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A 1.9 nW, Sub-1 V, 542 pA/V linear Bulk-Driven OTA with 154 dB CMRR for Bio-Sensing Applications

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Abstract: In this paper, a new technique for improvement on the DC voltage gain, while keeping the high-linearity in symmetrical operational transconductance amplifier (OTA) bulk-driven (BD) topology is proposed. These features are achieved by allying two topological solutions: enhanced forward-body-biasing self-cascode current mirror, and source degeneration. The proposed concept is demonstrated through simulations with typical process parameters and Monte Carlo analysis on nominal transistors of the CMOS TSMC 180 nm node. Results indicate that the proposed OTA can achieve a very small transconductance, only 542 pA/V while keeping a voltage gain higher than 60 dB, 150 dB CMRR, and high linearity of 475 mVpp (1% THD), consuming only 1.9 nW for a supply voltage of 0.6 V. This set of features allows the proposed OTA to be an attractive solution for implementing OTA-C filters for the analog front-ends in wearable devices and bio-sensing.

Keywords: bulk-driven OTA; transconductor; self-cascode mirror



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1. Introduction

The effort to develop implantable or bio-sensing battery-less biomedical instrumentation systems has been continuously challenging analog designers because of the intensified constraints arising from CMOS scaling [1–3]. Topological solutions for endowing operational transconductance amplifiers (OTAs) to process μV signals with common-mode swings in the range of tens of volts, allied to features like ultra-low power consumption, low-noise, enhanced linearity, high common-mode rejection ratio (CMRR), tiny silicon footprint, and large common-mode range (CMR) are frequently pursued by the analog circuit designers [4–18].

As a basic block in analog front-ends (AFEs) for biosensing, the OTA-C filter with large time constants is among the most important applications for OTAs with reduced transconductance [19]. Such circuits when used in implantable/wearable biomedical applications have their design challenged by the restricted-sized on-chip integrated capacitors. In order to decrease the size of such filters, OTAs must output a very small transconductance in the order of a few nA/V, which is achieved with very low biasing currents [20] at the cost of the OTA linearity.

Among the typical OTA design techniques to increase linearity is the use of non-unity gain current mirrors [21–23] to allow higher biasing currents and maintain a low transconductance. Another well-known technique that is used to improve both OTA linearity and input signal voltage swing is the bulk-driven differential pair [1,24–29]. Unlike the gate-driven OTA topologies, the bulk-driven OTAs outputs are an alternative for a relatively lower transconductance [20,30]. In this case, the main drawback of this approach is a poor DC voltage gain, which can be improved by using several techniques [25]. An interesting and widely employed technique relies on a self-cascode topology known

as trapezoidal or composite transistor [31–34]. Additionally, an improvement for the self-cascode transistor association was proposed in [13,35–37] allowing to increase voltage gain and decrease area usage. Therefore, in this paper, we propose a new symmetrical bulk-driven OTA topology that takes the advantages of previously described techniques, i.e., the combination of the topology presented by [23], with a bulk-driven differential pair [24], and the bulk-driven active source degeneration linearization technique adapted from [1,38]. Besides the employed combination of techniques in the OTA topology, we propose an innovative improved self-cascode current mirror (ISCCM) which is based on [35,37].

This paper is organized as follows: Section 2 describes current mirror topologies made of rectangular transistor arrays (composite transistor). The improved self-cascode current mirror that sources the proposed OTA is introduced. Section 3 presents the bulk-driven symmetrical OTAs topologies. Simulations and comparisons among the proposed BD topology, the conventional bulk-driven, and state-of-art transconductors are shown in Section 4. Finally, Section 5 presents the conclusions.

2. Current Mirrors

Current mirrors are the essential component of CMOS OTAs, and their output impedance improvement leads to OTAs with superior voltage gain and common-mode rejection. Implementing current mirrors with series-parallel associations of transistors is a design solution that allows for high current gain, reduced area usage, and less process variability compared to parallel-only current mirrors [22]. This technique was employed by [23] to achieve very low transconductance OTAs without sacrificing linearity and process variability tolerance [13,14]. Since the output transistor array has a large equivalent channel length (L_{eq}), the output current I_o is less dependent on the output voltage V_{out} variation.

Rectangular transistor arrays as illustrated in Figure 1 can be considered understood as a single transistor [31] with a higher output impedance [2,23,39]. The rectangular array, shown in Figure 1, is an m by n matrix of single transistors composed of m parallel columns of n series single transistors. The rectangular equivalent transistor aspect ratio S_{eq-R} is a function of the single transistor aspect ratio S_u , as shown in Equation (1). The total gate area of the rectangular array is $A_T = (mn)A_u$, where A_u is the gate area of the single transistor.

$$S_{eq-R} = \frac{W_{eq}}{L_{eq}} = \frac{mW_u}{nL_u} = \frac{m}{n} S_u. \tag{1}$$

Figure 2 represents an N-type improved composite transistor. It consists of a series connection of two independently forward-body-biased N-type MOS transistors M_{N1} and M_{N2} , as first proposed in [35], and described in detail in [13,14,37], by using the ACM (advanced compact model) all-region transistor model [40].

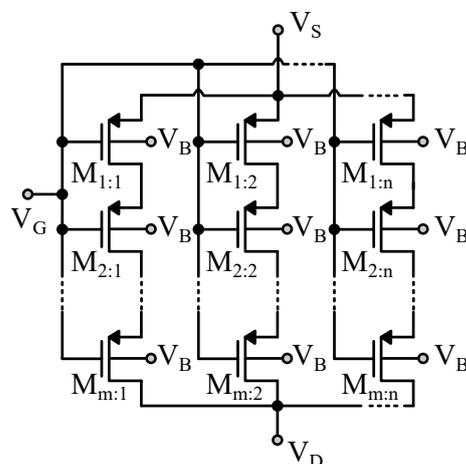


Figure 1. Rectangular $1 \times m : n$ transistor array.

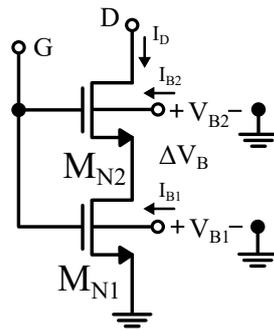


Figure 2. Improved composite transistor.

The improved composite transistor equivalent aspect ratio S_{eq} is defined as:

$$S_{eq} = \frac{S_{N1} \cdot \beta S_{N2}}{S_{N1} + \beta S_{N2}} = \frac{\beta k}{\beta k + 1} \cdot S_{N1} \tag{2}$$

where

$$\beta \approx e^{\frac{(n-1)\Delta V_B}{n\phi_t}} \tag{3}$$

represents a correction factor for the current drain I_D definition due to the difference between the body-bias of the series transistors M_{N2} and M_{N1} $\Delta V_B = V_{B2} - V_{B1}$, assuming the transistors are operating in weak inversion, and

$$k = \frac{S_{N2}}{S_{N1}} \tag{4}$$

is the ratio between transistors M_{N1} and M_{N2} and physical aspect ratios S_{N1} and S_{N2} . Figure 3a shows the conventional current mirror (CM). The ratio between transistors M_{1B} , and M_{1A} aspect ratios S_{1A} and S_{1B} define the current mirror gain $A_I = S_{1B}/S_{1A}$ and attenuation $1/A_I$. In order to have a better matching for non-unity current gain, the current mirror transistors should be replaced with rectangular transistor arrays [22].

A higher current attenuation is achieved by combining parallel transistor arrays at the current mirror input, and series transistor arrays at the output. This scheme is a desirable feature for ultra-low transconductance OTAs [21,23], as it provides transconductance attenuation without decreasing linearity.

The typical cascode current mirror is a variation of the Wilson current mirror first proposed by [41]. The topology increases the output impedance in order to decrease the output current gain error. On the other hand, its drawback is a lower output voltage swing, which will be solved by the proposed current mirror as follows.

