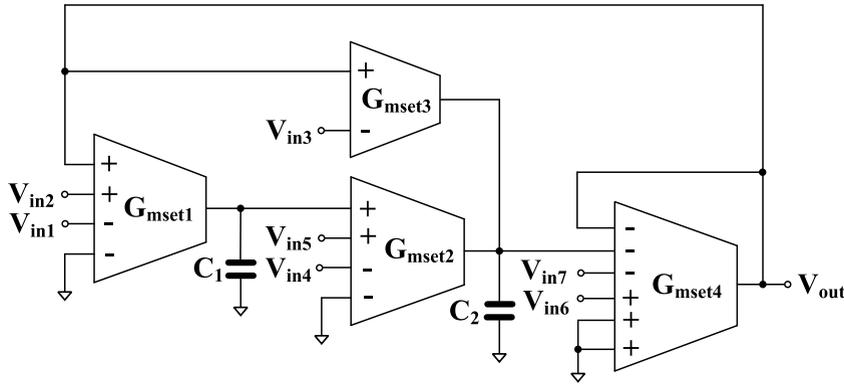
**FIGURE 4.** Proposed universal biquad filter using MI−OTAs.

The capacitor $C_C$ ensures the OTA stability. The negative feedback from the output terminal out-R to the input terminal of $M_1$ ensures transferring of the differential input voltages to the output terminal with the voltage gain equal to unity. The output terminal out−R is connected to a linear adjustable resistor $R_{set}$ that converts the output voltage to output current $I_{Rset}$. This current is mirrored by $M_7$ and $M_{13}$ to the output terminal out. Hence, the transconductance stage is obtained.

$$V_{out-R} = \left( \sum_{i=1}^{n} V_{i+} - \sum_{i=1}^{n} V_{i-} \right) \tag{3}$$

$$I_{out-R} = \frac{V_{out-R}}{R_{set}} = \frac{\left( \sum_{i=1}^{n} V_{i+} - \sum_{i=1}^{n} V_{i-} \right)}{R_{set}} \tag{4}$$

$$G_{meet} = \frac{1}{R_{set}} = \frac{I_{out-R}}{\left( \sum_{i=1}^{n} V_{i+} - \sum_{i=1}^{n} V_{i-} \right)}$$

$$= \frac{I_{out}}{\left( \sum_{i=1}^{n} V_{i+} - \sum_{i=1}^{n} V_{i-} \right)} \tag{5}$$

It is worth mentioning that the proposed MI−OTA enjoys high linearity and increased input voltage range thanks to the negative feedback connection and the input capacitive divider, that attenuate the input signal. Furthermore, thanks to the flipped voltage follower, the minimum voltage supply is given by only one gate-source and one drain-source voltage ($V_{DDmin} = V_{GS−M5} + V_{DS−M10}$).

## III. PROPOSED CIRCUITS

The proposed universal biquad filter using MI−OTAs is shown Fig. 4. It consists of four MI-OTAs and two grounded capacitors. If $V_{in1}$, $V_{in2}$, $V_{in3}$, $V_{in4}$, $V_{in5}$, $V_{in6}$ and $V_{in7}$ are input voltages, the output voltage of the proposed filter can be expressed by

$$V_{out} = \frac{\begin{array}{c} s^2 C_1 C_2 (V_{in6} - V_{in7}) + sC_1 G_{mset2} (V_{in4} - V_{in5}) \\ + sC_1 G_{mset3} V_{in3} + G_{mset2} G_{mset1} (V_{in1} - V_{in2}) \end{array}}{s^2 C_1 C_2 + sC_1 G_{mset3} + G_{mset2} G_{mset1}} \tag{6}$$

where $G_{mset1} = 1/R_{set1}$, $G_{mset2} = 1/R_{set2}$, $G_{mset3} = 1/R_{set3}$ and $G_{mset4} = 1/R_{set4}$.

From (6), five standard filtering responses and ten filtering functions can be obtained as

(1) The non-inverting LP response: $V_{in2} = V_{in3} = V_{in4} = V_{in5} = V_{in6} = V_{in7} = 0$ (grounded), $V_{in1} = V_{in}$.
(2) The inverting LP response: $V_{in1} = V_{in3} = V_{in4} = V_{in5} = V_{in6} = V_{in7} = 0$ (grounded), $V_{in2} = V_{in}$.
(3) The non-inverting BP response: $V_{in1} = V_{in2} = V_{in4} = V_{in5} = V_{in6} = V_{in7} = 0$ (grounded), $V_{in3} = V_{in}$.
(4) The non-inverting BP response: $V_{in1} = V_{in2} = V_{in3} = V_{in5} = V_{in6} = V_{in7} = 0$ (grounded), $V_{in4} = V_{in}$.
(5) The inverting BP response: $V_{in1} = V_{in2} = V_{in3} = V_{in4} = V_{in6} = V_{in7} = 0$ (grounded), $V_{in5} = V_{in}$.
(6) The non-inverting HP response: $V_{in1} = V_{in2} = V_{in3} = V_{in4} = V_{in5} = V_{in7} = 0$ (grounded), $V_{in6} = V_{in}$.
(7) The inverting HP response: $V_{in1} = V_{in2} = V_{in3} = V_{in4} = V_{in5} = V_{in6} = 0$ (grounded), $V_{in7} = V_{in}$.
(8) The non-inverting BS response: $V_{in2} = V_{in3} = V_{in4} = V_{in5} = V_{in7} = 0$ (grounded), $V_{in1} = V_{in6} = V_{in}$.
(9) The inverting BS response: $V_{in1} = V_{in3} = V_{in4} = V_{in5} = V_{in6} = 0$ (grounded), $V_{in2} = V_{in7} = V_{in}$.
(10) The non-inverting AP response: $V_{in2} = V_{in3} = V_{in4} = V_{in7} = 0$ (grounded), $V_{in1} = V_{in5} = V_{in6} = V_{in}$.
(11) The inverting AP response: $V_{in1} = V_{in3} = V_{in5} = V_{in6} = 0$ (grounded), $V_{in2} = V_{in4} = V_{in7} = V_{in}$.

