Int. J. Electron. Commun. (AEÜ) 65 (2011) 728–733

Contents lists available at ScienceDirect

International Journal of Electronics and Communications (AEÜ)

journal homepage: www.elsevier.de/aeue

A balanced output CMOS OTA with wide linear current tunable range

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ARTICLE INFO

Article history:
Received 22 March 2010
Accepted 28 October 2010

Keywords:
Balanced output operational transconductance amplifier (BOTA)
CMOS transconductor
Electronically tunable OTA
Linearly tuned

ABSTRACT

A new CMOS-based balanced output operational transconductance amplifier (BOTA) with very wide linear current tunable range is proposed in this paper. The design technique is achieved by the combination of a fully differential transconductor and an electronically variable current gain stage. The transconductance gain of the proposed BOTA can be linearly tuned by an external bias current for more than 4 decades, with less than 4% nonlinearity for the linear input-voltage range of about 0.2 V_{peak}. The OTA is designed in 0.5 μm AMIS technology. The performance of the proposed circuit is discussed and confirmed through application example and PSPICE simulation results.

1. Introduction

Operational Transconductance Amplifiers (OTAs) or voltage-to-current converter circuits are the essential active building blocks of many analog applications such as amplifiers, multiplier, active filters and sinusoidal oscillators. Moreover, they can also be applied in automatic gain control and analog multiplier circuits when the transconductance gain of the OTA can be electronically and linearly varied. There have been several approaches of designing tunable linear CMOS OTA. Those approaches are claimed to achieve the linear operation in the square law of the I–V characteristic of MOS transistors [1–4]. However, most of them are functioning in voltage-controlled mode. Their controllable voltage ranges are rather limited and only narrow linearly transconductance ranges are achieved. In the past, a current-controlled CMOS transconductor was presented in [5]. But the transconductance linear tunable range is narrow due to the MOS transistors are working in the weak-inversion region. On the other hand, an Electronically Tunable Operational Transconductance Amplifier (EOTA), where its g_m can be linearly tuned for a 3-decade range, has been proposed [6]. However, its structure is quite complicated due to the use of 3 CMOS OTAs.

Recently, a tunable CMOS-based balanced output OTA (BOTA) has received some attention [7–9]. It has been demonstrated that filters that realized by using BOTA provide more simplify structures and perform better performance in higher frequency range than the single output OTAs [10–12]. However, since there transconductance gains were controlled by DC voltages, narrow voltage control ranges were available. It is the objective of this paper to propose a new technique to realize a wide linearly current tunable CMOS BOTA. The design method is based on the used of a fully differential transconductor and a variable current gain cell that is modified from a current-mode translinear circuit. The proposed BOTA is simple and suitable for implementation. The BOTA characteristics are verified through PSPICE simulation results. In addition, a current-mode tunable BOTA-based biquad filter is used to demonstrate the usefulness of the proposed BOTA.

2. Circuit descriptions

Fig. 1(a) shows the symbol of a BOTA, which is a device that accepts two input voltages and provide balanced output currents. A new approach to realize a CMOS BOTA that its transconductance gain can be electronically and linearly varied in a very wide range is shown in the building block diagram of Fig. 1(b). It is designed by the cascade connection of a transconductance cell and an electronically tunable current gain cell. Since the key circuit building block in this approach is the electronically current-tunable current gain cell, we will therefore firstly introduce a method to realize a current gain cell that its current gain can be electronically varied in very wide range. Then to realize the BOTA, a fully balanced differential amplifier is use as the transconductance cell. In this work, it is generally assume that all MOS transistors are operating in the saturation region, hence the individual functions of the circuits are derived from the approximate square-law characteristic of MOS transistors in saturation.

