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0.5 V Fully Differential Universal Filter Based on Multiple Input OTAs

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ABSTRACT The design of a low-power, low-voltage, fully-differential universal biquad filter is presented in this work, which is constructed from four multiple-input gate-driven operational transconductance amplifiers (MI-OTAs) along with one passive resistor and two passive capacitors. The scheme of presented biquad filter has three high-input impedance voltage nodes and single output voltage node. Five unity gain filtering functions, all-pass (AP), low-pass (LP), band-pass (BP), high-pass (HP) and band-stop (BS) responses, are obtained. The selection of output filtering responses is obtained without the need of component matching condition, inverting or double input voltage. With this feature, it can be easily controlled with digital programming. The quality factor (Q) and angular frequency (ω_0) are electronically and independently tuned. Moreover, the adjustment of ω_0 and Q can be done without affecting the voltage gain. A workability of the design is confirmed via Cadence software and the Spectre simulator based on the 180 nm TSMC CMOS technology parameters. The proposed fully differential filter operates with 0.5 V supply voltage. The results verify that the proposed filter dissipates the total power of 53.3 nW. Additionally, the dynamic range (DR) of band-pass filtering function is 63 dB for 2% third intermodulation distortion (IMD). Also, the simulated RMS value of the band-pass filtering noise is 45 μV.

INDEX TERMS MI-OTA, low-power low-voltage circuit, universal filter, analog circuit, electronic control.

I. INTRODUCTION

Analog active filters are necessary in many fields of analog signal processing systems for examples sound systems, control systems, communication systems, instrumentation etc. The designs of second order or biquad filters are receiving a lot of attention, especially the universal biquad filter which provides all five filtering responses, low-pass (LP), band-pass (BP), high-pass (HP), all-pass (AP), and band-stop (BS) functions into single circuit scheme [1]. The universal biquad filter can be used in areas such as in touchtone telephone tone decoder, FM stereo demodulation, and audio crossover network circuit [1]–[3]. The design of analog active filters employing the single-ended active function blocks has been continuously presented in the open literature [4]–[8]. However, the use of single-ended active devices to realize the analog filter limits the dynamic range. With this limitation, the fully differential filters have been introduced to amend the analog active filtering performance. Not only the improvement of dynamic range but also the fully differential filter gives many advantageous features such as improving interference rejection and supply noise reduction or lower harmonic distortion. Moreover, it can reduce the coupling influence between various blocks [9], [10].

In the open literature, various fully differential voltage-mode universal biquad filters are available [10]–[20]. These filters are based on multiple input fully differential difference amplifier (MI-FDDA) [10], fully balanced differential difference amplifier (FBDDA) [11], [12], fully differential difference transconductance amplifier (FDDTA) [13], fully differential difference transconductor (FDDT)[14], fully balanced four-terminal floating nullor (FBFTFN) [15], operational transconductance amplifier (OTA) [16]–[18], second generation fully differential current conveyor (FDCCII)

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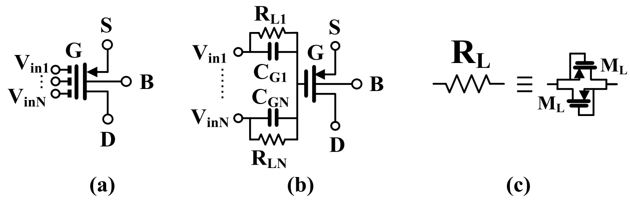

FIGURE 1. Multiple input MOST: a) symbol b) realization using passive elements (C, R) and c) practical realization of the large resistor by two MOSTs.

[19], [20], and digitally controlled fully differential current conveyor (DCFDCC) [21]. However, the filters in [10]–[13], [15], [21] contain large numbers of passive elements which increase the chip area. Some filters [12]–[21] operate at high power supply voltage (more than 0.5V) and are not suitable for low-voltage low-power applications. Several filters [10]–[13], [15], [19], [20] do not have the feature of electronic tuning of $\omega_0$and Q. In [11], the ω_0 and Q are not orthogonally controlled. The filters presented in [11]–[14], [16]–[21] cannot provide five filtering responses, AP, LP, BP, BS and HP functions. High impedance at input voltage nodes of the fully differential universal filters in [10]–[12], [15] is not obtained. Some filters require critical matching conditions [10], [11].