The parameters $\omega_o$ and $Q$ of all filtering responses can be expressed as

$$\omega_o = \sqrt{\frac{G_{mset2} G_{mset1}}{C_1 C_2}} \tag{7}$$

$$Q = \frac{1}{G_{mset3}} \sqrt{\frac{C_2 G_{mset2} G_{mset1}}{C_1}} \tag{8}$$

Letting $G_{mset1} = G_{mset2} = G_{mset}$, (7) and (8) can be rewritten as

$$\omega_o = \frac{G_{mset}}{\sqrt{C_1 C_2}} \tag{9}$$

$$Q = \frac{G_{mset}}{G_{mset3}} \sqrt{\frac{C_2}{C_1}} \tag{10}$$

From (9) and (10), the parameter $\omega_o$ for all filtering responses can be controlled electronically through $G_{mset}$ ($G_{mset} = G_{mset1} = G_{mset2}$) with constant $C_1 = C_2$ while the parameter $Q$ can be controlled orthogonally through $G_{mset3}$ with
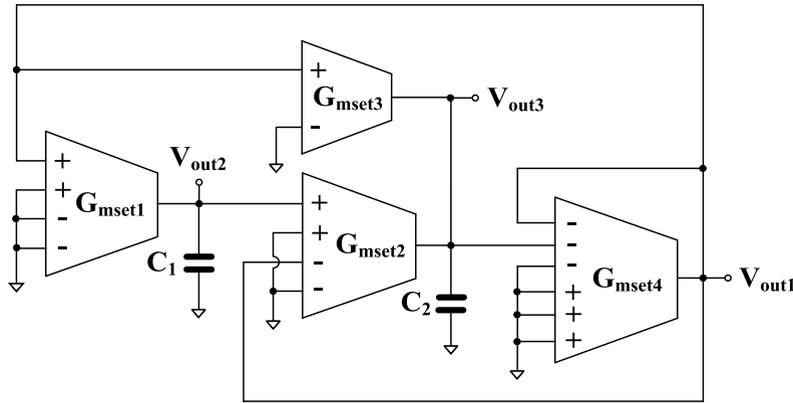
**FIGURE 5.** Quadrature oscillator developed from the universal filter.

constant $G_{mset}$ ($G_{mset} = G_{mset1} = G_{mset2}$,) and $C_1 = C_2$. In case of non−inverting and inverting all-pass responses, a condition $G_{mset2} = G_{mset3}$ is required.

The universal biquad filter in Fig. 4 can be transformed to a quadrature oscillator as shown in Fig. 5. To obtain orthogonal control of the condition of oscillation and the frequency of oscillations, the non-inverting band-pass response $V_{in4}$ could be used to create a positive feedback loop, with inputs $V_{in1}$, $V_{in2}$, $V_{in3}$, $V_{in5}$, $V_{in6}$ and $V_{in7}$ connected to ground. The characteristic equation of the quadrature oscillator can be expressed by

$$s^2 C_1 C_2 + sC_1 (G_{mset3} - G_{mset2}) + G_{mset1}G_{mset2} = 0 \quad (11)$$

The condition of oscillations (CO) and the frequency of oscillations (FO) can be expressed, respectively, by

$$G_{mset2} = G_{mset3} \quad (12)$$

$$\omega_o = \sqrt{\frac{G_{mset1}G_{mset2}}{C_1 C_2}} \quad (13)$$

From (12) and (13), it is evident that CO can be controlled electronically by $G_{mset3}$, while FO can be varied orthogonally by $G_{mset1}$ with $C_1 = C_2$ constant. Thus, the quadrature oscillator can be orthogonally controlled. It should be noted that the quadrature oscillator in Fig. 5 provides three output terminals $V_{out1}$, $V_{out2}$ and $V_{out3}$. The additional outputs still provide sinusoidal signals, with 90º phase shift, and the relationships between output signals can be expressed as:

$$\frac{V_{out2}}{V_{out1}} = \frac{G_{mset1}}{sC_1} \quad (14)$$

$$\frac{V_{out3}}{V_{out2}} = \frac{G_{mset2}}{sC_2} \quad (15)$$

With $s = j\omega_o$, eqns. (14) and (15) can be rewritten respectively, as $V_{out1} = j(\omega_o C_1/G_{mset1})V_{out2}$ and $V_{out2} = j(\omega_o C_2/G_{mset2})V_{out3}$ which indicates that $V_{out1}$ with $V_{out2}$, and $V_{out2}$ with $V_{out3}$ remain in quadrature.
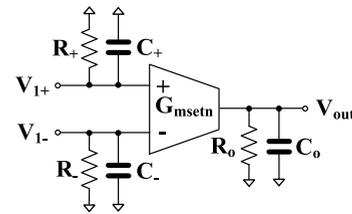