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doi:10.1016/j.aeue.2010.10.005

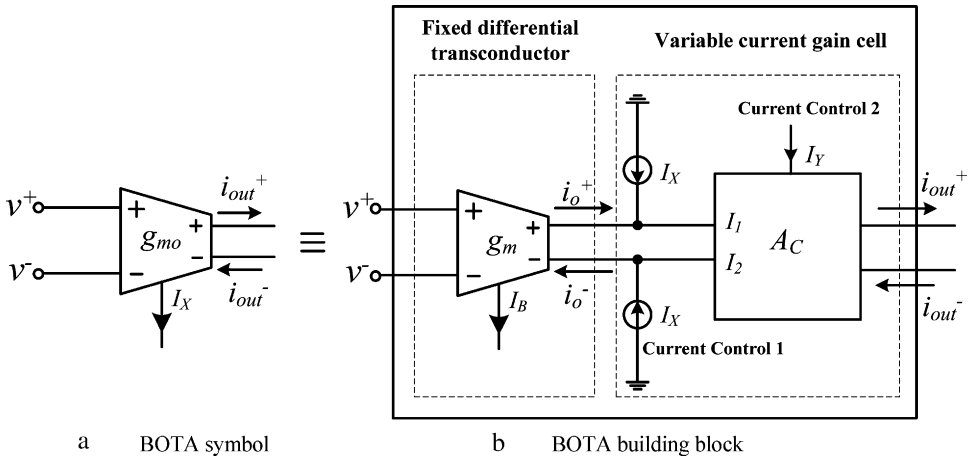

Fig. 1. Building block of the proposed balanced output CMOS OTA.

2.1. Electronically current-tunable current gain cell

Let us consider the schematic circuit diagram of the proposed electronically current-tunable current gain cell shown in Fig. 2, where the operation of the circuit is based on the square-law characteristic of MOS transistors biased in the saturation region. The circuit is modified from the current gain amplifier of Ref. [13] by adding the current source I_X and a dual-output current mirror in order to provide electronically property. Transistors M_1–M_3 and M_9–M_{11} function as current squarer circuits. Transistors M_{17} and M_{18} and the current source I_Y formed as the current-controlled bias circuit that provides a voltage bias V_{REF} to M_3 and M_{11}. The dual-output current mirrors M_4–M_6 and M_{12}–M_{14} have current gain ratios equal to 1:n. Let us assume that differential input signal currents i_{in} and $-i_{in}$ are injected into points B_1 and B_2, respectively. The input signals are added with the DC currents I_X form the current source I_X and the dual-outputs unity gain current mirrors M_{19}–M_{21}. Therefore, the differential input currents that flow into points C_1 and C_2 are $I_{in1} = I_X + i_{in}$ and $I_{in2} = I_X - i_{in}$, respectively. By using the square law of MOS transistors operating in the saturation region, the currents I_1 and I_2 of the current squarer circuits can be, respectively, expressed as [13]

$$I_1 = 2I_Y + \frac{(I_X + i_{in})^2}{8I_Y} \tag{1}$$

$$I_2 = 2I_Y + \frac{(I_X - i_{in})^2}{8I_Y} \tag{2}$$

The current I_1 and I_2 are multiplied n time by the current mirrors M_4–M_6 and M_{12}–M_{14}, respectively, and then subtracted to produce differential output currents i_{out}^+ and i_{out}^- at the ports A_1 and A_2, respectively. The differential output currents i_{out}^+ and i_{out}^- can be given by

$$i_{out}^+ = -i_{out}^- = n(I_1 - I_2) = \left(\frac{nI_X}{2I_Y}\right) i_{in} \tag{3}$$

or

$$i_{out}^+ = \left(\frac{nI_X}{2I_Y}\right) i_{in} = A_C i_{in} \tag{4}$$

where to keep all devices in the on state, the condition $|I_X| + |i_{in}| \leq 4I_Y$ must be satisfied. A_C is the current gain of the current gain cell and can be expressed as

$$A_C = \frac{nI_X}{2I_Y} \tag{5}$$

From (4), if I_Y is fixed, the output current i_{out} can now be electronically and linearly varied by the external DC currents I_X. Note that, the parameter n is included in order to increase the dynamic range of the current gain A_C. In this case the maximum value of A_C is limited by $A_{C(MAX)} \leq 2n$. From the circuit of Fig. 2, since it is

