A highly linear gilbert cell OTA with multiple gated transistors for non-coherent UWB receivers

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Abstract: This paper presents a highly linear Gilbert cell OTA by cancelling the second order transconductance derivative (gm'') of nMOS transistors near threshold voltage using the technique of multiple gated transistors. This gives rise to 5 dB improvement in IIP3 of OTA, using a $0.13\ \mu\text{m}$ CMOS technology with 1.2 V supply. Simulation results and analytical relations for THD and IIP3 in OTA illustrate the tradeoffs between linearity, bandwidth, and power dissipation. System level simulation of a non-coherent UWB receiver, with this OTA as a Gm-C integrator in receiver structure, illustrates an improvement of BER from $3e^{-2}$ to $1e^{-3}$. The power consumption, IIP3, and transconductance in OTA are 0.62 mW, 20.5 dBm, and $156\ \mu\text{A/V}$, respectively.

Keywords: CMOS OTA, linearization, multiple gated transistors

Classification: Integrated circuits

References

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1 Introduction

Non-coherent receivers in wireless sensor networks require high performance modules with low complexity and low power consumption. Among these modules is the integrator block which needs an OTA with linearity and power consumption around 20 dBm and 1 mW, respectively [1]. Considering these requirements, Gilbert cell OTA appears to be superior compared to fully differential and pseudo-differential OTA structures [2]. Nevertheless, Gilbert cell requires a simple low-power technique to meet the above linearity performance.

This paper deals with the design of highly linear Gilbert cell OTA, using the technique of multiple gated transistors to cancel the second order transconductance derivative (gm'') for linearity improvement.

2 Gilbert cell OTA nonlinearity

In Gilbert OTA shown in Fig. 1, two cross-coupled differential pairs of MOS transistors in saturation are used to cancel the common-mode input voltage dependency and to make the transconductance (G_m) tunable. The transconductance of differential pair is:

$$G_m = \frac{\partial I_{out}}{\partial v_{id}} = \mu_n \cdot C_{ox} \left(\frac{W}{L} \right) (V_{CM} - V_{Th}) \quad (1)$$

where V_{CM} is the common-mode input voltage, V_{Th} is transistor threshold voltage, μ_n represents the mobility, C_{ox} denotes the oxide capacitance per unit channel area and (W/L) is transistor size.

The output transconductance of Gilbert cell is given by:

$$G_m = \frac{\partial [(I_{D1} + I_{D3}) - (I_{D2} + I_{D4})]}{\partial v_{id}} = \mu_n \cdot C_{ox} (W/L) \cdot V_b \quad (2)$$

The value of G_m is controlled through the biasing voltage V_b (see Fig. 1), which also cancels the effect of V_{CM} on G_m .

Short channel effect is the main source for nonlinearity in Gilbert OTA structure. In a short channel device, the mobility is a function of critical electric field (E_c) [3]. Considering this, the current of MOS transistor in the saturation region is:

$$I_d = \frac{1}{2} \mu_0 \cdot C_{ox} \left(\frac{W}{L} \right) (V_{GS} - V_{Th})^2 \frac{1 + \lambda V_{DS}}{1 + \theta (V_{GS} - V_{Th})} \quad (3)$$

where $\mu_{eff} = \frac{\mu_0}{1 + E/E_c}$, $\theta = \frac{1}{LE_c} + \theta_0$, L is the device channel length, λ denotes the output impedance constant, μ_0 is the low-field mobility, E represents the longitudinal electric field, and E_c is the critical electric field [4]. Assume that: $\alpha = \frac{k\lambda}{1 + \lambda V_{DC}}$, $\gamma = \frac{\theta}{1 + \theta \cdot v}$, and $K = \frac{1}{2} \mu \cdot C_{ox} \left(\frac{W}{L} \right) \frac{1 + \lambda V_{DC}}{1 + \theta \cdot v}$, where k is small signal gain, $V_{DC} = V_{DD} - V_{DS}$ and $v = V_{CM} - V_{Th}$. By substituting these in (3):

$$I_d = K \left(v + \frac{v_{id}}{2} \right)^2 \cdot \left(1 - \alpha \frac{v_{id}}{2} \right) \frac{1}{1 + \gamma \frac{v_{id}}{2}} \quad (4)$$