**FIGURE 6.** Modeling the non-idealities in the OTA.

## IV. NON-IDEAL ANALYSIS

Considering non-idealities of OTA, the transconductance gain $G_{msetnj}$ can be expressed as [29]

$$G_{msetnj}(s) = \frac{G_{msetj}\omega_{gj}}{s + \omega_{gj}} \quad (16)$$

where $\omega_{gi}$ denote the pole frequency of OTA$_j$ ($j = 1, 2,.., n$). In the frequency range of a few MHz, $G_{msetnj}$ can be modified as

$$G_{msetnj}(s) \cong G_{msetj}(1 - \mu_j s) \quad (17)$$

where $\mu_j = 1/\omega_{gj}$. The pole frequency $\omega_{gj}$, results from the parasitic input and output resistances (R$_+$, R$_-$, R$_o$) and the input and output capacitances (C$_+$, C$_-$, C$_o$), as shown in Fig. 6. The high-resistance and small-capacitance values will result in high value of $\omega_{gj}$ and small value of $\mu_j$.

Using (17), the denominator of the transfer function of the universal filter can be expressed as

$$s^2 C_1 C_2 \left(1 - \frac{C_1 G_{mset3}\mu_3 - G_{mset1}G_{mset2}\mu_1\mu_2}{C_1 C_2}\right)$$
$$+ sC_1 G_{mset3} \left(1 - \frac{G_{mset1}G_{mset2}\mu_1 + G_{mset1}G_{mset2}\mu_2}{C_1 G_{mset3}}\right)$$
$$+ G_{mset2}G_{mset1} \quad (18)$$

It can be made negligible by satisfying the following, condition:

$$\frac{C_1 G_{mset3}\mu_3 - G_{mset1}G_{mset2}\mu_1\mu_2}{C_1 C_2} \ll 1 \quad (19)$$

$$\frac{G_{mset1}G_{mset2}\mu_1 + G_{mset1}G_{mset2}\mu_2}{C_1 G_{mset3}} \ll 1 \quad (20)$$

**TABLE 1.** Parameters of the components of MI-OTA in Fig. 2.

| Transistor | W/L (μm/ μm) |
| --- | --- |
| $M_1$, $M_2$, $M_5$ | 9×9/0.3 |
| $M_b$, $M_{10}$, $M_{11}$ | 12/3 |
| $M_{12}$, $M_{13}$ | 2×12/3 |
| $M_6$, $M_7$ | 2×25/2 |
| $M_R$ | 4/5 |
| $C_G$=0.5 pF, $C_c$= C = 2.6pF | |

The various passive and active sensitivities of the parameters $\omega_o$ and $Q$ of the universal filter can be expressed as

$$S_{G_{mset1}}^{\omega_o} = S_{G_{mset2}}^{\omega_o} = -S_{C_1}^{\omega_o} = -S_{C_2}^{\omega_o} = \frac{1}{2} \quad (21)$$

$$S_{G_{mset1}}^{Q} = S_{G_{mset2}}^{Q} = S_{C_2}^{Q} = -S_{C_1}^{Q} = \frac{1}{2} \quad (22)$$

$$S_{G_{mset3}}^{Q} = -1 \quad (23)$$

Thus, all the incremental parametric sensitivities for parameters $\omega_o$ and $Q$ are below 1.

Using (17), the characteristic equation of a quadrature oscillator can be written as

$$s^2 C_1 C_2 \left( 1 \right.$$
$$\left. - \frac{C_1 G_{mset3}\mu_3 - C_1 G_{mset2}\mu_2 - G_{mset1}G_{mset2}\mu_1\mu_2}{C_1 C_2} \right)$$
$$+ sC_1 \left( G_{mset3} - G_{mset2} \right)$$
$$\times \left( 1 - \frac{G_{mset1}G_{mset2}\mu_2 - G_{mset1}G_{mset2}\mu_1}{C_1 \left( G_{mset3} - G_{mset2} \right)} \right)$$
$$+ G_{mset1}G_{mset2} = 0 \quad (24)$$

It can be made negligible by satisfying the following condition:

$$\frac{C_1 G_{mset3}\mu_3 - C_1 G_{mset2}\mu_2 - G_{mset1}G_{mset2}\mu_1\mu_2}{C_1 C_2} \ll 1 \quad (25)$$

$$\frac{G_{mset1}G_{mset2}\mu_2 - G_{mset1}G_{mset2}\mu_1}{C_1 \left( G_{mset3} - G_{mset2} \right)} \ll 1 \quad (26)$$

## V. SIMULATION RESULTS

The circuit was designed to work with 1.2 V supply, with 5 $\mu$A bias current and 24 $\mu$W power consumption. The Cadence environment has been used to design and simulate the circuit using a 0.18$\mu$m CMOS technology from TSMC. The parameters of the components are shown in Table. 1.

For the filter design the capacitor $C_1 = C_2 = 10$ nF and $R_{set1} = R_{set2} = R_{set3} = R_{set4} = 15$ k$\Omega$ were selected for natural frequency of 1 kHz. These $R_{set}$ resistors can be integrated on chip using a high resistance poly resistor while these 10 nF capacitors should be off-chip capacitors.

