Electronically Tunable Current-Mode Quadrature Oscillator Using Single MCDTA

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Abstract. *This paper presents a modified current differencing transconductance amplifier (MCDTA) and a MCDTA based quadrature oscillator. The oscillator is current-mode and provides current output from high output impedance terminals. The circuit uses only one MCDTA and two grounded capacitors, and is easy to be integrated. Its oscillation frequency can be tuned electronically by tuning bias currents of MCDTA. Finally, frequency error is analyzed. The results of circuit simulations are in agreement with theory.*

Keywords

Quadrature sinusoidal oscillator, current-mode circuit, low-component count, MCDTA.

1. Introduction

Recently, current differencing transconductance amplifiers (CDTAs) have been extensively used as building blocks in signal generation applications and current-mode signal processing. The current-mode filters, oscillators, and rectifiers in literature [1] – [16] have well explained this viewpoint. However, the oscillators in [5] – [8] use exces-sive number of passive components, including floating capacitors. Thus, they are not suitable for monolithic in-tegration. The oscillators in [9] – [11] employ two ground-ed capacitors, they are resistor-less and the tuning laws for the oscillation condition and the oscillation frequency are independent. But these circuits use too many active devices, so they are not oscillators with low component count. The oscillators in [12] – [15] are overlarge due to partial utili-zation of the CDTA terminals, and there is an external linear resistor appearing at the *z* terminal of the first CDTA in Fig. 2(a) of [12] and [13], which is simulated by means of a CDTA in [13] and [14]. This is redundant, since a resistor can be easily simulated by only one OTA, rather than using a CDTA.

In this paper, we present a modified current differencing transconductance amplifier (MCDTA). The proposed MCDTA element consists of two well-known and mutually independent building blocks, namely Z-Copy CDTA and OTA. OTA is relatively independent. This is different from CDTA, MO-CDTA, and MCDTA in existing literature. As an application, the MCDTA based quadrature oscillator is given. The circuit uses only one MCDTA and two grounded capacitors, and is easy to be integrated. Its oscillation frequency can be tuned electronically by tuning bias currents of MCDTA. The outputs of the circuit possess high output impedances, and it enjoys low passive sen-sitivities. The results of circuit simulations are in agreement with theory.

2. Circuit Description

2.1 MCDTA

By adding a transconductance amplifier and a current mirror in CDTA, the number of *x* ports and *z* ports is extended, and a MCDTA is realized. It is noted that the second stage transconductance amplifier in MCDTA is relatively independent, which is different from the existing MO-CDTA. The new MCDTA circuit representation and equivalent circuit are shown in Fig. 1. Its terminal relationships can be characterized by the following set of equations

$$\begin{aligned} &V_p = V_n = 0, I_{z1} = I_{zc} = I_p - I_n, \\ &I_{x1} = g_{m1} V_{z1}, I_{x2} = g_{m2} V_{z2}, \end{aligned} \tag{1}$$

and

$$g_{m1} = \frac{I_{B1}}{2V_T}, g_{m2} = \frac{I_{B2}}{2V_T} \tag{2}$$

where I_{B1} and I_{B2} are bias currents of MCDTA, V_T is the thermal voltage and g_m is the transconductance gain of MCDTA.

2.2 Proposed Oscillator

Fig. 2 shows MCDTA-based current-mode quadrature sinusoidal oscillator. From routine analysis of the circuit in Fig. 2, the loops gain of the oscillator is obtained:

$$L(s) = \frac{sC_1 - g_{m1}}{sC_1 + g_{m1}} \times \frac{g_{m2}}{sC_2}.$$

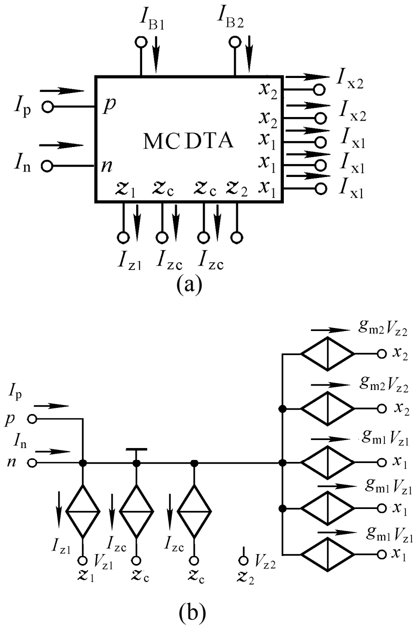

Fig. 1. MCDTA (a) symbol and (b) equivalent circuit.