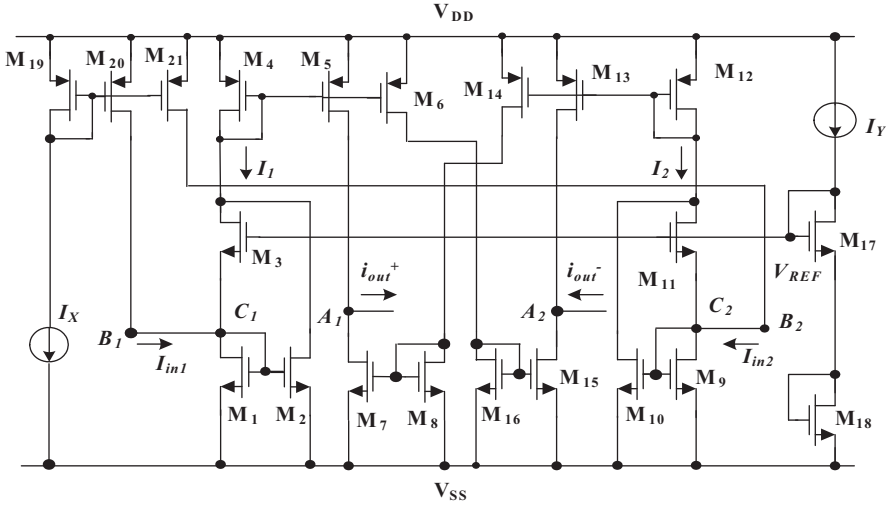

Fig. 2. Electronically current-tunable current gain cell

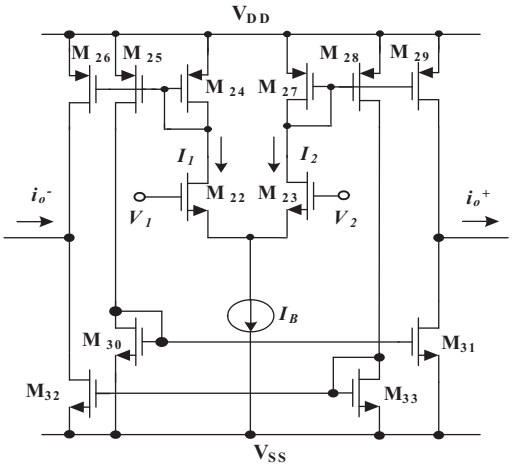

Fig. 3. CMOS differential transconductor.

mostly composed of current mirror circuits, therefore the error of the current gain cell is due to mismatch between the transistors [13,14]. This error can be improved if high-performance current mirrors such as a Wilson current mirror or a cascode current mirror is employed.

2.2. Fully balanced differential transconductor

A fully balanced differential transconductor structure is shown in Fig. 3. Let M_{22} and M_{23} are perfectly matched, the current mirrors have unity current gains and all transistors are operated in saturation region. v_{in} is the small signal differential input voltage ($v_{in} = V_1 - V_2$), i_o^+ and i_o^- are the differential output currents and I_B is the bias current. The differential output currents can be given by

$$i_o^+ = -i_o^- = I_2 - I_1 = \frac{\mu_n C'_{ox}}{2}\left(\frac{W}{L}\right) v_{in}\sqrt{\frac{4I_B}{\mu_n C'_{ox}(W/L)} - {v_{in}}^2} \tag{6}$$

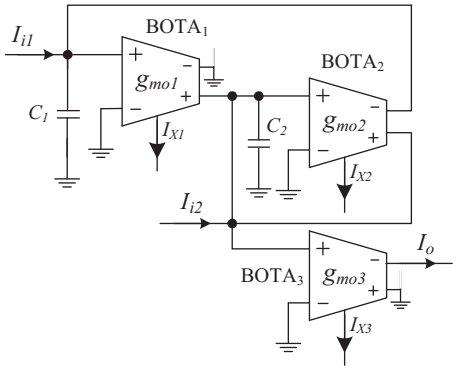

Fig. 5. An electronically and linearly tunable current-mode biquad filter.