The simulated magnitude frequency responses of the universal filter showing the non-inverting LP, HP, BP, and BS responses are shown in Fig. 7. The simulated natural frequency is 1.03 kHz. The magnitude frequency response and the phase characteristic of the non-inverting AP filter are shown in Fig. 8. The total power consumption of the filter is 96 $\mu$W. Fig. 9 shows the tuning capability of the $Q$ factor
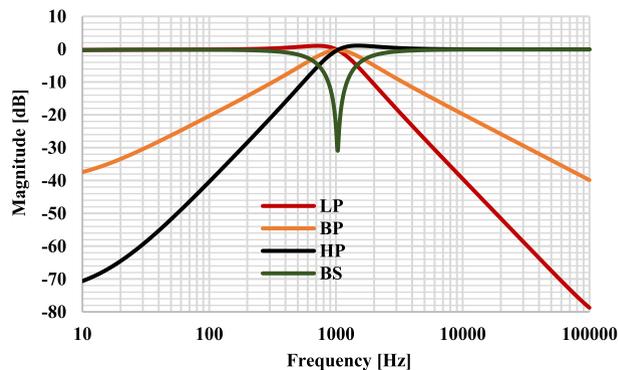
**FIGURE 7.** The simulated magnitude frequency responses of the universal filter.
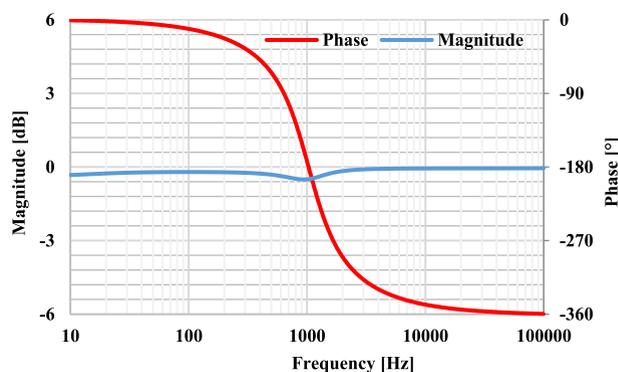
**FIGURE 8.** The simulated magnitude frequency response and phase characteristic of the AP filter.
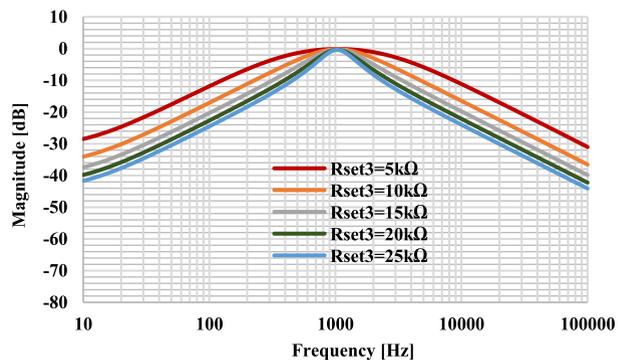
**FIGURE 9.** The simulated tuning capability of the BP filter.

for the BP filter by tuning $R_{set3} = 5$k$\Omega$, 10k$\Omega$, 15k$\Omega$, 20k$\Omega$ and 25k$\Omega$, while $R_{set1} = R_{set2} = R_{set4} = 15$k$\Omega$.

A sine wave with different amplitudes (0.05V, 0.1V, 0.15V, 0.2V, 0.25V) and 1 kHz frequency was applied to the input of the BP filter. The output signals are shown in Fig. 10. The total harmonic distortion (THD) is less than 1.67 % for input amplitude of 0.3V as shown in Fig. 11. This confirms the high linearity of the filter with low THD.

To determine the third-order distortion products produced by the circuit nonlinearity, two tones that are close in frequency are applied to the input of the BP filter. The first
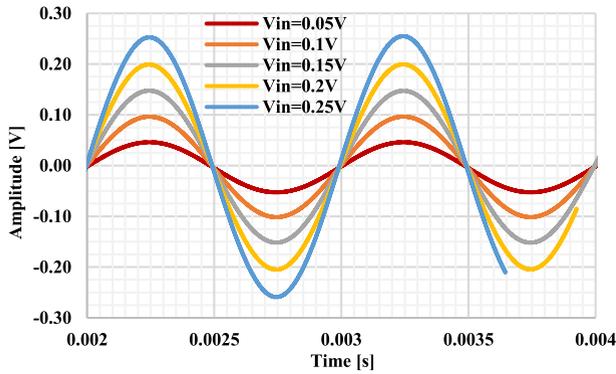
**FIGURE 10.** The simulated transient analysis of the BP filter with different amplitude of input signal at 1kHz.
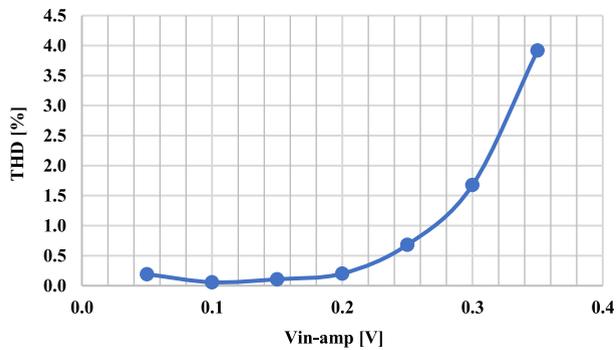