In this paper, a MI-OTAs based low-voltage low-power (LV-LP) fully differential versatile biquad filter is proposed. The designed filter consists of four MI-OTAs, two capacitors and one resistor. The feature of electronic and independent tuning of angular frequency and quality factor is obtained. Also, the tuning of angular frequency and quality factor does not affect the voltage gain. To provide five filtering responses, the proposed filter does not need an extra active device. The simulation results with Cadence software and the Spectre simulator verify the performance of the design.

II. PRINCIPLE OF THE PROPOSED CIRCUIT

A. MULTIPLE-INPUT CMOS OTA

The universal filter proposed in this work is based on multiple-input OTAs (MI-OTAs), that allows simplifying its overall structure and improve the dynamic range (DR). In order to simplify the structure of the MI-OTA itself, a concept of Multiple-Input MOS transistor (MI-MOST) is introduced.

The symbol and practical realization of the MI-MOST are shown in Figs. 1(a) and 1(b) respectively. The MI-MOST can be seen as multiple-gate device, where the multiple-gate is realized using an analog summing circuit composed of capacitors C_{Gi} (i=1…N) and an input capacitance of an "internal" MOS transistor seen from its gate terminal (C_{in}). In order to provide proper biasing of the gate terminal for DC, the capacitors C_{Gi} are shunted with the large resistors R_{Li}. The resistors should be sufficiently large, in order to not deteriorate the input resistance of the transistor. The possible realization of the resistance R_L composed of two MOS transistors operating in the cutoff region is shown in Fig. 1(c).

Assuming $1/\omega C_{Gi} \ll R_{Li}$, the gate potential of the MI-MOST can be expressed as follows:

$$V_G = \sum\nolimits_{i=1}^{N} \frac{C_{Gi}}{C_{\sum}} V_{ini}. \tag{1}$$

where C_{Σ} is the sum of capacitances given as:

$$C_{\sum} = C_{in} + \sum\nolimits_{i=1}^{N} C_{Gi}. \tag{2}$$

Thus, as it is seen from Eqs. (1) and (2), the voltage signal at the internal gate is attenuated by the factor of C_{Σ}/C_{Gi}, being the result of the voltage divider added to the gate terminal of the MOST.

Substituting the gate potential into typical equations describing an MOS transistor we obtain the description of the MI-MOST. For instance, the dynamic large-signal transfer characteristic of the p-channel MI-MOST operating in subthreshold region can be expressed as follows:

$$I_D = I_o \exp\left[\left(V_S - \sum\nolimits_{i=1}^{N} \frac{C_{Gi}}{C_{\sum}} V_{ini}\right) \cdot \frac{1}{n_p U_T}\right]. \tag{3}$$

From Eq. (3) the small-signal transconductance of the MI-MOST from i-th input (with other inputs shorted to ground for AC signal) is given by:

$$g_{mi} = g_m \frac{C_{Gi}}{C_{\sum}}. \tag{4}$$

where $g_m = I_D/n_p U_T$ is the transconductance of internal MOS.

As it can be concluded from the above considerations, the MI-MOST is a multiple-input device, where the transconductance seen from i-th input is attenuated by the factor of C_{Σ}/C_{in}. It is obvious that both, the large-signal range, as well as the input-referred noise (from i-th input) are increased in the same proportion, therefore, the DR remains the same as for an internal MOS.

The intrinsic voltage gain of the MI-MOST device, from i-th input, is given by:

$$A_{oi} = \frac{g_{mi}}{g_{ds}} = \frac{g_m}{g_{ds}} \cdot \frac{C_{Gi}}{C_{\sum}}. \tag{5}$$

thus, it is attenuated by the factor of C_{Σ}/C_{Gi} as well.