The transconductance gain (g_m) of the fully differential transconductor can be derived by taking the derivative of (6) with respect to v_{in}, yielding

$$\left.\frac{di_o^+}{dv_{in}}\right|_{v_{in}=0} = \sqrt{\mu_n C'_{ox}\frac{W}{L} I_B} \tag{7}$$

Therefore, the transconductance gain can be defined as

$$g_m = \sqrt{2KI_B}, \quad -\sqrt{\frac{I_B}{K}} \le v_{in} \le \sqrt{\frac{I_B}{K}} \tag{8}$$

where $K = (\mu_n C'_{ox}/2)(W/L)$ is the transconductance parameter, μ_n is the mobility of the carrier, C_{ox} is the gate-oxide capacitance per unit area, W is the effective channel width, L is the effective channel length. Thus we can obtain the following equation for a small signal input:

$$i_o^+ = g_m v_{in} = \sqrt{2I_B K}\cdot v_{in} \tag{9}$$

It should be noted that the transconductance gain (g_m) of the transconductor shown in Fig. 3 will provide low harmonic distortion if the input voltage is limited in the range of

$$-\sqrt{\frac{I_B}{K}} \le v_{in} \le \sqrt{\frac{I_B}{K}} \tag{10}$$

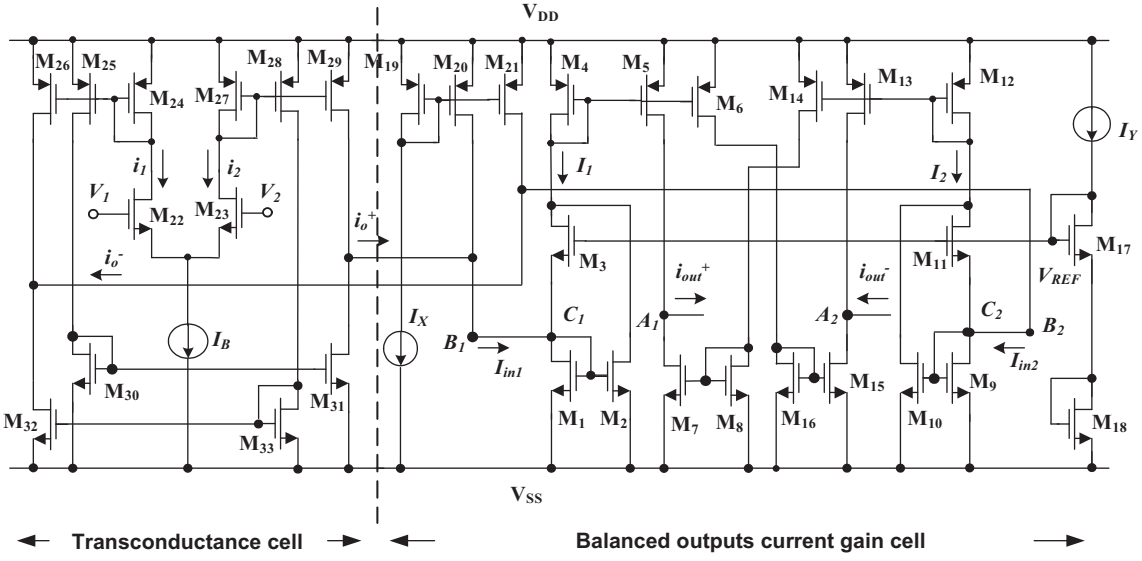

Fig. 4. Complete schematic diagram of the proposed BOTA.