**FIGURE 11.** The simulated transient analysis of the BP filter with different amplitude of input signal at 1kHz.
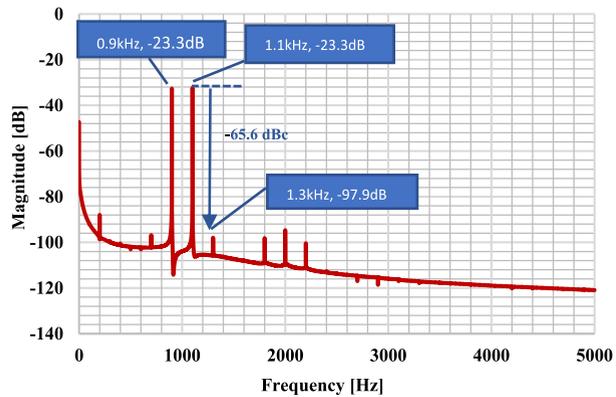


**FIGURE 12.** The simulated spectrum of the output signal of the BP filter.



**FIGURE 13.** The simulated IMD3 versus the input signal ($V_{in-pp}$) for the BP filter.
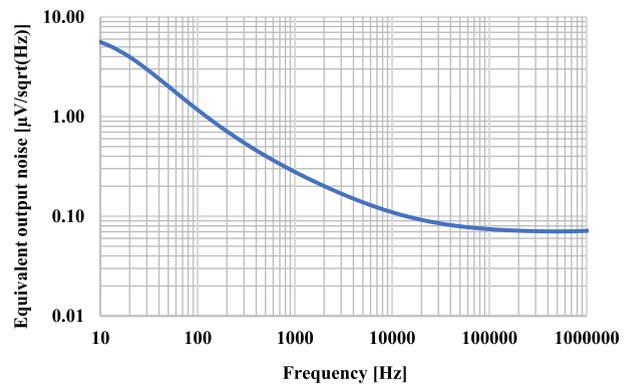


**FIGURE 14.** The simulated equivalent output noise of the BPF.
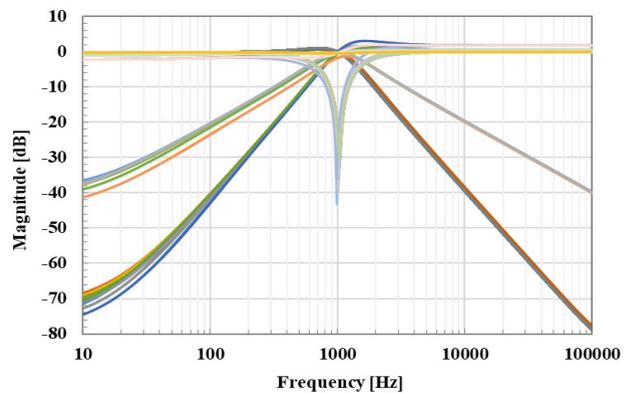


**FIGURE 15.** The simulated magnitude frequency responses of the universal filter with PVT corners.

tone is a sine wave with amplitude of 25mV@ 0.9kHz and the second one with 25mV@ 1.1kHz. The spectrum of the output signal is depicted in Fig. 12. The third intermodulation distortion (IMD3) is only -65.6 dBc. The relation between the IMD3 and the peak-to-peak value of the input signal is shown in Fig. 13. The IMD3 is around -33.7 dB for 350mV$_{pp}$. The equivalent output noise of the BPF is shown in Fig. 14. The RMS output noise of the BP filter integrated in the band pass from 432Hz−2.523kHz was 12$\mu$V that results in 78.6dB dynamic range for 1% IMD3.

To confirm the robustness of the design under process, voltage, temperature (PVT) corners, the MOS transistor (slow−slow, slow −fast, fast−slow, fast−fast), MIM-capacitor (slow−slow, fast−fast), voltage supply (V$_{DD}$±10%), and temperature (-10°C−60°C) corners were selected. Fig. 15 shows the magnitude frequency responses for the circuits of Figs. 7 and 8 with PVT corners and confirm the robustness of the design.
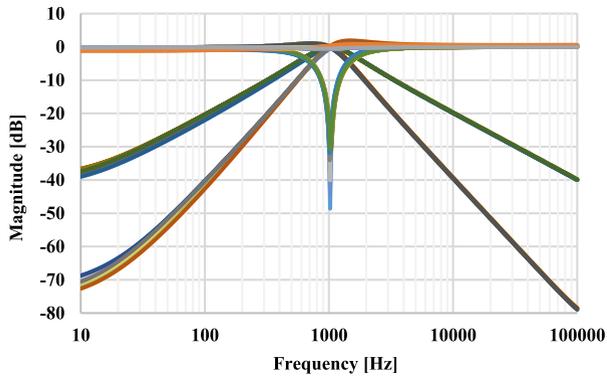
**FIGURE 16.** The simulated magnitude frequency responses of the universal filter with 200 runs MC analysis.



**FIGURE 17.** The simulated growing oscillations of the quadrature oscillator output voltages.



**FIGURE 18.** The simulated steady−state output waveforms of the oscillation.



**FIGURE 19.** The simulated spectrums of the output signals.



**FIGURE 20.** The simulated oscillator frequency versus $R_{set1}$.
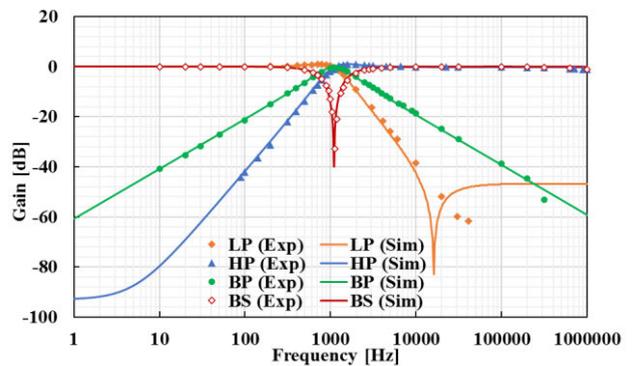


**FIGURE 21.** The measured and simulated frequency magnitude responses of the LP, BP, HP and BS filters.

The low sensitivity of the design to process and mismatch variation was also proved by Monte Carlo (MC) analysis with 200 runs as shown in Fig. 16.