An alternative topology to a typical cascode, is the self-biased self-cascode current mirror (SCCM), first proposed by [42], which uses composite transistor arrays in a trapezoidal shape, which are equivalent to single transistors with increased output impedance. The trapezoidal geometry means that the top composite transistors, i.e., those related to drain portion must have a greater aspect ratio than the bottom transistors, i.e., corresponding to source portion, so this kind of composite transistor can be made by arranging their drain transistors in an array connected to a series array corresponding to source transistors (the smaller base of the trapezoid) [43]. This topology is recommended for low input currents and unity current mirror gain, but it is not appropriate for higher currents or very large current gains, since it would require a very large area. Nevertheless, the trapezoidal current mirror can still use the parallel-series technique for current attenuation [21,22] by replacing the output series transistor array with a trapezoidal transistor array, as shown in Figure 3b. This is possible because there is no need for trapezoidal arrays at the mirror input for non-unity gains.

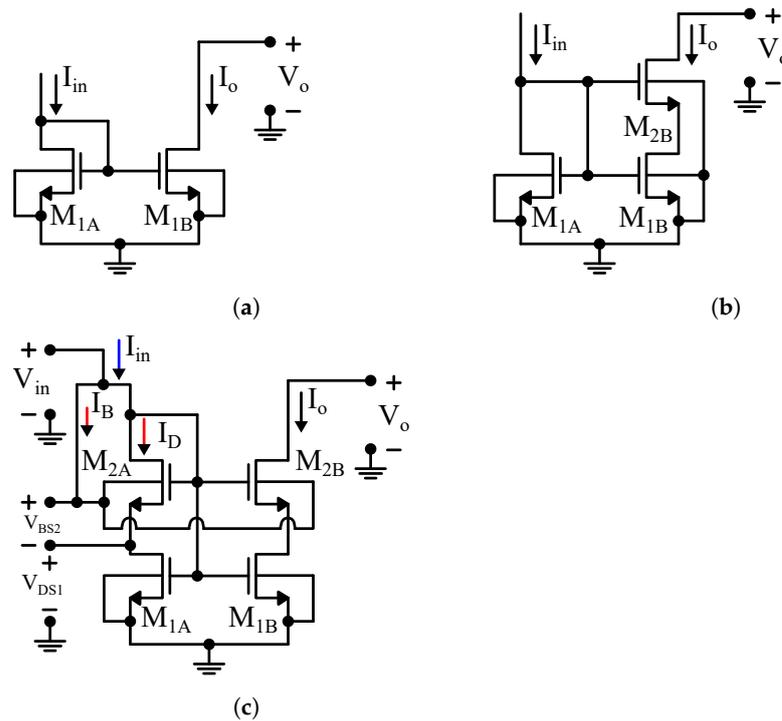


Figure 3. Self-biased current mirrors: (a) conventional current mirror with rectangular transistor arrays (CM) [22], (b) trapezoidal output current mirror (SCCM) [42], and (c) improved self-cascode current mirror (ISCCM).

By taking (2), $\beta = 1$, and since M_{1B} and M_{2B} bulk terminals are connected to each other, the current gain A_I can be expressed as

$$A_I = \left(\frac{S_{2B}}{S_{1B} + S_{2B}} \right) \cdot \frac{S_{1B}}{S_{1A}} \tag{5}$$

For $S_{2B} \gg S_{1B}$, the SCCM current gain is approximately S_{1B}/S_{1A} , as in the conventional parallel-series current mirror. However, this current mirror has a relatively larger output resistance, consequently, it is more tolerant to output voltage variation.

The SCCM output resistance can be further increased by independently forward-body-biasing transistors M_{1B} and M_{2B} by connecting their shared gate terminals to their shared bulk-terminals [37], as shown in Figure 3c. In its turn, the k factor is increased by a β factor function of the bulk-to-source voltage V_{BS2} , accordingly to (3), and hence the gain of the current mirror, A_I is defined as

$$A_I = \frac{S_{1B} \cdot \beta S_{2B}}{S_{1B} + \beta S_{2B}} \cdot \frac{S_{1A} + \beta S_{2A}}{S_{1A} \cdot \beta S_{2A}} = \left(\frac{S_{1A} + \beta S_{2A}}{S_{1B} + \beta S_{2B}} \cdot \frac{S_{2B}}{S_{2A}} \right) \cdot \frac{S_{1B}}{S_{1A}} \tag{6}$$

Again, considering a high value of β , the current gain A_I is approximately S_{1B}/S_{1A} .

For proof of concept, the above current mirrors were designed for the TSMC 180 nm technology and simulated for typical process parameters and room temperature. Table 1 summarizes the transistor arrays dimensions for each circuit.

First, by considering a fixed 1.6 nA input current I_{in} and an output voltage V_o sweeping from 0 to 600 mV, Figure 4a shows the output current mirrors. According to the transistor arrays dimensions, the conventional rectangular parallel-series current mirror (CM) should attenuate the input current by a $16\times$ factor, and provide a 100 pA current. However, due to non-ideal behavior, it outputs about 125 nA, which is close to $13\times$ attenuation. The self-cascode current mirror (SCCM) behaves similarly to CM, as $S_{2B} = 16 \times S_{1B}$. The improved self-cascode current mirror (ISCCM) has a slightly smaller attenuation, close to $12\times$. The

main difference between these current mirrors is the output resistance $R_o = 1/(dI_o/dV_o)$, shown in Figure 4b. At the saturation region, the SCCM R_o is much higher than CM, while the ISCCM is more than one order of magnitude higher.

Table 1. Transistor sizes (Figure 3).

Mirror		Size (m × (M:N) × W/L)		Size (m × (M:N) × W/L)
RCM	M _{1A}	4 × (4:1) × 1.0 μm/8.0 μm	M _{1B}	4 × (1:4) × 1.0 μm/8.0 μm
TCM	M _{1A}	4 × (4:1) × 1.0 μm/8.0 μm	M _{1B}	4 × (1:4) × 1.0 μm/8.0 μm
	M _{2B}	4 × (4:1) × 1.0 μm/8.0 μm		
ISCCM	M _{1A}	4 × (4:1) × 1.0 μm/8.0 μm	M _{1B}	4 × (1:4) × 1.0 μm/8.0 μm
	M _{2A}	4 × (4:1) × 1.0 μm/8.0 μm	M _{2B}	4 × (4:1) × 1.0 μm/8.0 μm

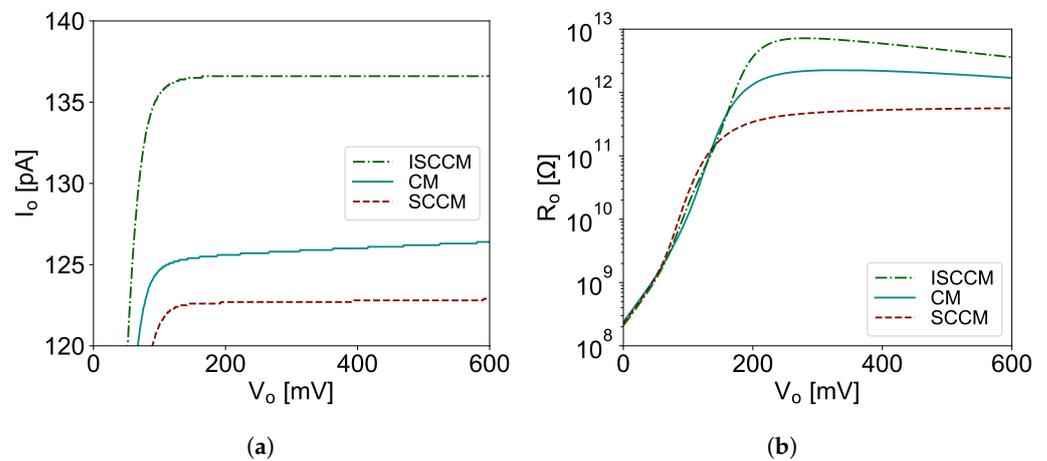


Figure 4. Current mirrors comparison: (a) output current × output voltage, and (b) output resistance × output voltage for $I_{in} = 1.6$ nA.