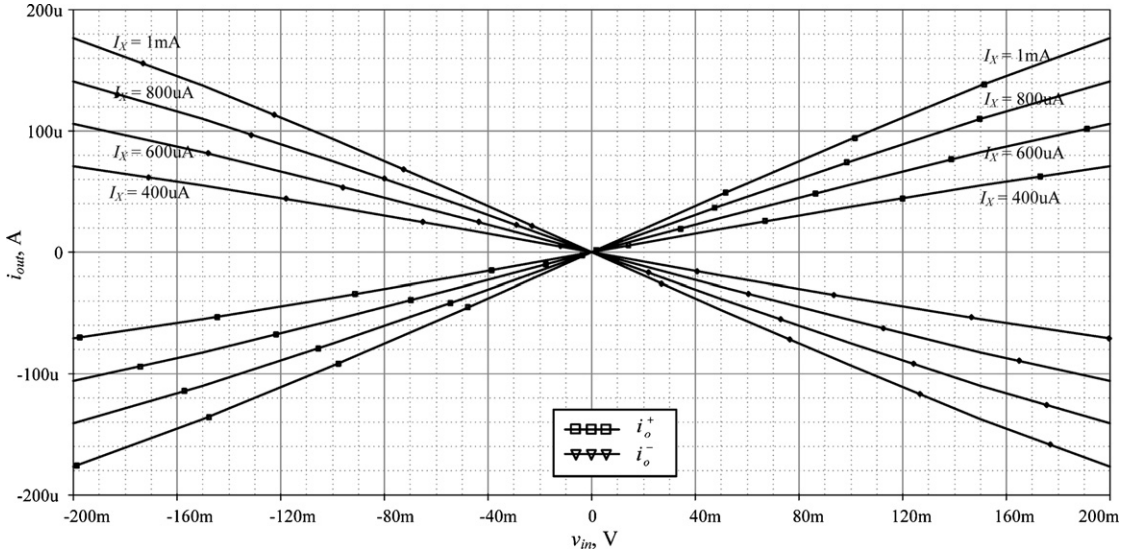

Fig. 6. DC transfer characteristics of the proposed BOTA.

2.3. Complete balanced output CMOS OTA

Based on the approach of the Fig. 1, the schematic circuit diagram of the proposed wide linearly and electronically tunable range CMOS BOTA is shown in Fig. 4. The circuit is simple and suitable for implementing in integrated circuit form. From Eqs. (3) and (4), and since $i_o^+ = i_{in}$ the output current of the proposed circuit can be expressed as

$$i_{out}^+ = A_C \cdot i_o^+ \tag{11}$$

From Eqs. (5), (9) and (11) the output current of the proposed BOTA can now be expressed as

$$i_{out}^+ = \sqrt{2KI_B} \cdot \frac{nI_X}{2I_Y} \cdot v_{in} \tag{12}$$

If g_{mo} is the transconductance gain of the BOTA, it can be shown that

$$g_{mo} = \frac{n\sqrt{2KI_B}}{2I_Y} I_X = K_T I_X \tag{13}$$

where $K_T = n\sqrt{2KI_B}/2I_Y$. The term K_T can usually be kept constant. Eq. (13) clearly indicates that the transconductance gain of the proposed transconductor can be electronically and linearly tuned by the bias current I_X. Since the current $|I_x| + |i_{in}|$ should be limited to be less than $4I_Y$, therefore, for a wide current tunable range, the bias current I_Y should be a high constant current. It worth noting that the input dynamic and nonlinearity range is not affected by the adjustment of I_X.

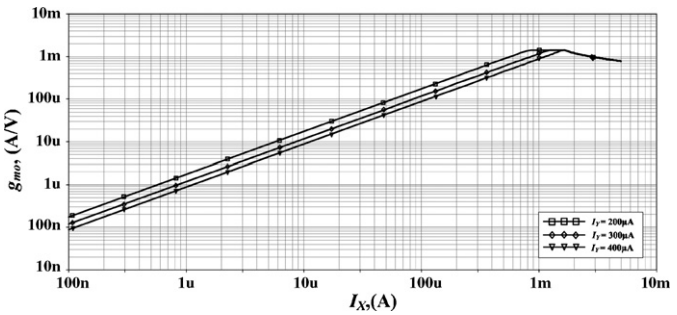

Fig. 7. Transconductance gain linearly tuned range.