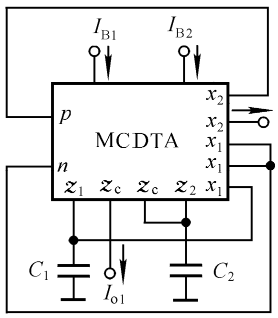

Fig. 2. MCDTA-based current-mode quadrature sinusoidal oscillator.

The characteristic equation of the oscillator can be written as

$$\Delta = 1 - L(s) = 1 - \frac{sC_1 - g_{m1}}{sC_1 + g_{m1}} \times \frac{g_{m2}}{sC_2} = 0.$$

That is

$$s^2 + s\frac{g_{m1}C_2 - g_{m2}C_1}{C_1C_2} + \frac{g_{m1}g_{m2}}{C_1C_2} = 0. \tag{3}$$

The modified oscillation condition and oscillation frequency can be obtained as

$$g_{m2}C_1 \ge g_{m1}C_2\ , \tag{4}$$

$$f_o = \frac{1}{2\pi}\sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} = \frac{1}{4\pi V_\mathrm{T}}\sqrt{\frac{I_\mathrm{B1}I_\mathrm{B2}}{C_1C_2}}. \tag{5}$$

From the circuit in Fig. 2, for sinusoidal steady state, the current transfer function from I_{o1} to I_{o2} is

$$\frac{I_{o2}}{I_{o1}} = -j\sqrt{\frac{g_{m2}C_1}{g_{m1}C_2}} = -j. \tag{6}$$

Taking (4) into account, one can conclude that the oscillator will provide quadrature signals of equal magnitudes. From (5), if $I_\mathrm{B1} = I_\mathrm{B2} = I_\mathrm{B}$ and $C_1 \ge C_2$, f_o can be tuned by adjusting bias current I_B (about this question, a simple solution has been given in the appendix). Therefore, the proposed circuit is an electronically tunable current-mode quadrature sinusoidal oscillator with low component-count.

All the active and passive sensitivities of the oscillator can be expressed as

$$S^{f_o}_{C_1,C_2} = -\frac{1}{2},\ S^{f_o}_{I_{B1},I_{B2}} = \frac{1}{2}\cdot \tag{7}$$

(7) shows that active and passive f_o sensitivities are less than unity in magnitude and hence the circuit exhibits a good sensitivity performance.

2.3 Non-Ideal Analysis

If the parasitic resistances at terminal p and n are negligible (ideally they are zero), then the parasitic impedances appearing at the x terminals would be connected between virtual grounds and actual ground and thereby eliminating their effect. Parasitic capacitances appearing at the high output impedance z_1 terminal, z_c terminal, z_2 terminal, and x_1 terminal and ground are absorbed into the external capacitors as they are shunt with them. This feature of grounded capacitor based circuits makes them par-ticularly desirable for monolithic integration. In practice, in the non-ideal case, to alleviate the effects of the parasitic impedances, the parasitic conductances and parasitic capac-itors at terminal z_1, z_c, z_2 and x_1 should be take into account.

Re-analysis of the circuit in Fig. 2 yields the following modified loops gain

$$L(s) = \frac{s(C_1+C_\mathrm{p1})+G_1-g_{m1}}{s(C_1+C_\mathrm{p1})+G_1+g_{m1}} \times \frac{g_{m2}}{s(C_2+C_\mathrm{p2})+G_2}.$$

where G_1 denotes parasitic conductance in parallel with C_1, G_2 denotes parasitic conductance in parallel with C_2, C_p1 and C_p2 denote parasitic capacitors in parallel with C_1 and C_2, respectively.