The schematic of the MI-OTA based on MI-MOST is shown in Fig. 2. The circuit can be seen as a typical fully-differential current mirror OTA. The input differential pair (M_1, M_2) is realized with MI-MOST transistors. The current mirrors M_3-M_5 and M_6-M_8 are realized using self-cascode connections, which increases the DC output resistance (voltage gain), while still maintaining large output swing. Similar self-cascode connections were used in the upper current mirrors (M_{10}-M_{14}), (M_{11}-M_{13}), however, the upper transistors of the self-cascode connections form a bulk-driven common-mode feedback (CMFB) circuit. The transistors with index "c" operate in triode region, with V_{DS} voltage drops of only 14 mV, therefore, they don't affect the output voltage swing. The output common-mode level is equal to the reference voltage V_{CM}. The principle of operation of the CMFB circuit is described in more detail in [22]. As it can be concluded from

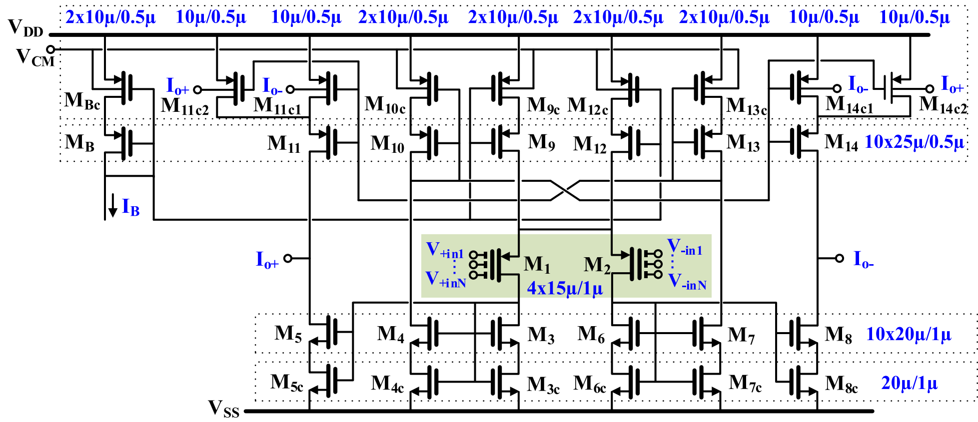

FIGURE 2. CMOS structure of the multiple input OTA with enhanced gain.

the above considerations, the MI-MOST is a multiple-input device, where the transconductance seen from i-th input is attenuated by the factor of C_Σ/C_{in}. It is obvious that both, the large-signal range, as well as the input-referred noise (from i-th input) are increased in the same proportion, therefore, the DR remains the same as for an internal MOS.

The MI-OTA operates with nanoampere biasing currents ($I_B = 4.3$ nA), therefore, all transistors operate in subthreshold region. Neglecting the second-order effects and assuming unity gain of all current mirrors, the small-signal output current of the OTA can be expressed as:

$$I_{o+} - I_{o-} = \sum_{i=1}^{N} g_{mi} \left(V_{+ini} - V_{-ini}\right), \tag{6}$$

where g_{mi} is the small-signal transconductance of the MI-MOSTs M_1 and M_2, described by (4), where $I_D = I_B$.

The DC voltage gain, from i-th input, is given by:

$$A_{vd} = g_{mi} r_{out}, \tag{7}$$

where r_{out} is the output resistance of the OTA, which can be approximated as:

$$r_{out} \cong g_{m5,8} r_{ds5,8} r_{ds5c,8c} \parallel g_{m11,14} r_{ds11,14} \left(r_{ds11c1,14c1}/2\right). \tag{8}$$