3. Application example

In this section, in order to demonstrate its application and usefulness, the proposed BOTA is employed to realize a current-mode electronically tuned biquad filter. Fig. 5 shows the current-mode biquad filter that constructed by using 3 BOTAs [15]. The circuit analysis yields the following transfer functions.

$$I_o(s) = \frac{I_{i1}g_{mo1}g_{mo3}/C_1C_2 + I_{i2}g_{mo3}s/C_2}{s^2 + g_{mo2}s/C_2 + g_{mo1}g_{mo2}/C_1C_2} \tag{14}$$

where I_{i1} and I_{i2} are the input signal currents and I_o is the output signal current of the filter. From Eq. (14), the circuit can realize a low-pass filter if $I_{i2} = 0$ and the band-pass filter if $I_{i1} = 0$, where there transfer function can be, respectively, expressed as

$$\text{Low-pass filter}: \frac{I_o(s)}{I_{i1}(s)} = \frac{g_{mo1}g_{mo3}}{C_1C_2s^2 + C_1g_{mo2}s + g_{mo1}g_{mo2}} \tag{15}$$

$$\text{Band-pass filter}: \frac{I_o(s)}{I_{i2}(s)} = \frac{C_1g_{mo3}s}{C_1C_2s^2 + C_1g_{mo2}s + g_{mo1}g_{mo2}} \tag{16}$$

The cut-off frequency ω_0 and the quality factor Q of the filter can be given as

$$\omega_0 = \sqrt{\frac{g_{mo1}g_{mo2}}{C_1C_2}} \quad \text{and} \quad Q = \sqrt{\frac{g_{mo1}C_2}{g_{mo2}C_1}} \tag{17}$$

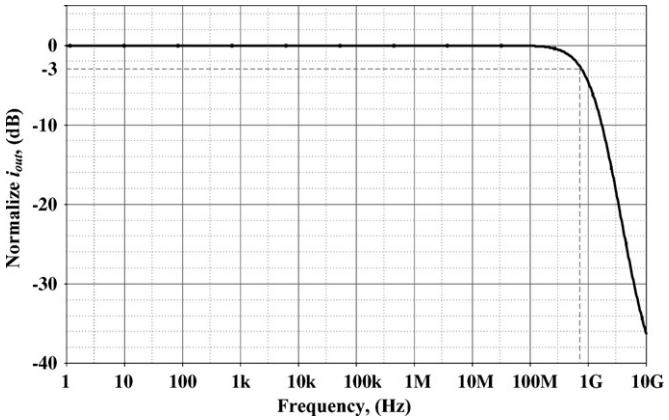

Fig. 8. The frequency response of the proposed transconductor.

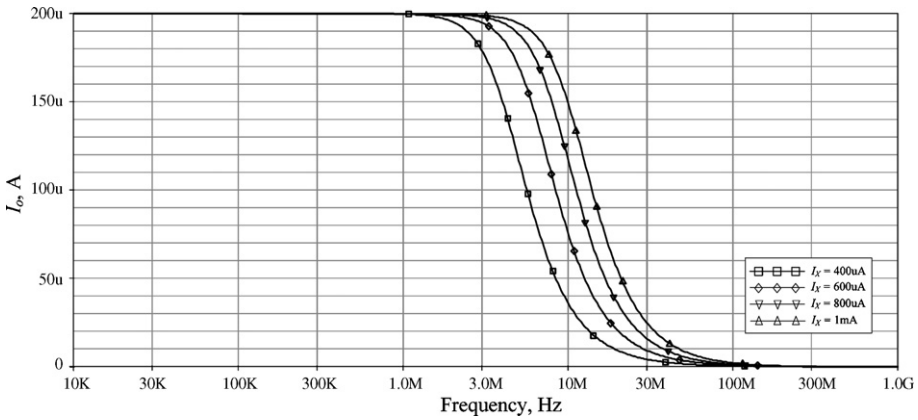

Fig. 9. Simulated frequency response of the BOTA-based low-pass filter with tuning f_0.