By using Taylor series expansion in (4) and using this in Gilbert cell structure (Fig. 1 without auxiliary transistors), the output current of Gilbert OTA is:

$$\begin{aligned} I_{out} &= (I_{D1} + I_{D3}) - (I_{D2} + I_{D4}) \\ &= K \cdot \left\{ 2 \left[(2v - V_b) - \frac{(\alpha + \gamma)}{2} \left(v^2 - \frac{(v - V_b)^2}{(1 - V_b)} \right) \right] \cdot v_{id} \right. \\ &\quad + (0)v_{id}^2 + 2 \left[\frac{1}{8} \alpha \frac{V_b}{1 - V_b} + \frac{1}{8} \gamma \cdot \frac{V_b}{1 - V_b} + \frac{1}{4} \alpha \cdot \gamma \cdot \left(v - \frac{v - V_b}{(1 - V_b)^2} \right) \right. \\ &\quad \left. \left. - \frac{1}{8} \alpha \cdot \gamma^2 \cdot \left(v^2 - \frac{(v - V_b)^2}{(1 - V_b)^3} \right) + \frac{1}{4} \gamma^2 \cdot \left(v - \frac{v - V_b}{(1 - V_b)^2} \right) - \frac{1}{8} \gamma^3 \cdot \left(v^2 - \frac{(v - V_b)^2}{(1 - V_b)^3} \right) \right] v_{id}^3 \right. \\ &\quad + (0)v_{id}^4 - \frac{1}{16} \gamma^2 \cdot \left[(\alpha + \gamma) \left(1 - \frac{1}{(1 - V_b)^3} \right) - 2\alpha \cdot \gamma \cdot \left(v - \frac{v - V_b}{(1 - V_b)^4} \right) \right. \\ &\quad \left. \left. + \gamma^2 \cdot \alpha \left(v^2 - \frac{(v - V_b)^2}{(1 - V_b)^5} \right) - 2\gamma^2 \cdot \left(v - \frac{v - V_b}{(1 - V_b)^4} \right) \right. \right. \\ &\quad \left. \left. + \gamma^3 \left(v^2 - \frac{(v - V_b)^2}{(1 - V_b)^5} \right) \right] v_{id}^5 + \dots \right\} \quad (5) \end{aligned}$$

Even harmonic distortions are canceled out in both differential and Gilbert OTA structures. Odd harmonic distortions by using $v_{id} = V_{Peak} \cdot \cos(\omega t)$ are given by:

$$HD_3(Diff) = \frac{Z}{16} V_{Peak}^2 \approx \frac{\theta}{4v \cdot (1 + \theta \cdot v)^2 \cdot (2 + \theta \cdot v)} V_{rms}^2 \quad (6)$$

$$HD_5(Diff) = \frac{\gamma^2 \cdot Z}{256} V_{Peak}^4 \approx \frac{\theta^3}{32 \cdot v \cdot (1 + \theta \cdot v)^4 \cdot (2 + \theta \cdot v)} V_{rms}^4 \quad (7)$$

$$HD_3(Gilbert) \approx \frac{Z'}{16} V_{Peak}^2 \approx \frac{\theta \cdot V_b \cdot [(1 + \theta \cdot v)^2 + 2\theta]}{4v \cdot (1 + \theta \cdot v)^2 \cdot (2 + 2\theta \cdot v - \theta \cdot V_b)} V_{rms}^2 \quad (8)$$

$$HD_5(Gilbert) \approx \frac{\gamma^2 \cdot Z''}{256} V_{Peak}^4 \approx \frac{\theta^3 \left[\left(1 - \frac{1}{(1 - V_b)^3} \right) (1 + \theta \cdot v)^2 - 2\theta \cdot V_b \right]}{32 \cdot v \cdot (1 + \theta \cdot v)^4 \cdot (2 + 2\theta \cdot v - \theta \cdot V_b)} V_{rms}^4 \quad (9)$$

$$\text{Where } Z = \frac{-\left(\frac{1}{v} - 2\gamma + \gamma^2 \cdot v\right) (\alpha + \gamma)}{\left(1 - \frac{\alpha + \gamma}{2} v\right)}, \quad Z' = \frac{-(\alpha + \gamma) \left(\frac{-V_b}{2 \cdot v} - 2\gamma \frac{V_b}{2v} + \gamma^2 \cdot V_b\right)}{\left(1 - \frac{\alpha + \gamma}{2} \cdot V_b\right)}$$

$$\text{and } Z'' = \frac{-(\alpha + \gamma) \left(\frac{\left(1 - \frac{1}{(1 - V_b)^3}\right)}{2 \cdot v} - 2\gamma \frac{V_b}{2v} + \gamma^2 \cdot V_b\right)}{\left(1 - \frac{\alpha + \gamma}{2} \cdot V_b\right)}.$$