The modified characteristic equation of the oscillator is

$$\begin{aligned}\Delta &= 1 - L(s)\\ &= 1 - \frac{s(C_1+C_\mathrm{p1})+G_1-g_{m1}}{s(C_1+C_\mathrm{p1})+G_1+g_{m1}} \times \frac{g_{m2}}{s(C_2+C_\mathrm{p2})+G_2} = 0.\end{aligned}$$

That is

$$\begin{aligned}&s^2 + \frac{s[(G_2-g_{m2})(C_1+C_\mathrm{p1})+(G_1+g_{m1})(C_2+C_\mathrm{p2})]}{(C_1+C_\mathrm{p1})(C_2+C_\mathrm{p2})}\\ &+\frac{G_1G_2+G_2g_{m1}-G_1g_{m2}+g_{m1}g_{m2}}{(C_1+C_\mathrm{p1})(C_2+C_\mathrm{p2})} = 0.\end{aligned} \tag{8}$$

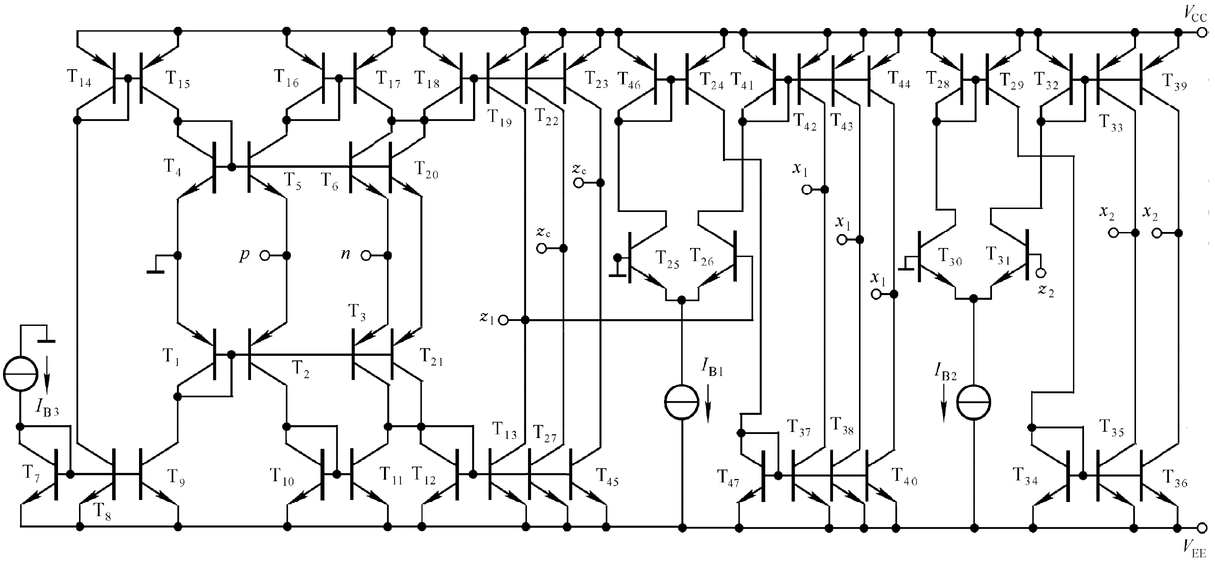

Fig. 3. Internal construction of MCDTA.

The modified oscillation condition is

$$g_{m2}(C_1+C_{p1}) \ge (G_1+g_{m1})(C_2+C_{p2})+G_2(C_1+C_{p1})\cdot \tag{9}$$

The modified oscillation frequency is

$$f_o' = \frac{1}{2\pi}\sqrt{\frac{G_1G_2+G_2g_{m1}-G_1g_{m2}+g_{m1}g_{m2}}{(C_1+C_{p1})(C_2+C_{p2})}}. \tag{10}$$