Table 1
The transistor dimensions.

MOSFETs	W (μm)	L (μm)
M_1–M_4, M_7–M_{12}, M_{15}–M_{33}	2.5	0.5
M_5, M_6, M_{13}, M_{14}	25	0.5

Setting $g_{mo1} = g_{mo2} = g_{mo}$ by given that $I_{X1} = I_{X2} = I_X$, then ω_0 and Q can be rewritten as

$$\omega_0 = \frac{g_{mo}}{\sqrt{C_1 C_2}} = \frac{K_T I_X}{\sqrt{C_1 C_2}} \quad \text{and} \quad Q = \sqrt{\frac{C_2}{C_1}} \tag{18}$$

Eq. (18) indicates that the filter cut-off frequency ω_0 can be electronically and linearly tuned. In addition, it is also provide the advantage that the cut-off frequency ω_0 and the quality factor Q can be orthogonally tuned. This can be achieved by firstly determine the quality factor Q by setting the capacitance ratio C_2/C_1 and then the cut-off frequency is tuned through g_{mo} by the DC bias current I_X.

4. Simulation results

The performance of the proposed BOTA is verified through the PSPICE simulation. All the CMOS transistor is simulated by using the model parameters of MOSIS 0.5 μm AMIS CMOS process with $v_{tn} = 0.67$ V and $v_{tp} = -0.91$ V. For the proposed circuit in Fig. 4, the parameter n is set to $n = 10$ and the dimensions of all CMOS transistors are listed in Table 1. The power supply voltages are set to $V_{DD} = -V_{SS} = 1.5$ V.

Fig. 6 shows the simulated transfer characteristic of the proposed BOTA. The plots show the output current i_{out} versus the input voltage v_{in} for the controlled current (I_X) in the cases of 400 μA, 600 μA, 800 μA and 1 mA, where I_B and I_Y were set to 100 μA and 400 μA, respectively. These plots demonstrate that the transconductor can linearly convert the input voltage into output signal current with nonlinearity of less than 4% for the input voltage (v_{in}) in the ranges of −200 mV_{peak} to 200 mV_{peak}. The results were agreed with the prediction value from Eq. (14). For example, in the case of the DC bias current $I_X = 1$ mA, $I_B = 100$ μA, $I_Y = 400$ μA, $v_{in} = 0.2$ V, $\mu_n C'_{ox}/2 = 5.78 \times 10^{-5}$ A/V^2 and $W/L = 10$, the transconductance gain $g_{mo} = 8.5 \times 10^{-4}$ A/V where $i_{out} = 170$ μA_{peak}.

In the Fig. 7 is the plot of the relation between the transconductance gain g_{mo} and the bias current I_X. It is measured by fixing $v_{in} = 0.2$ V_{peak}, $I_B = 100$ μA and in the case of $I_Y = 200$ μA, 300 μA and 400 μA, respectively, by varying I_X from 100 nA to 5 mA. The result shows that the transconductance gain g_{mo} can be linearly tuned by the bias current I_X over the current range of 100 nA to 1.6 mA, where the simulated conversion errors are less than 3.7%. It is also shown that the simulated and calculated data are in a good agreement. The results clearly demonstrated that the BOTA transconductance gain can be linearly and electronically tuned for a 4-decade range. Noting for the cases of $I_Y = 200$ μA, $I_Y = 300$ μA and $I_Y = 400$ μA, respectively, the tunable ranges are limited by about $I_X \approx 0.8$ mA, $I_X \approx 1.2$ mA and $I_X \approx 1.6$ mA, respectively. This confirm the condition that the current $|I_X| + |i_{in}|$ should be limited to be less than $|4I_Y|$. The achieved THD of the output current is found to be less than 1.4% for an input sinusoid with peak amplitude of 0.2 V and 50 MHz frequency. The total power consumption is 20 mW with ±1.5 V supply voltage in typical situation. The frequency response of the BOTA is also demonstrated in Fig. 8 where the −3 dB bandwidth of about 780 MHz is achieved. In addition, the CMRR of the proposed transconductor is depended on the differential CMOS transconductor [16], where $CMRR = 2g_{m(M23)}R_B g_{m(M27)}(r_{o(M29)}//r_{o(M31)})$ and R_B is