The quadrature oscillator was also simulated with $C_1 = C_2 = 10nF$, $R_{set1} = R_{set2} = R_{set4} = 15k\Omega$. To start the oscillation the $R_{set3}$ was selected to $16k\Omega$. The simulated oscillation frequency was 1.04kHz. Fig. 17 shows the growing oscillations of the oscillator output voltage $V_{out1}$, $V_{out2}$ and $V_{out3}$ while Fig. 18 shows the steady-state waveforms. The spectrum of the output signals was shown in Fig. 19, the THD of the output signals is 0.33%, 0.35% and 0.1% for $V_{out1}$, $V_{out2}$, $V_{out3}$, respectively.

Fig. 20 shows the frequency versus $R_{set1}$. For all simulated frequencies, the THD was less than 0.4%.

## VI. EXPERIMENTAL RESULTS

The proposed universal filter was also tested experimentally. The prototype circuit was realized using commercially available integrated circuits LM13700N [30]. Note, that the macro-model of the LM13700N has been also used, hence the simulation results based on macro-model and the measured results can be compared. The multiple input OTA was realized by a parallel connection of OTAs (LM13700N) as shown in Fig. 1 (c).
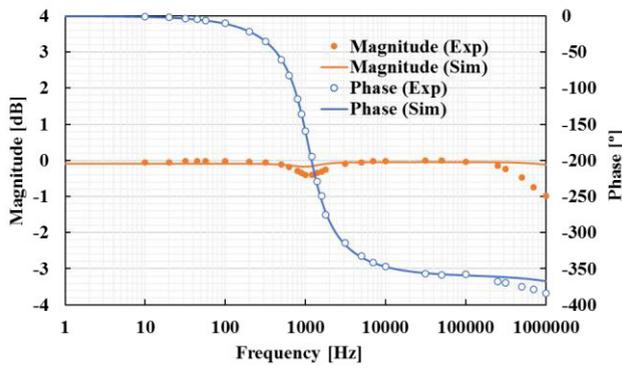
**FIGURE 22.** The measured and simulated frequency magnitude and phase responses of the AP filter.
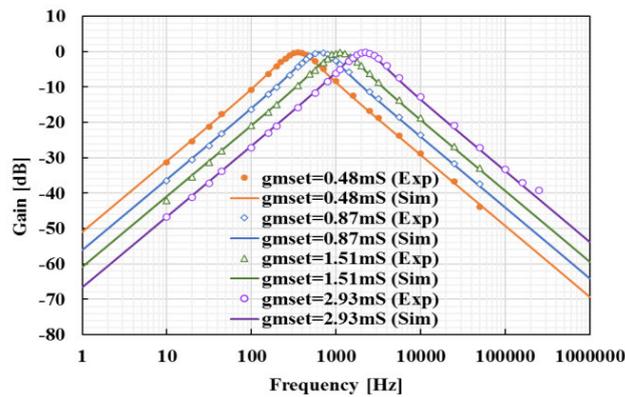


**FIGURE 23.** The measured and simulated magnitude responses of the BP filter with different parameter $\omega_o$.
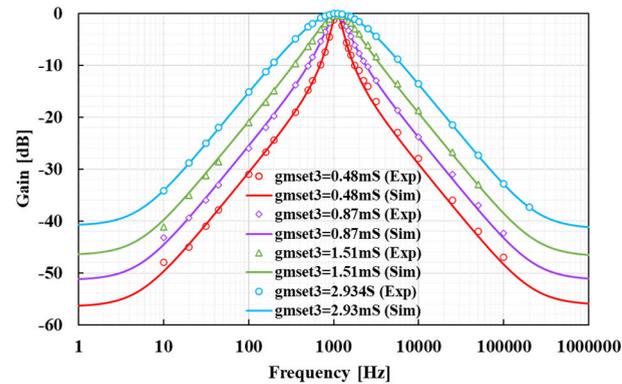


**FIGURE 24.** The measured and simulated magnitude responses of the BP filter with different parameter Q.



**FIGURE 25.** The experimental result of the quadrature outputs $V_{out1}$, $V_{out2}$, $V_{out3}$.
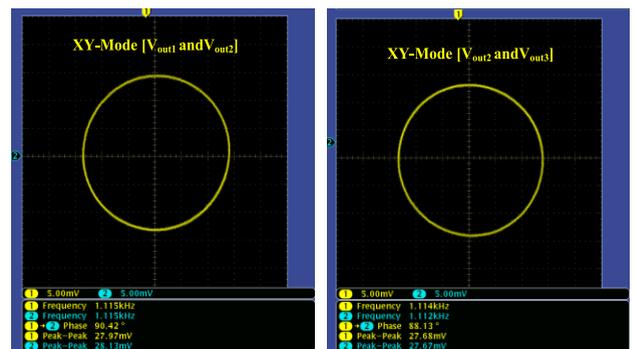