Nonetheless, the ISCCM is not perfect. Figure 5 shows the current attenuation $1/A_I$ as a function of the input current I_{in} . As can be seen, the current attenuation is practically constant for the CM and SCCM, but it varies for the ISCCM, as the β is indirectly a function of the input current.

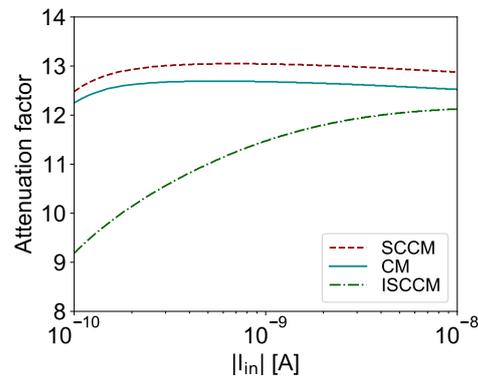


Figure 5. Current mirrors attenuation as a function of input current for $V_{out} = 0.3$ V.

The ISCCM differential bulk voltage is defined as $\Delta V_B = V_{BS2} = V_{in} - V_{DS1}$. As the input current I_{in} increases exponentially, V_{in} increases linearly, as shown in Figure 6a. For 1 nA input, ΔV_B is approximately 100 mV. As V_{BS2} is always positive, the transistors M_{2A} and M_{2B} are forward-body-biased. In spite of that, the drain current I_D is orders of magnitude higher than I_B (see Figure 6b), so $I_{in} \approx I_D$.

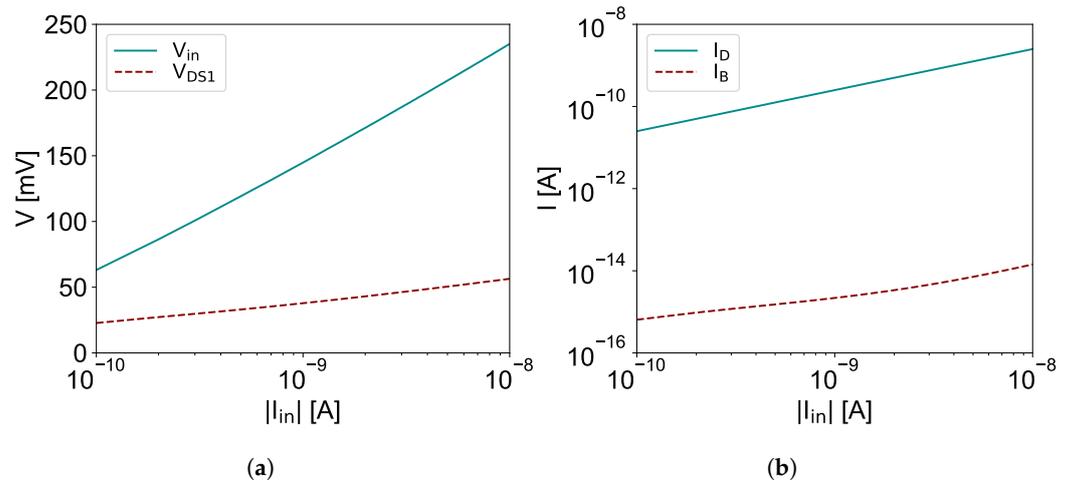


Figure 6. Improved self-cascode current mirror: (a) V_{in} and V_{DS1} voltages \times input current, and (b) drain (I_D) and bulk currents (I_B) \times input current.

3. Bulk-Driven Symmetrical Operational Transconductance Amplifiers

Bulk-driven OTA topologies as illustrated in Figure 7 take advantage of the transistor bulk terminal of the differential pair to achieve higher transconductance linearity and input range rather than conventional gate-driven topologies [25,28].

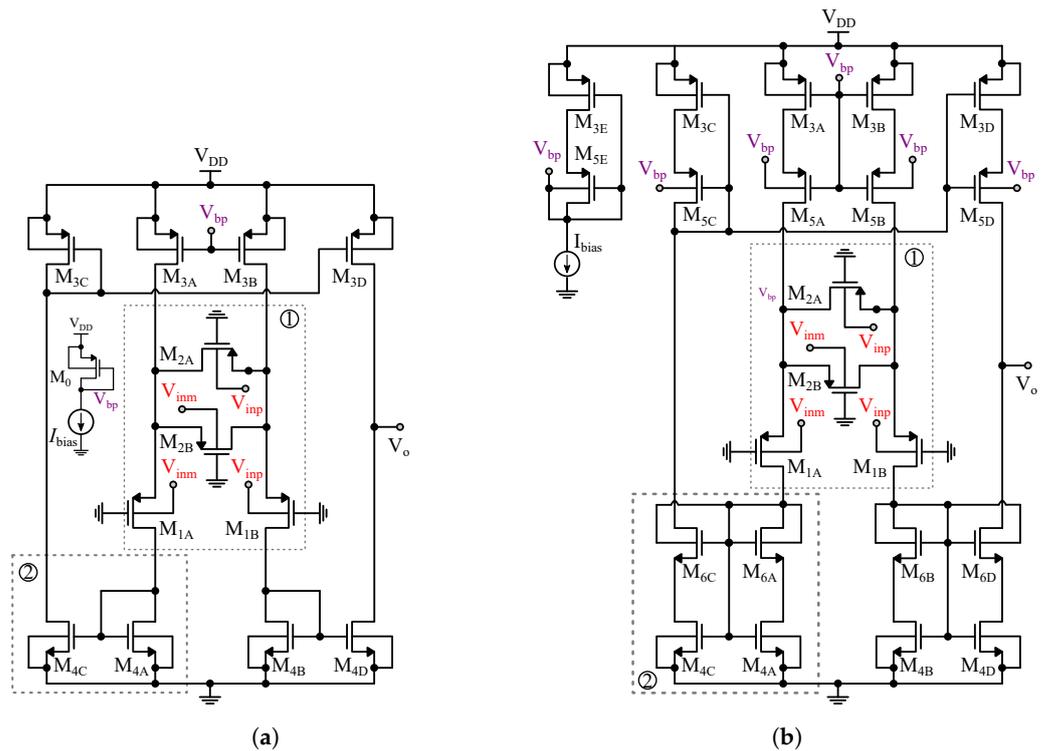


Figure 7. Symmetrical OTA topologies: (a) with parallel-series current mirrors [21], bulk-driven inputs and active source degeneration [1,38], and (b) proposed topology with the addition of improved current mirrors.

The intrinsic drawback of this scheme is the reduced transconductance due to its equivalent gate-driven OTA, hence, a lower DC voltage gain. Nonetheless, biomedical applications frequently involve slow varying quantities and the supposed disadvantage, i.e., the very-low transconductance turns beneficial as analog signal filters with very low cut-off frequencies using relatively small-sized integrated capacitors are essential. Moreover, the lower voltage gain can be addressed with techniques such as positive feedback [25], cascode gain stages [44], and transistor arrays [31,45].

Figure 7b shows the proposed topology which relies on the conventional symmetrical OTA shown in Figure 7a with a key aspect. The current mirrors are built by improved self-cascode configuration [36], according to Figure 3c. This scheme allows increasing the OTA DC voltage gain as also the CMRR. Further, in this work, the conventional BD-OTA (see Figure 7b) makes use of the same active source degeneration technique [1,38,46] employed in the input differential pair to keep fair comparisons between the topologies.