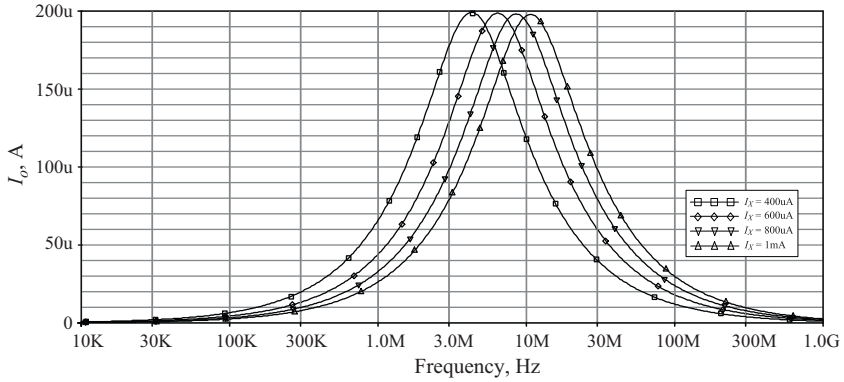

Fig. 10. Simulated frequency response of the BOTA-based band-pass filter with tuning f_0.

the output impedance of the current source. From the simulation result, the CMRR is about 90 dB.

The frequency responses of the BOTA-based second-order biquad filter, for a low-pass function and band-pass function, are shown in Figs. 9 and 10, respectively. The quality factor is kept constant to $Q = 0.707$ by chosen $C_1 = 20$ pF and $C_2 = 10$ pF and the cut-off frequency f_0 is tuned by the bias current I_X (given $I_{X1} = I_{X2} = I_{X3} = I_X$). The responses show that in the case of the input signal current is 200 μA_{peak} and $I_X = 400$ μA, 600 μA, 800 μA and 1 mA where $I_Y = 400$ μA is kept constant, the cut-off frequency (f_0) of the low-pass and band-pass filter are about 4 MHz, 6 MHz, 8 MHz and 10 MHz, respectively. In comparison with the predicted values, the f_0 that calculated from Eq. (18) using the same cases as the simulation values are at 3.8 MHz, 5.7 MHz, 7.6 MHz and 9.6 MHz, respectively. We can found that the simulation values are deviated from the calculated values with the error of less than 5%. It should be noted that the parasitic poles of CMOS BOTA influence the performances of the filter, however this results also demonstrate that the cut-off frequency can be linearly tuned by the DC bias current I_X of the transconductors without effecting the Q.

5. Conclusions

In this work, a novel circuit technique based on the combination of a transconductance cell and the new current-tunable current amplifier to realize a CMOS BOTA has been proposed. The BOTA transconductance gain g_{mo} can be electronically and linearly tuned over more than 4 decades by the external DC bias current I_X, with the transconductor's nonlinearity of less than 4%. The simulated responses with PSPICE have been quite good over the frequency range of about 780 MHz, with low power consumption. We also demonstrate that an electronically tuned current-mode biquad filter using the proposed BOTA provides the response closed to the theoretical prediction.

Acknowledgments

This work was funded by the Thailand Research Fund (TRF), under the Research Grant for New Scholar Program (Grant No. MRG5180123), and the Commission on Higher Education, the Ministry of Education, Thailand.

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