**FIGURE 26.** Lissajous pattern: (a) $V_{out1}$ and $V_{out2}$ outputs, (b) $V_{out2}$ and $V_{out3}$ outputs.
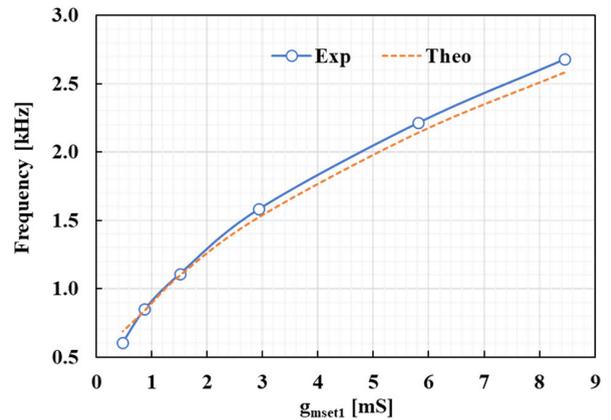


**FIGURE 27.** The measured and experimental frequency of oscillations versus $g_{mset1}$.

The supply voltages were $V_{DD} = -V_{SS} = 5$ V and the capacitances $C_1$ and $C_2$ were of 220 nF. The sinusoidal input signal and the output waveforms were measured using Agilent Technologies DSOX 1102G oscilloscope. The transconductances $g_{mset1} = g_{mset2} = g_{mset3} = g_{mset4} = 1.512$ mS were designed to obtain the filter with the natural frequency $f_o \cong 1.09$ kHz and the quality factor $Q \cong 1$. Fig. 21 shows the measured and simulated magnitude responses of the non-inverting LP, HP, BP and BS filter with natural frequency $f_o = 1.09$ kHz. Fig. 22 shows magnitude and phase responses

of the AP filter. Fig. 23 shows magnitude responses of the BP filters for different values of the transconductance $g_{mset}$ ($g_{mset} = g_{mset1} = g_{mset2} = g_{mset3}$) equal to 0.481 mS, 0.873 mS, 1.512 mS and 2.934 mS.

Fig. 24 shows the magnitude responses of the BP filters for $g_{mset3}$ equal to 0.481 mS, 0.873 mS, 1.512 mS and 2.934 mS.

**TABLE 2.** Properties comparison of this work with those of OTA-based universal filters.

| Factor | [6] | [7] | [8] | [9] | [10] | [11] | Proposed |
|---|---|---|---|---|---|---|---|
| Number of active devices | 3–OTA/1–INV | 5–OTA | 5–OTA | 5–OTA | 5–OTA | 1-DO–OTA, 3–OTA | 4-MI–OTA |
| Realization | CMOS structure | commercial IC | commercial IC | commercial IC | CMOS structure | commercial IC | CMOS structure & commercial IC |
| Number capacitors | 2–C | 2–C | 2–C | 2–C | 2–C | 2–C | 2–C |
| Total capacitance (nF) | 0.002 | 4 | 4.4 | 2 | 0.02736 | 4.4 | 20 |
| Type of filter | MISO | MISO | MIMO | MIMO | MISO | MIMO | MISO |
| Operation mode | MM | VM | VM | VM | MM | VM | VM |
| Number of offered responses | 20 (5–VM) | 5 | 13 | 24 | 20 (5-VM) | 9 | 11 |
| Offer inverting and non-inverting of five standard responses | Yes | No | No | No | No | No | Yes |
| All grounded capacitors | Yes | Yes | Yes | Yes | Yes | Yes | Yes |
| Unnecessary inverting input conditions | Yes | Yes | Yes | Yes | Yes | Yes | Yes |
| High input impedances | Yes | Yes | Yes | Yes | Yes | Yes | Yes |
| Orthogonal control of $\omega_o$ and $Q$ | Yes | Yes | Yes | Yes | Yes | Yes | Yes |
| Offer modified into oscillator | No | Yes | Yes | Yes | No | Yes | Yes |
| Number phases of quadrature output | – | 2 | 2 | 2 | 2 | 2 | 3 |
| Simulated power supply (V) | ±0.5 | ±15 | ±15 | ±15 | ±0.9 | ±15 | 1.2 |
| Natural frequency (kHz) | 2500 | 238.73 | 217 | 159.16 | 3390 | 144.7 | 1 |
| Simulated power dissipation (mW) | 0.035 | - | 860 | 861 | – | 920 | 0.096 |
| Total harmonic distortion (%) | 2@60mV$_{pp}$ | 1.99@65mV$_{pp}$ | 1.93@200mV$_{pp}$ | 2@100mV$_{pp}$ | – | 3.83 @170mV$_{pp}$ | 1.67@600mV$_{pp}$ |
| IMD3 (dBc) | – | -33@40mV$_{pp}$ | -34@120mV$_{pp}$ | -34@60mV$_{pp}$ | – | -48.67@63mV$_{pp}$ | -52.1@200mV$_{pp}$ |
| Dynamic range (dB) | – | – | – | – | – | – | 78.6 |
| Verification of result | Sim | Sim/Exp | Sim/Exp | Sim/Exp | Sim | Sim/Exp | Sim/Exp |