For very low biasing currents and large transistor channels areas, usually the thermal noise component is dominant. Assuming perfect symmetry of the OTA, its input-referred thermal noise (referred to i-th input) can be expressed as:

$$\overline{v_{nti}^2} = 2\left(\frac{U_T}{I_B}\right)^2 \left(\frac{C_\Sigma}{C_{Gi}}\right) \left[\overline{\iota_{1,2}^2} + 3\overline{\iota_{3-8}^2} + 2\overline{\iota_{10,14}^2}\right], \tag{9}$$

where

$$\overline{\iota_{1,2}^2} = 2qI_B \tag{10}$$

$$\overline{\iota_{3-8}^2} = 4kTg_{ds3-8c}\left(1 + \frac{2}{3}\frac{g_{ds3-8c}}{g_{m3-8}}\right) \times \frac{\left(g_{m3-8} r_{ds3-8c}\right)^2}{\left(1 + g_{m3-8} r_{ds3-8c}\right)^2} \tag{11}$$

$$\overline{\iota_{10-14}^2} = 4kTg_{ds10-14c}\left(1 + \frac{2}{3}\frac{g_{ds10-14c}}{g_{m10-14}}\right) \times \frac{\left(g_{m10-14} r_{ds10-14c}\right)^2}{\left(1 + g_{m10-14} r_{ds10-14c}\right)^2} \tag{12}$$

where it was assumed, that $r_{ds11c1}||r_{ds11c2} = r_{ds14c1}||r_{ds14c2} = r_{ds10c} = r_{ds13c}$.

As it is easy to note, the total input referred noise is increased by the factor of C_Σ/C_{Gi}, as referred to the i-th input, due to the input capacitive divider composed of capacitors C_{Gi}. Since the input linear range is increased in the same proportion, then the DR remains unchanged.

It can be easily concluded from (9)-(12) and previous considerations, that if the multiple-input OTA would be realized with N traditional differential pairs with common output and biased with the current of I_B/N, then, assuming $C_\Sigma/C_{Gi} = N$, the input referred thermal noise and small-signal transconductance for such a circuit would be exactly the same as for the considered MI-OTA, but its linear range in weak inversion would be the same as for a single differential pair. Since for the MI-OTA proposed in this work the linear range is increased by the factor of C_Σ/C_{Gi}, then also the DR of the MI-OTA will be increased in the same proportion, as compared to the multiple-input OTA, with multiple differential pairs at the output. This can be considered as the main advantage of the proposed approach, achieved at the cost of additional silicon area occupied by the capacitances C_{Gi}.

B. PROPOSED FULLY DIFFERENTIAL UNIVERSAL BIQUAD FILTER USING MULTIPLE INPUT OTAs

The proposed fully differential universal second order filter is illustrated in Fig. 3. This configuration comprises of four MI-OTAs, one passive resistor and two passive capacitors. Three differential input voltage nodes, V_{iL}, V_{iB} and V_{iH} are high impedance and one differential output voltage is V_o. By analyzing the presented fully differential filter, the following output voltage V_o is obtained

$$V_o = \frac{s^2 C_1 C_2 V_{iH} - sC_1 g_{m2} g_{m3} R V_{iB} + g_{m2} g_{m1} V_{iL}}{s^2 C_1 C_2 + sC_1 g_{m2} g_{m3} R + g_{m2} g_{m1}} \tag{13}$$

From Eq. (13), five filtering responses with unity voltage gain can be realized with following statements:

- Non-inverting voltage mode lowpass function: input voltage signal is V_{iL}, $V_{iH} = V_{iB} = 0$ (grounded).
- Inverting voltage mode bandpass function: input voltage signal is V_{iB}, $V_{iH} = V_{iL} = 0$ (grounded).

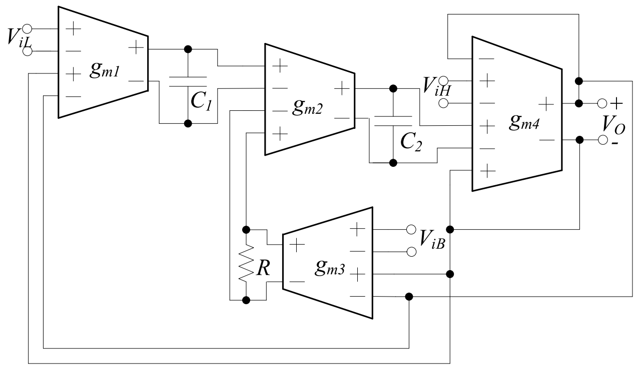

FIGURE 3. Proposed fully differential universal filter.