To describe the topology behavior, we use the ACM transistor model (more details in [47]), hence the BD-OTA topology can be explained as follows: the transconductance G_{mB} is a function of the differential pair transconductance g_{mb1} , the source degeneration factor a [46], and the current mirror factor N , as defined by (7). The differential pair transconductance g_{mb1} , defined by (8) and is attenuated relative to the gate-driven OTA by a factor of $(n - 1)$.

$$G_{mB} = \frac{g_{mb1}}{aN}, \tag{7}$$

$$g_{mb1} \approx \frac{n-1}{n} g_{ms1} \approx \frac{n-1}{n} \frac{2I_S}{\phi_t} (\sqrt{1+i_f} - 1). \tag{8}$$

Another advantage of the bulk-driven topology over the gate-driven approach is the reduced minimum supply voltage needed for operation, since the differential pair transistors M_{1A-B} and source degeneration transistors M_{2A-B} gate terminals are connected to the ground instead of to the input signal voltages, which has a typical common-mode voltage of half the supply voltage. It is worth noting that, in order to M_{1A-B} operate in the saturation region, V_{GS1} should be greater than the sum of V_{GS4} and V_{DSAT1} , which is achieved by assuring that i_{f1} is sufficiently greater than i_{f4} [38,46].

The differential pair is composed of the transistors M_{1A-B} , and the active source degeneration transistors M_{2A-B} . The ratio between the differential pair and the source degeneration transistor aspect ratios S_1/S_2 is 4 for the earlier explained reasons and is achieved by using rectangular transistor arrays with the same area. The ratio between the differential pair and the tail current source transistors aspect ratios S_1/S_3 is sixteen, consequently, since the drain currents are the same, the ratio of their forward inversion level i_{f1}/i_{f3} is also 16. The current mirrors use the series-parallel technique [23] to achieve a $16\times$ current attenuation.

As sketched in Figure 7, the conventional BD-OTA, and the proposed one differs because of their body biasing and their current mirror schemes according to Figure 7a,b, respectively. Both the designed conventional and proposed OTAs are composed of the same transistors with the same dimensions. The transistors' sizes of both topologies are summarized in Table 2.

Table 2. Transistor sizes (Figure 7).

Trans.	Size (m × (M:N) × W/L)	Trans.	Size (m × (M:N) × W/L)
M_{1A-B}	$4 \times (1:4) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$	M_{2A-B}	$2 \times (1:8) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$
M_{3A-B}	$4 \times (4:1) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$	M_{3C-E}	$4 \times (1:4) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$
M_{4A-B}	$4 \times (4:1) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$	M_{4C-D}	$4 \times (1:4) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$
M_{5A-B}	$4 \times (4:1) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$	M_{5C-E}	$4 \times (1:4) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$
M_{6A-B}	$4 \times (4:1) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$	M_{6C-D}	$4 \times (1:4) \times 3.0 \mu\text{m}/8.0 \mu\text{m}$

Figure 8 illustrates the layout of the conventional and the proposed BD-OTA. It is possible to observe a very small difference between both topologies, with the tiny occupied area of only 0.00867 mm^2 and 0.0143 mm^2 , for conventional BD-OTA and the proposed BD-OTA, respectively.

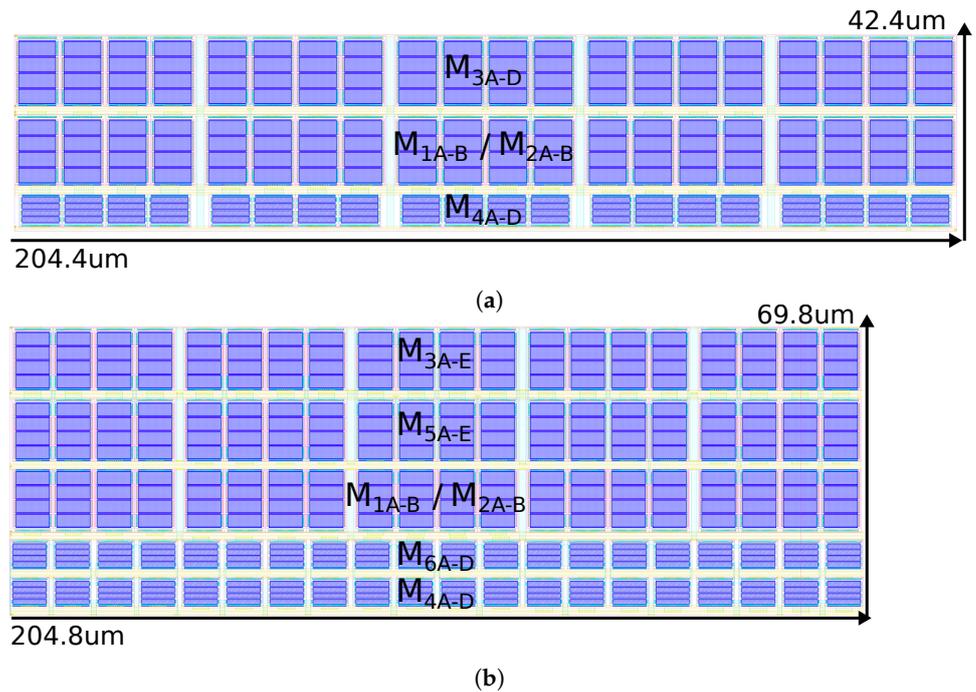


Figure 8. Layout designs of the BD-OTAs. (a) Conventional BD-OTA layout. (b) Proposed BD-OTA layout.

4. Simulation Results

In this section, the post-layout simulation results referring to a TSMC 180 nm CMOS process for the conventional BD-OTA, and the proposed one, are reported. The circuits are considered to operate under the same conditions, i.e., 27 °C temperature, V_{DD} equal to 0.6 V, I_{bias} equal to 100 pA, besides the typical process parameters. Characteristics from both OTAs were obtained by simulating the four testbenches shown in Figure 9. Figure 9a shows the integrator test bench used in the AC and DC simulations. This scheme allows the evaluation and comparison of DC open-loop gain, as also the gain-bandwidth product (GBW) of each OTA version. Then, Figure 10a shows the open-loop gain AC simulation results, and Figure 10b shows the DC simulation results.

It can be noted that the use of improved mirrors increases DC gain without changing considerably the gain-bandwidth product of the OTA versions using the same differential pair, as they are biased with the same current. As expected, the proposed BD OTA with the enhanced mirror has lower transconductance, while keeping higher gain and the same linearity than the typical BD topology.

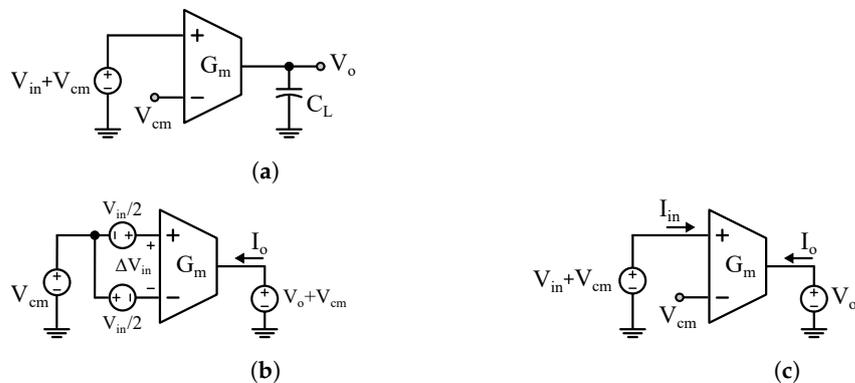


Figure 9. Cont.