**Note**: MISO = multiple-input single output, MIMO = multiple–input multiple–output, INV = inverter, MM = mixed–mode, VM = voltage–mode, Sim = simulation, Exp = experimental, DO–OTA = dual–output OTA.
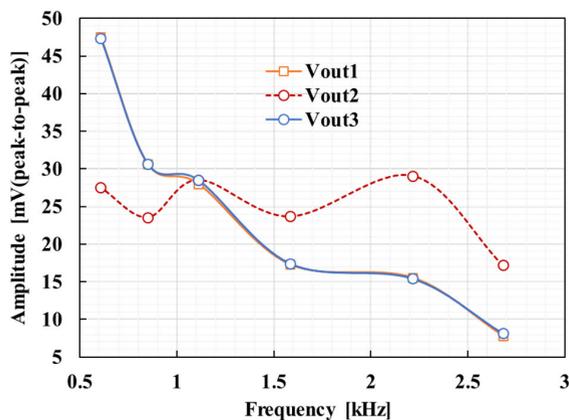


**FIGURE 28.** The measured output amplitudes versus frequency of oscillations.



**FIGURE 29.** The measured THD of oscillations versus frequency.

The value of $g_{mset3}$ was varied to obtain different values of the parameter $Q$, while the input signal was applied to $V_{in3}$, with $g_{mset1} = g_{mset2} = 1.512$ mS.

The proposed quadrature oscillator was experimentally tested as well. The capacitances $C_1$ and $C_2$ were 220nF. The measured output waveforms were taken using Tektronix MSO 4034 mixed signal oscilloscope. Fig. 25 shows the measured output wave forms of $V_{out1}$, $V_{out2}$ and $V_{out3}$ for $g_{mset1} = g_{mset2} = g_{mset4} = 1.512$mS and variable $g_{mset3}$, for controlling the condition of oscillation. The circuit generates the output waveform with frequency of 1.11 kHz, while the theoretical value was 1.09 kHz. The amplitudes were nearly equal in this case. The quadrature output waveform

in Fig. 25 was verified through the XY mode to show the quadrature relationship. The quadrature relationships between $V_{out1}$ and $V_{out2}$ and between $V_{out2}$ and $V_{out3}$ were shown in Fig. 26, (a) and (b), respectively. The measured frequency of oscillations as a function of $g_{mset1}$ is shown in Fig. 27. The achieved results agree well with (13). The frequency of oscillations was 0.607 kHz, 0.849 kHz, 1.11 kHz, 1.58 kHz, 2.21 kHz and 2.68 kHz for $g_{mset1}$ equal to 0.481 mS, 0.873 mS, 1.512 mS, 2.934 mS, 5.81 mS and 8.45 mS, respectively. The plot showing the output amplitude versus frequency of oscillations is shown in Fig. 28. The THD of the output signals $V_{out1}$, $V_{out2}$ and $V_{out3}$ is shown in Fig. 29. The phase error between $V_{out1}$ and $V_{out2}$, and between $V_{out2}$ and $V_{out3}$, referred to 90º, is shown in Fig. 30.
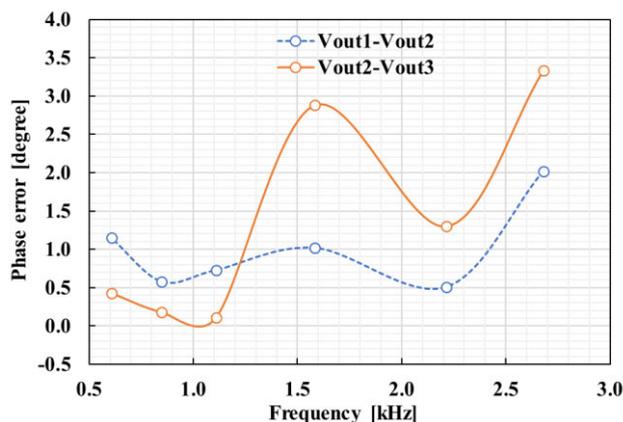
**FIGURE 30.** The measured phase error as a function of frequency of oscillations.

Table. 2. compares the proposed circuit with most recent published works [6]–[11]. While the works in [7]–[9], [11] use only commercially available integrated circuit without CMOS structure, the works in [6], [10] use only a conventional CMOS OTA structure (i.e. well known OTA with one input).

Compared with [6], the proposed filter can be easily transformed into a quadrature oscillator. Although using the topology in [6] one can obtain five standard filtering responses, they are achievable only with the use of different modes of operation (voltage−mode, current−mode, transresistance−mode, transconductance−mode). Compared with [8]–[10], the proposed filter provides full non-inverting and inverting realization of five standard filtering responses and offers a three−phase quadrature oscillator. Furthermore, thanks to innovative and simple CMOS structure, the proposed filter offers improved linearity, lowest THD and IMD3, wide dynamic range and low power consumption and less amounts of active elements when it is realized using MI−OTAs. Finally, it is worth to mention that due to absence of MI-OTA fabrication the commercially available integrated circuit LM13700N is used only to confirm the functionality and benefits of the multiple-input technique on these new presented applications.

## VII. CONCLUSION

In this paper, a new voltage-mode universal biquad filter employing four MI−OTAs and two grounded capacitors is proposed. It offers eleven filtering responses into single topology which can be possible using the MI−OTA−based circuit. The filter also provides orthogonal control of the natural frequency and the quality factor. By slightly modifying the proposed universal filter, a two−phase quadrature oscillator with orthogonal control of the condition and frequency of oscillation can be obtained. The performance of the proposed filter and oscillator circuit were evaluated in Cadence environment using the TSMC 0.18 $\mu$m CMOS technology and investigated by experiment tests using LM13600 discrete

component integrated circuit as OTAs. The simulation results were in consistent with the experimental ones.

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