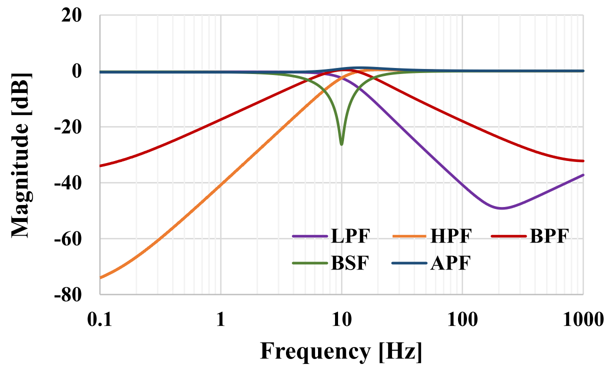

FIGURE 4. Gain response of the topology in Fig. 3.

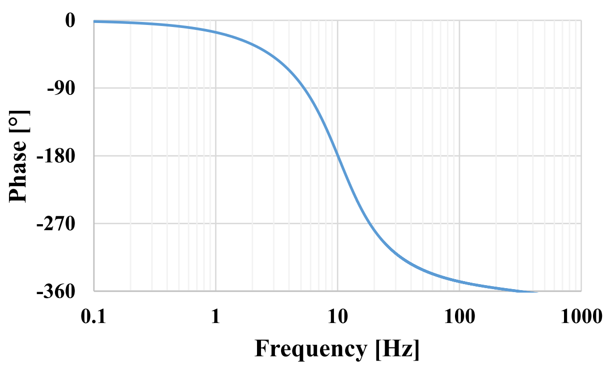

FIGURE 5. Simulated phase and gain response of all-pass function.

- Non-inverting voltage mode highpass function: input voltage signal is V_{iH}, $V_{iB} = V_{iL} = 0$ (grounded).
- Non-inverting voltage mode bandstop function: input voltage signals are V_{iH} and V_{iL}, $V_{iB} = 0$ (grounded).
- Non-inverting voltage mode allpass function: input voltage signals are V_{iH}, V_{iL} and V_{iB}.

From the above statement, it is found that the presented fully differential biquad filter can realize five filter responses without the requirement of element matching conditions and additional circuits (inverting amplifier or double gain amplifier). With this feature, it can be easily controlled with digital programming. From Eq. (13), the angular frequency

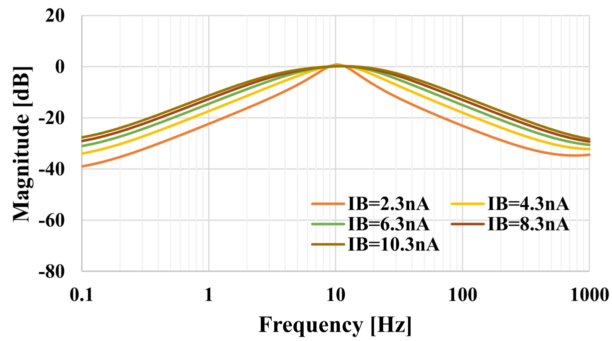

FIGURE 6. BP response for difference values of I_{B3}.