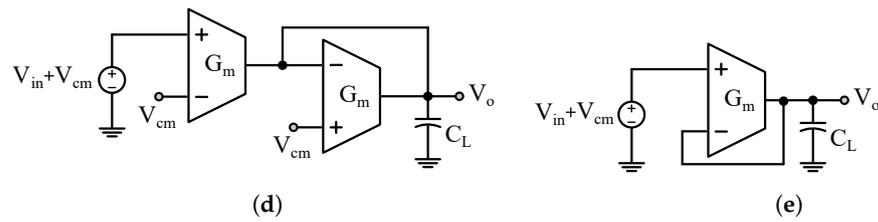


Figure 9. OTA testbenches. (a) OTA-C integrator. (b) Transconductor (Symmetrical). (c) Transconductor (asymmetrical). (d) OTA-C low-pass filter. (e) Unity gain buffer.

As the power supply rejection ratio (PSRR) is equal to the OTAs DC gain, there is a unity gain voltage between supply and output voltages. The common-mode rejection ratio (CMRR) is inherently increased by the use of improved mirrors, as the current source transistors also use improved self-biased cascode configuration. The CMRR and PSRR can be noted in Figure 10c,d, respectively. Table 3 summarizes the AC simulation results.

Table 3. Integrator simulation results summary.

	Conventional BD-OTA	Proposed BD-OTA
DC gain (dB)	44.5	64.2
CMRR (dB)	114	154
PSRR (dB)	88.7	124
GBW (Hz)	78.47	83.14

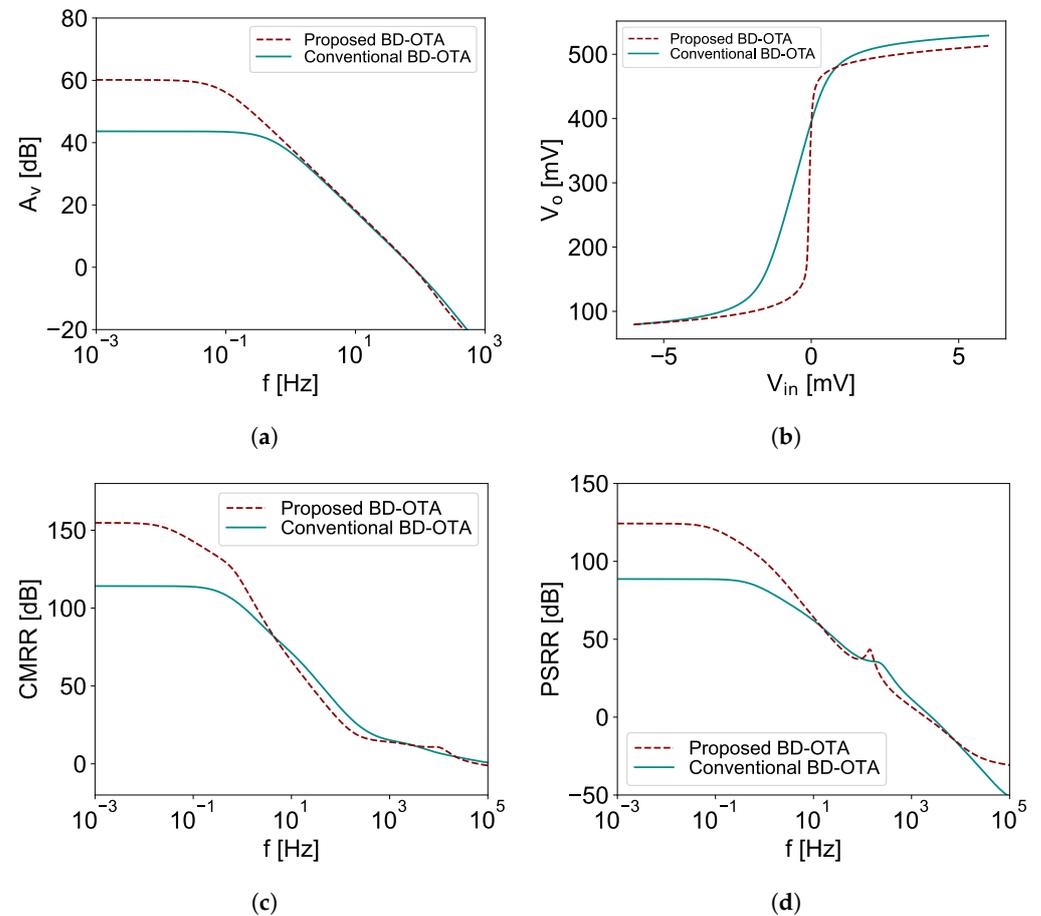


Figure 10. Integrator test bench simulation results. (a) AC voltage gain transfer function. (b) DC voltage gain transfer function. (c) AC CMRR transfer function. (d) AC PSRR transfer function.

Figure 9c shows the testbench used in the DC simulations to compare the transconductance linearity of each OTA version. Figure 11a–c show, respectively, the output current, transconductance, and transconductance error for the conventional BD-OTA and for the proposed one.

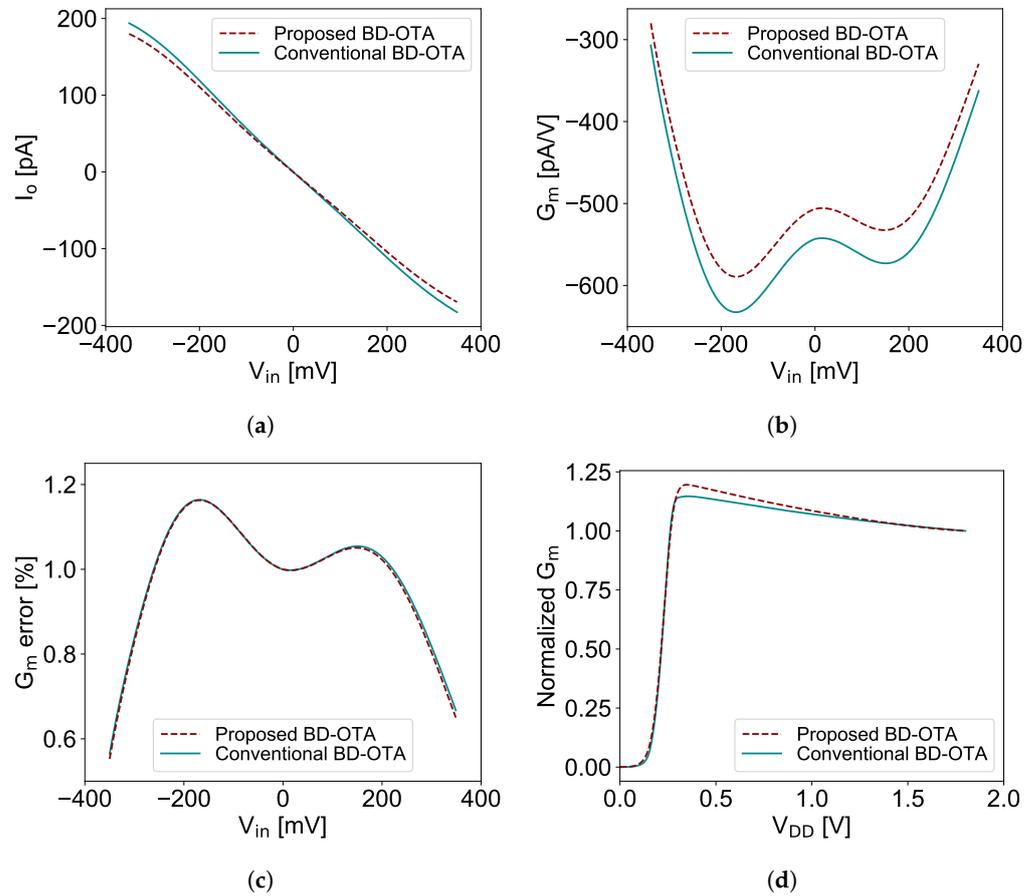


Figure 11. Transconductor testbench simulation results. (a) Output current. (b) Transconductance. (c) Transconductance error. (d) Normalized transconductance.