Right sides of (6-9) are extracted by neglecting channel length modulation with $\alpha = 0$ (for $k = 1$ and $\lambda < \theta$). Right sides of (8-9) are extracted by assuming that $V_b \ll v$, $V_b \ll 1$, $(1 - V_b) \approx 1 \approx (1 - V_b)^2 \approx (1 - V_b)^3$, and $v + (v - V_b)/(1 - V_b)^2 \approx 2v - V_b \approx 2v$. From these equations, we have $HD_3 \gg HD_5$, so $THD \approx HD_3$.

The IP3 of differential OTA is:

$$IP3(Diff) = \sqrt{\frac{4}{3} \cdot \frac{\alpha_1}{\alpha_3}} = \sqrt{\frac{4}{3} \cdot \frac{1}{HD_3}} = \sqrt{\frac{64}{3Z \cdot V_{Peak}^2}} \approx \sqrt{\frac{16v \cdot (1 + \theta \cdot v)^2 \cdot (2 + \theta \cdot v)}{3\theta \cdot V_{rms}^2}} \quad (10)$$

Also, the IP3 of Gilbert OTA is:

$$IP3(Gilbert) \approx \sqrt{\frac{64}{3Z' \cdot V_{Peak}^2}} \approx \sqrt{\frac{16v \cdot (1 + \theta \cdot v)^2 \cdot (2 + 2\theta \cdot v - \theta \cdot V_b)}{3\theta \cdot V_b \cdot [(1 + \theta \cdot v)^2 + 2\theta] \cdot V_{rms}^2}} \quad (11)$$

These equations show the tradeoffs between linearity, frequency response and power consumption. In single MOS and differential pair, THD and power consumption are functions of ν . For high linearity, we should decrease θ ($\theta \propto 1/L$), which at the same time increase the parasitic capacitances, leading to narrow band frequency response. Increasing ν also improves the linearity at the expense of extra power consumption.

By comparing IP3 in (10) and (11), it can be shown that $V_b \cdot [(1 + \theta \cdot \nu)^2 + 2\theta] \ll 1$; because $V_b < \nu \ll 1$ and $\theta < 1$; consequently $IIP_3(Gilbert) \gg IIP_3(Diff)$. By comparing harmonic distortion of differential and Gilbert cell in (6) and (8), we conclude that $HD_3(Diff) \gg HD_3(Gilbert)$. Therefore, Gilbert OTA can result in a relatively high linearity.

Fig. 1. Schematic of Gilbert cell OTA with one auxiliary transistor

3 Proposed OTA with multiple gated transistors

A simple low-power solution for linearity enhancement, without using on-chip inductors, is the technique of multiple gated transistors [5]. In this technique, an auxiliary MOS is used to cancel the nonlinearity effect of the main MOSFET.

The drain current of a common source MOS is expressed by:

$$i_{ds} = g_m v_{gs} + \frac{g'_m}{2!} v_{gs}^2 + \frac{g''_m}{3!} v_{gs}^3 + \dots \quad (12)$$

where v_{gs} is gate-to-source voltage, $g_m^{(n)}$ denotes the n^{th} order g_m derivative. To reduce the power consumption, $(v_{gs}-v_{Th})$ is usually set near $0.1 \sim 0.2$ [5]. In this bias region, g_m'' has a negative peak and degrade the linearity of OTA. Negative g_m'' of the main transistor is cancelled by positive g_m'' of auxiliary transistor. Main transistor is biased at v_g , but auxiliary transistor is biased at v_g-v_{B1} (Fig. 2. d), so the transfer characteristic curve for auxiliary transistor is shifted by v_{B1} as shown in Fig. 2. b. This figure shows the ADS simulation result with $0.13 \mu\text{m}$ CMOS technology and for $v_{B1} = 0.15 \text{ V}$. The main and auxiliary transistors are designed with the same size.