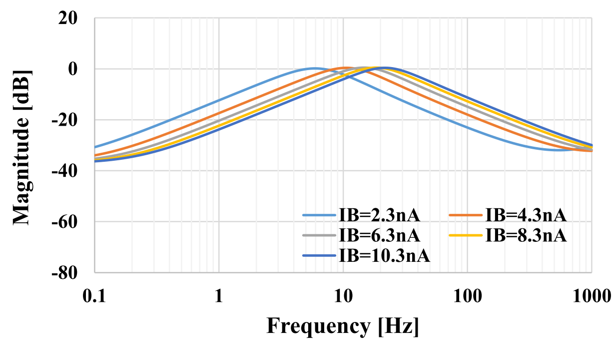

FIGURE 7. BP response for difference values of $I_{B1} = I_{B2} = I_B$.

is given as

$$\omega_0 = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} \tag{14}$$

Subsequently, the quality factor is

$$Q = \frac{1}{g_{m3}R}\sqrt{\frac{C_2g_{m1}}{C_1g_{m2}}} \tag{15}$$

From Eqs. (14) and (15), if $C_1 = C_2 = C$ and $g_{m1} = g_{m2} = g_m$, the angular frequency can be expressed as

$$\omega_0 = \frac{g_m}{C} \tag{16}$$

Later, the quality factor is given to be

$$Q = \frac{1}{g_{m3}R} \tag{17}$$

Eqs. (16) and (17) confirm that the linear adjustability of the ω_0 can be electronically and independently achieved from the Q via g_m. Additionally, the Q is tuned electronically and independently from the ω_0 via g_{m3}.

III. SIMULATION RESULTS

To prove the functionality of presented fully differential biquad universal filter designed in Fig. 3, the Cadence software and the Spectre simulator of the Analog Design Environment were used to design and simulate with the 0.18 μm TSMC CMOS technology. The internal structure of the multiple input gate-driven OTAs was realized as depicted in Fig. 2. For the purpose of simulation, the bias currents

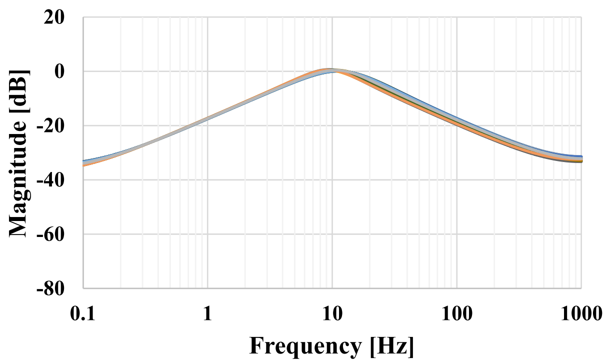

FIGURE 8. BP filtering performance for PV corners.

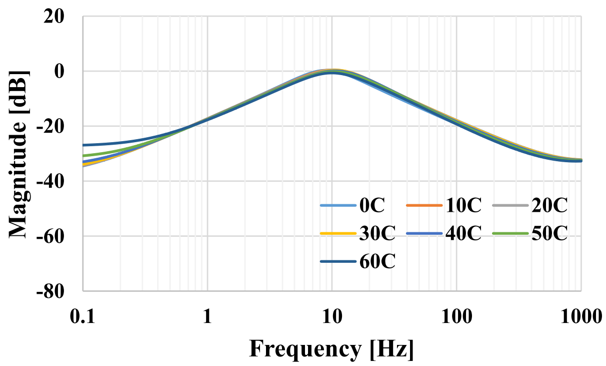

FIGURE 9. Frequency characteristic of BP response for different values of temperature.

for g_{m1}, g_{m2} and g_{m3} were $I_{B1} = I_{B2} = I_{B3} = 4.3$nA and the passive elements values were selected as R= 50 MΩ and $C_1 = C_2 = 450$ pF. The proposed filter was supplied by a symmetrical voltage supply of ±0.25 VDC. By using the referred active and passive element values, the expected angular frequency from Eq. (14) and quality factor from Eq. (15) are f_0 =10Hz, Q= 1. The simulation in Fig. 4 are the magnitude responses of LP, HP, BP, BS and AP filtering functions obtained from the presented biquad filter. The angular frequency is approximately 10Hz. The result of simulated frequency response for phase of AP function is depicted in Fig. 5. From this simulation result, the simulated phase response against frequency of the all-pass filtering function is from 0° to -360°. The simulation in Figs. 4 and 5 insist that the presented fully differential biquad universal filter can offer five filtering responses for the same circuit architecture. The Q adjustment without disturbing the ω_0 was tested as exhibited in Fig. 6 where the value of bias current I_{B3} was set for five values, 2.3 nA, 4.3 nA, 6.3 nA, 8.3 nA and 10.3 nA. The tuning of the ω_0 without affecting Q was proven in Fig. 7, where I_{B3} =4.3 nA and $I_{B1} = I_{B2} = I_B$ were set for five values, 2.3 nA, 4.3 nA, 6.3 nA, 8.3nA and 10.3 nA, the simulated angular frequency were respectively located at 5.9 Hz, 10Hz, 14.7 Hz, 18.4 Hz and 21.6 Hz. The influence of process, voltage and temperature (PVT) variation on filtering performance for BP filtering function was investigated.