Table 4 summarizes the transconductance and impedance simulation results for I_{bias} equal to 100 pA. Notice that BD OTAs have finite DC input impedances ($1/G_i$) as large as their output impedances ($1/G_o$), which reduces considerably the effectiveness of the gain improving technique in practical use, where the OTAs are cascaded in OTA-C filters.

Table 4. BD-OTAs DC results.

	Conventional BD-OTA	Proposed BD-OTA
G_i (fA V^{-1})	77.3	78.1
G_m (pA V^{-1})	506	542
G_o (fA V^{-1})	3024.5	311.6
A_v (G_m/G_o)	167.3	1739.4

In Figure 11d, the transconductance normalized with respect to the supply voltage V_{DD} is shown. It is possible to note that both OTA versions work properly from a minimum V_{DD} of about 300 mV, which is feasible for implants and wearable biomedical trends. Unlike conventional gate-driven OTA topologies, which are limited by the minimum common-mode input voltage V_{cmi} , and in which frequently are set to half V_{DD} to allow the current source transistors to operate in saturation, according to mentioned this limitation is mitigated in BD topologies. Besides the mentioned aspects, it is worth noticing that the

minimum operational voltage, V_{DD} , is directly influenced by the current source, and the differential pair transistors channel inversion, hence which are themselves a function of the bias current, i.e., I_{bias} . In this way, a higher biasing current would result in a larger linear input range and greater transconductance, on the other hand, also a higher minimum V_{DD} .

Figures 12a,b and 13a,b show the nominal output current and its resulting transconductance for symmetrical and asymmetrical input voltage, according to the testbenches shown in Figure 9b,c, respectively. For the asymmetrical test, the inverting input is kept constant at $V_{cm} = V_{DD}/2$, so $-300 < \Delta V_{in} < 300$ mV, while, for the symmetrical input, both OTA inputs are at V_{cm} for $V_{in} = 0$ V, and the differential input voltage excursion is doubled to $-600 < \Delta V_{in} < 600$ mV. Moreover, for the asymmetrical testbench, the common mode input voltage V_{cmi} varies with the input voltage V_{in} , so $V_{cmi} = V_{in}/2 + V_{cm}$. For the symmetrical testbench, V_{cmi} is constant, as the average of the inverting and non-inverting input voltages are the same. It can be noticed, for both cases, that as the biasing current I_{bias} increases, the transconductance G_m increases almost proportionally.

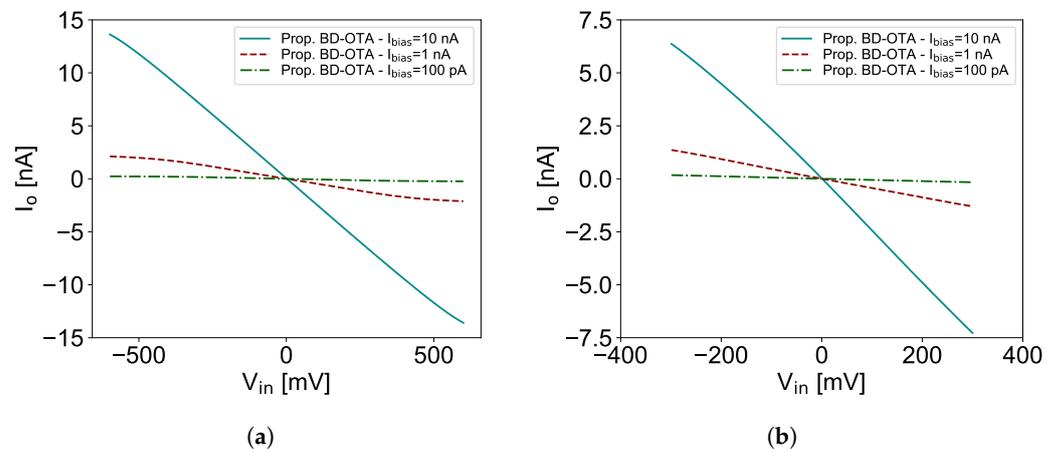


Figure 12. Output current I_o for (a) symmetrical, and (b) asymmetrical input voltage V_{in} , as a function of I_{bias} .

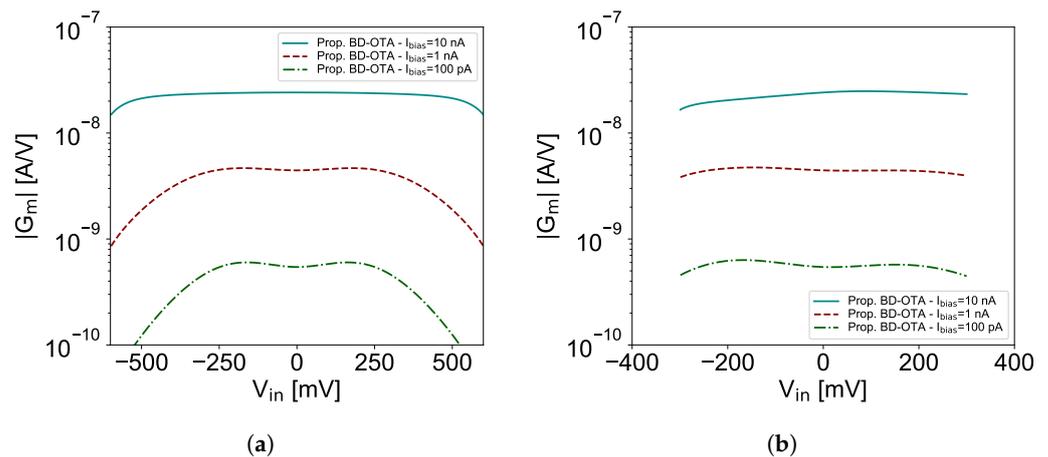


Figure 13. Nominal transconductance G_m for (a) symmetrical, and (b) asymmetrical input voltage V_{in} , as a function of I_{bias} .

For a better comparison, for different biasing currents, the transconductances were normalized for $\Delta V_{in} = 0$, as shown in Figure 14a,b. It is clear for the asymmetrical input that the error is larger for $\Delta V_{in} < 0$. This happens for two reasons: the parasitic substrate current at the differential pair is extremely non-linear and the common-mode input voltage goes above the limit for $I_{bias} = 10$ nA. For symmetrical input, the resulting G_m is also symmetrical and the range is twice as high. It can also be noted that the shape

of the curve changes as the current increases, which is expected, as the differential pair inversion increases.

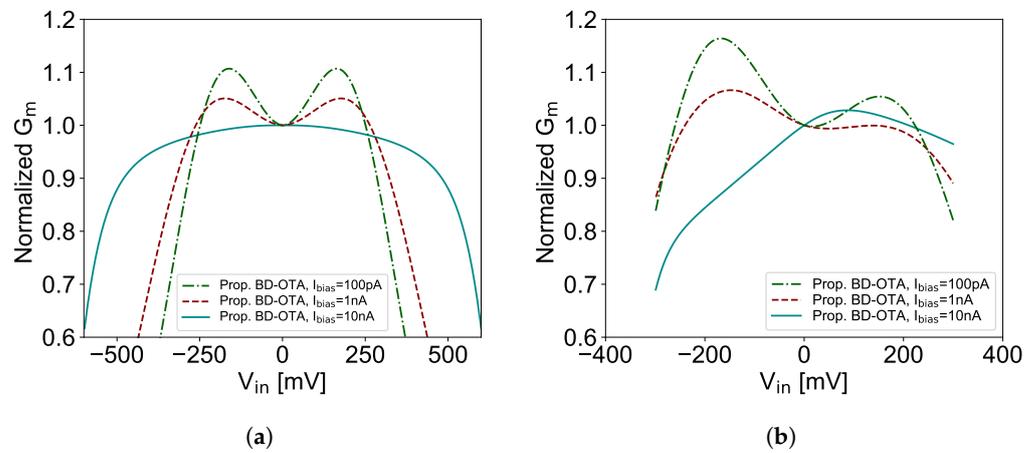


Figure 14. Normalized transconductance G_m for (a) symmetrical, and (b) asymmetrical input voltage V_{in} , as a function of I_{bias} .