The schematic of highly linear multiple gated Gilbert cell OTA is shown in Fig. 1. In this figure, M1-M4 are biased in the saturation region to satisfy (2) for G_m . The sizes of M1-M4 are the same to cancel out the common-mode voltage effect at the output. The auxiliary transistors M1_{aux}-M4_{aux} are biased in the region with positive g_m'' to cancel the negative g_m'' of the main transistor in OTA cell. PMOS transistors M5-M8 are used to provide

Fig. 2. Simulation results a) g_m of main, auxiliary FET and their summation b) third-order nonlinear currents of main transistor, one auxiliary FET, and their summation c) decreasing sensitivity of $ip3$ to bias voltage, by using two auxiliary Transistors with $(W/L)_{aux}=0.5(W/L)_{main}$ with 0.15 V and 0.3 V bias shift d) Multiple gated with one auxiliary Transistor e) Multiple gated with two auxiliary Transistors with half size

OTA biasing current, and to form the active load at the output. The amount of bias voltage (V_b) establishes a tradeoff between transconductance and linearity. A high value of V_b gives high G_m , but low IP3 as shown in (10).

4 Simulation results

This section presents the results of Advanced Design System (ADS) simulator using a $0.13\ \mu\text{m}$ CMOS technology with 1.2 V supply. The IIP3 of OTA was obtained by two-tone test with frequency separation of 100 kHz in 100 MHz. Harmonic balance (HB) and budget simulations were used for nonlinearity calculation. Simulation illustrates that by using multiple gated transistors in Gilbert OTA, the IIP3 is improved by 5 dB, giving rise to 20.5 dBm in 0.56 V bias voltage, as shown in Fig. 3 (a). This figure shows the IIP3 in OTA versus bias voltage before and after linearization. The value of θ is determined by a best fit to the simulated device characteristics for $0.13\ \mu\text{m}$ CMOS technology $\theta = 0.35 \sim 0.4\ \text{V}^{-1}$. Analytical result for $\theta = 0.38\ \text{V}^{-1}$, $V_b = 0.05$, and $v_{Th} = 0.43\ \text{V}$ is illustrated in Fig. 3. a. Therefore, by using the linearization technique we have 5 dB improvement in IIP3 for $V_g = 0.56\ \text{V}$.

As shown in Fig. 2 (b) and 3 (a), the IIP3 is sensitive to bias voltage. To decrease this sensitivity, we use two auxiliary transistors with half size and voltage shift of 0.15 V and 0.3 V. As shown in Fig. 2 (c), G_m'' summation of main and two auxiliary transistors is constant in wide range of bias voltage from 0.55 V to 0.64 V, while IIP3 is 20.5 dBm as shown in Fig. 3 (a). The power consumption of OTA with two auxiliary transistors is 0.62 mW for 0.56 V bias voltage as represented in Fig. 3 (b). Also, OTA has $156\ \mu\text{A}/\text{V}$ transconductance while drawing only 0.516 mA from 1.2 V supply. Increasing V_g also improves the transconductance at the expense of power consumption. In this case, $P_{1\text{-dB}}$ is equal to 10.18 dBm.

An Energy detection non-coherent UWB receiver modeled in IEE802.15.3.a with BPPM modulation. Transmitter composed of channel coding, source coding, modulation block, UWB pulse generator and power amplifier. Receiver consists of LNA, AGC, squarer module, integrator (I&D) and synchronizer [1]. System level simulation of a non-coherent receiver in Simulink of MATLAB with OTA as a Gm-C integrator improves BER of receiver from

Fig. 3. Simulation result a) The IIP3 of Gilbert OTA, linearization performance and theoretical result b) power consumption (mW) of proposed OTA

3e^{-2} to 1e^{-3} .

5 Conclusion

Low-power linearization with multiple gated transistors was presented for a CMOS Gilbert OTA using a 0.13 μm CMOS technology. In this linearization technique, negative $g_{m''}$ of main transistor is cancelled by positive $g_{m''}$ of auxiliary transistor. Simulations results illustrate 5 dB improvement of IIP3, giving rise to IIP3 of 20.5 dBm. The OTA dissipates 0.62 mW, and $P_{1\text{-dB}}$ (in) is 10.18 dBm for transconductance of $156 \mu\text{A/V}$. System level simulation of a non-coherent receiver with OTA as a Gm-C integrator illustrates improvement of BER from 3e^{-2} to 1e^{-3} .

Acknowledgments

This work was supported by ITRC (Iran telecommunication research center).