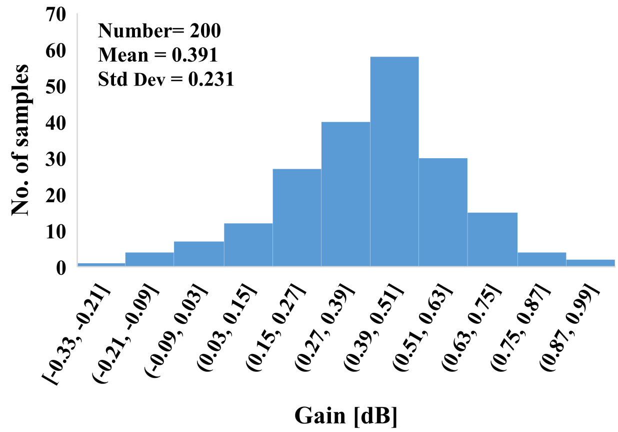

FIGURE 10. Monte-Carlo simulation of the BP filter at f0 = 10 Hz.

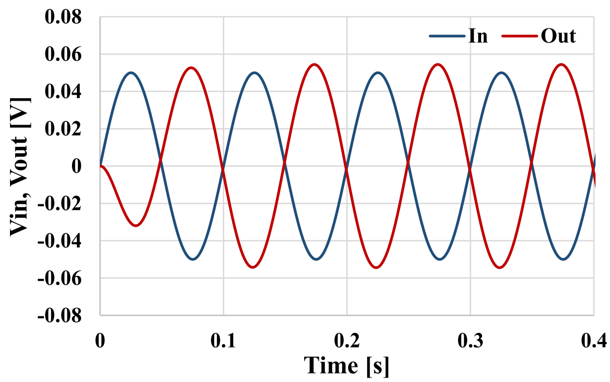

FIGURE 11. Simulated time domain response of the BP filter.

The different process corners were simulated including fast-fast, fast-slow, slow-fast and slow-slow, where voltage supply corners were ±2% around the nominal value of the power voltage supplies. The simulation in Fig. 8 proves acceptable low sensitivity of the BP filter to process and voltage supply variations. Also, the frequency response of the BP filter under different temperature corners from 0 °C to 60 °C is shown in Fig. 9 and this result proves satisfactory low sensitivity to temperature variations. Moreover, Monte Carlo (MC) simulation of the BP filter at $f_0 = 10$ Hz was done with 200 runs. The MC simulation results are illustrated in Fig. 10 where the standard deviation of the voltage gain was 2.96 mdB. The simulated output voltage sinusoidal signal in time-domain for BP filter is depicted in Fig. 11. In this simulation, the sine wave input voltage with 100 mV_{pp}, f = 10Hz was fed at input node V_{iB}. The total harmonic distortion (THD) for f=10 Hz is plotted in Fig. 12.

IV. COMPARISON TABLE

A comparative study of the proposed fully differential biquad filter with several fully differential filters previously published in the literature [10]–[21] is shown in Table 1. As it can be concluded from Table 1, the proposed fully differential versatile filter operates at lowest power supply voltage and power consumption. In addition, the proposed

TABLE 1. Comparison of features of proposed universal filter with several fully differential filters.