It is also important to notice that for single-ended OTA applications, normally, the input is not symmetrical. This is the case with most OTA-C filters, such as those based on integrators and active loads, as depicted in the testbenches shown in Figure 9a,d. For wider range and linearity, the single-ended OTA should be converted to its fully differential version, which needs extra biasing circuits for its output common-mode definition.

As previously explained, the parasitic input current is one of the causes of transconductance asymmetry. This parasitic current is shown in Figure 15a, and is a function of the input voltage and biasing current. There is a single point where the input current is zero, which happens when the differential pair PMOS transistor bulk-terminal voltage is equal to its source-terminal voltage. For input voltages below this point, the transistor is forward-body-biased and the parasitic current grows exponentially. For voltages above this point, the parasitic current is almost constant, consequently, the input conductance is very small. Figure 15b shows the OTA output current for both inputs at $V_{DD}/2$ and the output sweeping from 0 to 600 mV. As can be seen, the output current, even considering that the current mirrors attenuate the differential pair output current, is considerably larger than the parasitic current for a large range.

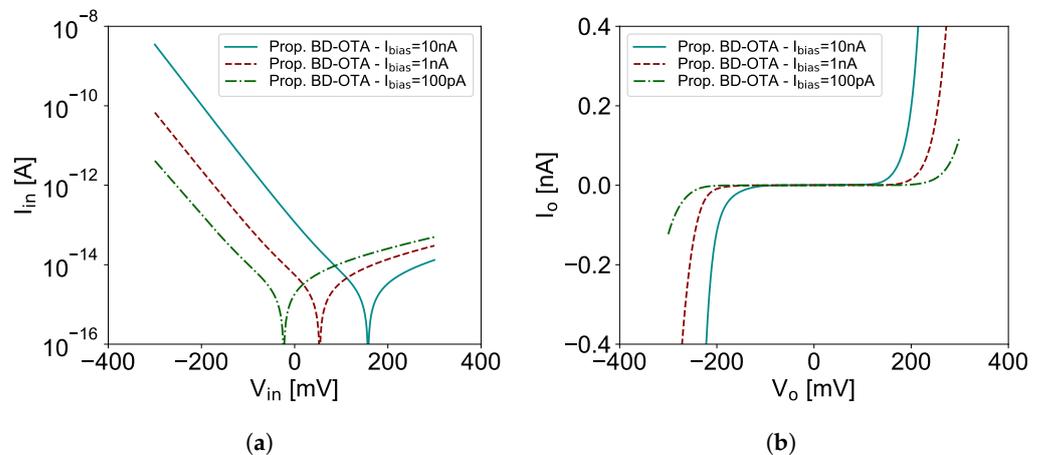


Figure 15. Nominal (a) input current $I_{in} \times$ input voltage V_{in} , and (b) output current $I_o \times$ output voltage V_o , as a function of I_{bias} .

The input and output conductances can be derived from the input and output currents, as shown in Figure 16a,b, respectively. It is worth noting that for OTA-C filter applications, the OTA outputs terminals will be connected to other OTAs input terminals. The main

advantage of the proposed improved self-cascode current mirror is to decrease the output conductance as it increases the output resistance. If the input conductance of the subsequent stage is greater than the output conductance, the technique effectiveness is reduced.

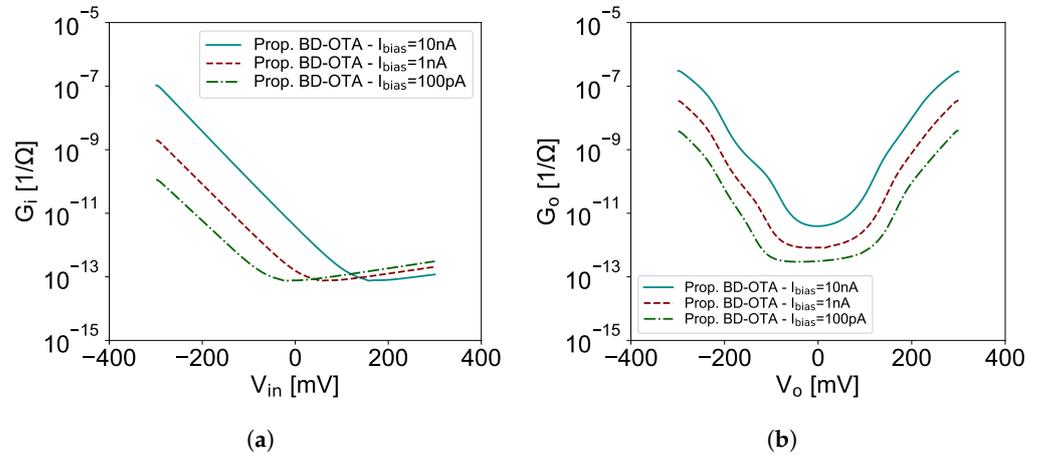


Figure 16. Normalized (a) input conductance G_i , and (b) output conductance G_o as a function of I_{bias} .

In order to compare the linearity OTAs, the unity gain low-pass OTA-C filter test-bench shown in Figure 9d was used in DC and transient simulations. Figure 17a,b show, respectively, the DC transfer functions, transient, and the total harmonic distortion (THD) for both OTAs. It is possible to observe that the BD OTAs have almost the same full input range. Figure 17b shows the total harmonic distortion versus input plotted as a function of input signal amplitude. For both OTAs, one can observe that THD is lower for smaller signal amplitudes. They exhibit approximately the same amount of distortion of 0.07% as a result of a 300 mV amplitude input signal at 100 mHz. As the input voltage amplitude increases, the proposed OTA reaches $\approx 1\%$ THD (-39.8 dB), SNR equal to 56.6 dB for a $V_{in-pp} = 405$ mV at 100 mHz signal.

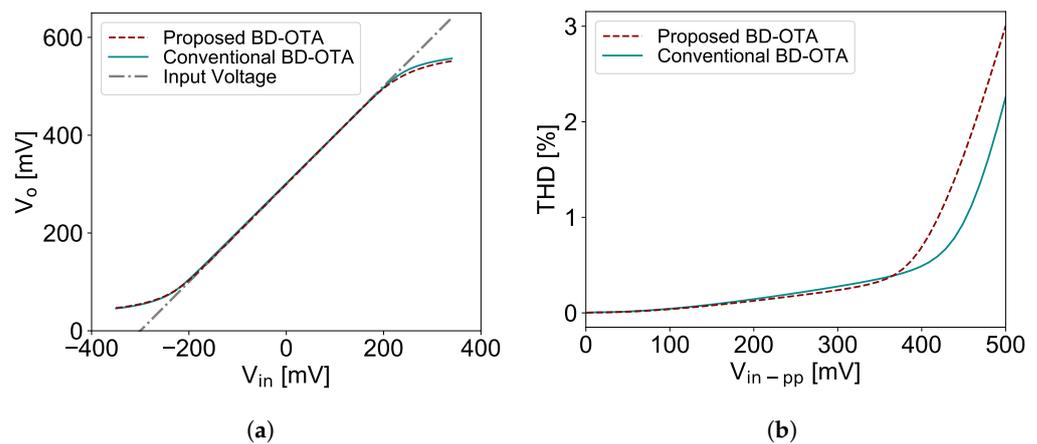


Figure 17. Low-pass testbench simulation results. (a) Transfer functions. (b) Total harmonic distortion.