Ref.	Technology (μm)	Supply (V)	Power	f_0 (Hz)	No. of devices	DR (dB)	Functions	Electronical tune of f_0 and Q	Orthogonal tune of f_0 and Q	THD
[10]	0.18	0.5	0.74 μW	100	3 FDDA	73	LP, BP, HP, BS, AP	no	yes	<1%@330 mVpp
[11]	0.18	0.5	1 μW	270	3 FDDA	-	LP, BP, HP	no	yes	-
[12]	1.2	±1.5	3420 μW	-	3 FBDDA	-	LP, BP	no	yes	-
[13]	0.18	0.8	45μW	30k	2 FDDTA	-	LP, BP, HP	no	no	-
[14] (Fig. 13)	2	±2.5	-	500k	5FDDT	-	LP, BP, HP	yes	yes	-
[15]	1.2	3	2400 μW	-	2 FBFTFN	-	LP, BP, HP, BS, AP	no	yes	-
[16]	0.5	±1.5	-	-	5 OTA	-	LP, BP, HP	yes	Yes	-
[17]	0.13	0.9	0.55-1 μW	15	4 OTA	70	LP, BP	yes	Yes	-63dB (@5mV)
[18]	0.8	2	10 μW	-	4 OTA	-	LP, BP	yes	yes	<1%@2V
[19]	0.25	±1.5	-	13k	3 FDCCII	-	LP, BP	no	yes	-
[20]	1.2	±1.5	-	1.59M	3 FDCCII	-	LP, BP	no	yes	-
[21]	0.5	±1.5	-	450k	5 DFDCC	-	LP, BP, HP	yes	yes	
Proposed	0.18	0.5	53.3 nW	10	4 OTA	63	LP, BP, HP, BS, AP	yes	yes	<1%@100 mVpp

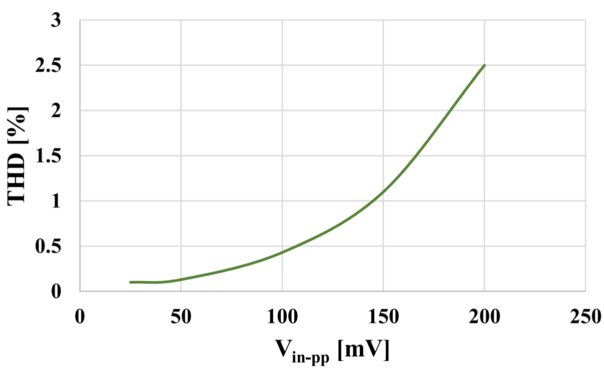

FIGURE 12. THD against the amplitude of the input voltage at f=10 Hz.

fully-differential universal biquad filter can provide more filtering responses compared with [10]–[13], [15], [19], [20] with electronic and orthogonal control of f_0 and Q. Thus, the presented fully differential versatile second-order voltage-mode filter is suitable to be used in LV-LP systems.

V. CONCLUSION

This work proposes a fully-differential three-input and single output biquad filter using multiple-input gate-driven OTA. The proposed filter consists of four MI-OTAs, two capacitors and one resistor. The input voltage nodes, V_{iL}, V_{iB} and V_{iH}, are high-impedance nodes. The features of the proposed filter are proved with simulation. The obtained results indicate that the Q can be electronically tuned via I_{B3} without disturbing the ω_0. Also, the control of ω_0 is electronically and linearly done without disturbing the Q by simultaneously setting I_{B1} and I_{B2}. The simulation with 0.5V supply voltage using the Cadence software and the Spectre simulator verified the good functionality and superiority of the proposed filter.

The simulated dynamic range for BP filtering function is 63 dB for 2% 3rd IMD. The power consumption of MI-OTA based fully-differential universal biquad filter is in nano-watts range thus it is the best choice for LV-LP analog integrated circuits for signal processing applications. Moreover, the simulated results including Monte Carlo and PVT variation analysis, which confirm the theoretical analysis, are compared with the characteristics of some previous fully differential biquad filters.

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