Figure 18 shows the input-referred noise (IRN) for both OTA versions configured as a unit-gain buffer. Since both OTA versions have the same transistor dimensions, differing only by the adopted current mirror topology, there is a slight difference in IRN of conventional BD-OTA and the proposed one. The IRN in the proposed topology is equal to $246 \mu V_{RMS}$, and $237 \mu V_{RMS}$ in the other, both obtained by integrating noise from 10 mHz–1 kHz.

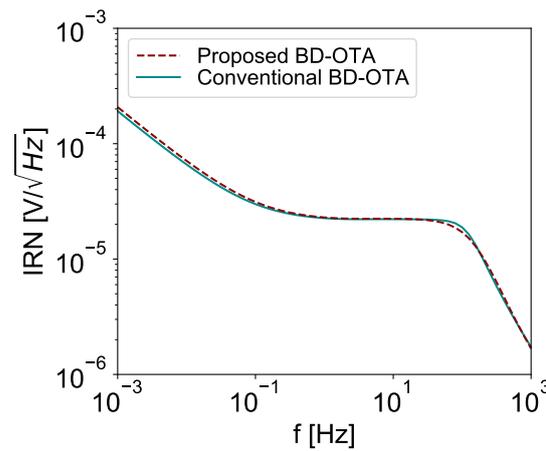


Figure 18. Unity gain buffer testbench simulation results—input-referred noise.

By using the transconductor (Figure 9c) and low-pass filter (Figure 9d) testbenches, 500 runs of Monte Carlo have been carried out for evaluation of transconductance and offset voltage, respectively. Figure 19a,b show the results for the Monte Carlo process and mismatch analysis of the proposed BD-OTA. These results are summarized in Table 5. On this basis, it is possible to conclude that the proposed BD-OTA besides a lower transconductance feature, has a considerably less input voltage offset than the conventional BD-OTA.

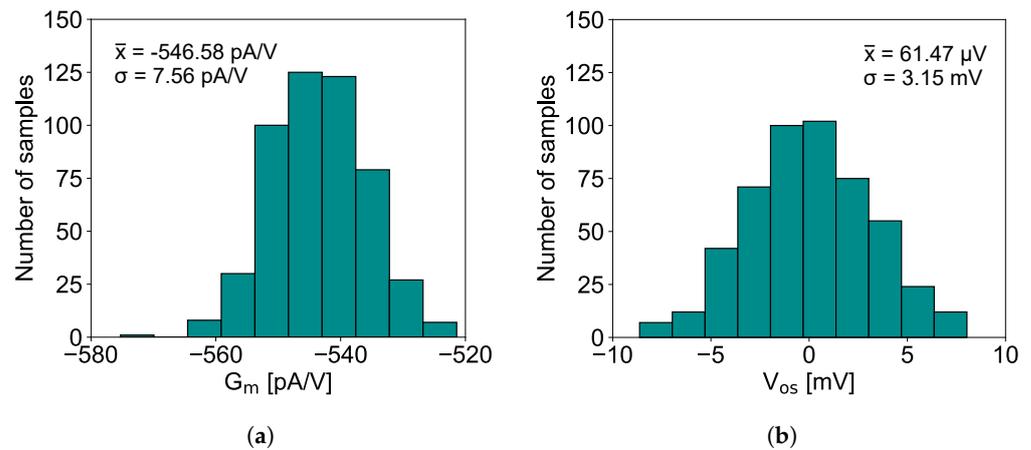


Figure 19. Five hundred runs of Monte Carlo simulations analyzing process and mismatch with the proposed BD-OTA. (a) Transconductance. (b) Offset.

Table 5. Monte Carlo simulation results.

		Conventional BD-OTA	Proposed BD-OTA
G_m	\bar{x} (pA/V)	506.6	546.58
G_m	σ (pA/V)	7.61	7.56
G_m	σ/μ (%)	1.5	1.38
V_{os}	\bar{x} (μ V)	429.8	61.47
V_{os}	σ (mV)	3.16	3.15

Table 6 compares the performance of the proposed OTA with the state-of-art low-transconductance OTAs. The previous work proposed by us [48] presented a 450 pA/V OTA with small power consumption but lower gain, CMRR, and PSRR, despite being based on non-unitary current gain through the splitting current technique, it achieved a poorer performance with respect to the present work. In [49], a low-transconductance amplifier has been proposed based on the channel-length-modulation effect (Early effect). This solution

shows a high IRN. Such an IRN is 3x smaller in the proposed topology while keeping lower transconductance, power consumption, and higher CMRR as also PSRR features. The OTA proposed by [28] is similar to the conventional OTA presented in this work. The difference is in the rectangular arrays used to increase the gain and in no linearization technique employed. Another low- G_m topology presented by [50] uses a linearization technique that relies on a combination of source degeneration with an active attenuator. Despite the valuable linearity and gain achieved, the power consumption, and transconductance may not be suitable to the constraints of biomedical implants or bio-sensing operations. The architecture proposed by [51] is another channel length modulated OTA which contains the same V-I conversion scheme as presented in [49] but requires a higher supply voltage.

Table 6. Comparison of Low- G_m OTA topologies.

Feature	This Work (S)	[48] (S)	[49] (M)	[28] (M)	[50] (M)	[51] (M)
Year	2021	2021	2020	2014	2014	2009
Tech. (nm)	180	180	180	130	350	350
Supply (V)	0.6	1	1	0.25	5	5
G_m (nA/V)	0.542	0.45	0.62–6.28	22	39.5–367.2	0.03–25,000
A_v (dB)	64	37	-	-	52.3–64.7	-
Power (μ W)	0.0019	0.032	0.028–0.270	0.01	160	<300
CMRR (dB)	154	56	56	-	>44.8	>80
PSRR (dB)	124	36	47	-	-	>80
GBW (Hz)	83.14	6.9	-	-	-	-
V_{os} (mV)	0.06 ± 3.15	± 20	25–50	± 10.82	-	-
IRN (μ V _{RMS})	246	-	760	100	-	635
Linear range sym. input (V_{pp})	0.475	0.3	2	-	2	2.6
THD (%) @ input (V_{pp})	1@0.475	1@0.3	0.18@2	0.53@0.1	0.13@2	<1@2.6
SNR (dB) @ THD (%)	54.8@1	-	59.3@0.47	-	66.5@0.13	62@1
Layout area (mm ²)	0.0143	-	0.027	0.053	0.006	0.046

THD @ $V_{in} = 250$ mV_{pp}, (M): Measured, (S): Simulated.

5. Conclusions

The paper presented a bulk-driven symmetrical OTA based on a self-cascode current mirror with source degeneration to provide a high-gain, high linearity, and low- G_m topology. The proposed low- G_m topology shown a valuable performance that is suitable for new biomedical IC applications. In particular, the new topology achieved the lowest power consumption compared with the state-of-art topologies, in addition to high gain, linearity, low transconductance, and IRN. Moreover, the circuit obtained the highest CMRR and high PSRR, which turns the proposed OTA into an interesting topology as a basic block for OTA-C filters.

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Abbreviations

The following abbreviations are used in this manuscript:

ACM	Advanced Compact Model
BD	Bulk-driven
CMOS	Complementary Metal-Oxide Semiconductor
CM	Current Mirror
CMR	Common-Mode Range
CMRR	Common-Mode Rejection Ratio
GBW	Gain–Bandwidth Product
OTA	Operational Transconductance Amplifier
SCCM	Self-Cascode Current Mirror
IRF	Equivalent Input Referred Noise
ISCCM	Improved Self-Cascode Current Mirror
PSRR	Power Supply Rejection Ratio
SNR	Signal-to-Noise Ratio
THD	Total Harmonic Distortion

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