Assuming $g_{m1} = g_{m2} = g_m$, ignoring the second-order infinitesimal G_1G_2 and $C_{p1}C_{p2}$, and using $\sqrt{1+x} \approx 1+x/2$, for $|x| << 1$, (10) becomes

$$f_o' \approx f_o(1-\frac{G_1-G_2}{2g_m}-\frac{C_{p1}}{2C_1}-\frac{C_{p2}}{2C_2})\cdot \tag{11}$$

(9) shows that if $g_{m1} = g_{m2} = g_m$, the modified oscillation condition changes into

$$C_1 \ge (C_2+C_{p2})(G_1+g_{m1})/(g_{m2}-G_2)-C_{P1},$$

hence, C_1 must be slightly greater than C_2 due to non-ideal factors. (11) shows that that if $g_{m1} = g_{m2} = g_m$, the modified oscillation frequency changes into

$$f_o' = f_o\,[1-(G_1-G_2)/2g_m-C_{p1}/2C_1-C_{p2}/2C_2],$$

therefore, taking into account non-ideal factors, the oscillation frequency becomes smaller.

From (11), the deviation for oscillation frequency is

$$\frac{f_o'-f_o}{f_o} = -\frac{G_1-G_2}{2g_m}-\frac{C_{p1}}{2C_1}-\frac{C_{p2}}{2C_2}. \tag{12}$$

The sensitivity study of the modified oscillation frequency indicates that

$$S_{C_1}^{f_o'} = -\frac{1}{2(1+C_{p1}/C_1)}, \quad S_{C_2}^{f_o'} = -\frac{1}{2(1+C_{p2}/C_2)},$$

$$S_{C_{p1}}^{f_o'} = -\frac{1}{2(1+C_1/C_{p1})}, \quad S_{C_{p2}}^{f_o'} = -\frac{1}{2(1+C_2/C_{p2})},$$

$$S_{g_{m2}}^{f_o'} = S_{I_{B2}}^{f_o'} = \frac{-G_1g_{m2}}{2(G_1G_2+G_2g_{m1}-G_1g_{m2}+g_{m1}g_{m2})},$$

$$S_{g_{m1}}^{f_o'} = S_{I_{B1}}^{f_o'} = \frac{g_{m1}g_{m2}}{2(G_1G_2+G_2g_{m1}-G_1g_{m2}+g_{m1}g_{m2})},$$

$$S_{G_1}^{f_o'} = \frac{G_1(G_2-g_{m2})}{2(G_1G_2+G_2g_{m1}-G_1g_{m2}+g_{m1}g_{m2})},$$

$$S_{G_2}^{f_o'} = \frac{G_2(G_1+g_{m1})}{2(G_1G_2+G_2g_{m1}-G_1g_{m2}+g_{m1}g_{m2})}. \tag{13}$$

From the above expressions, it is seen that all passive and active sensitivities of the circuit are low.

3. Simulation Results

In order to test the performances of the proposed circuit, the MO-CCCDTA of literature [17] is modified to MCDTA of Fig. 3, and the sub-circuit for MCDTA is created on transistor QNL(BF = 100) and QPL(BF = 100) by ELECTRONICS WORKBENCH 5.0 software (EWB 5.0), then Fig. 2 is created. Finally, the Fig. 2 circuit is simulated with ± 1.5 V power supplies, $C_1 = 1.08$ nF, $C_2 = 1$ nF, $I_{B1} = I_{B2} = I_B$ =25 μA, and $I_{B3} = 1000$ μA, which yields parasitic resistor $R_p = R_n = V_T/2I_{B3} = 13\ \Omega$, hence, terminals p and n are at virtual ground. The simulation results are shown in Fig. 4. This shows that the circuit really created a quadrature sinusoidal oscillation. Using (5) yields the design value for oscillation frequency f_o = 73.6657 kHz; using the pointer in EWB5.0 yields the actual value for oscillation frequency f_o' = 72.1671 kHz, so the deviation for oscillation frequency is (72.1671 – 73.6657) / 73.6657 = -2.03 %.

To illustrate the controllability of the oscillation frequency by adjusting I_B, let $C_1 = C_2 = 1$ nF, $I_{B1} = I_{B2} = I_B = 50$ μA, $I_{B3} = 1000$ μA. Simulation result of the

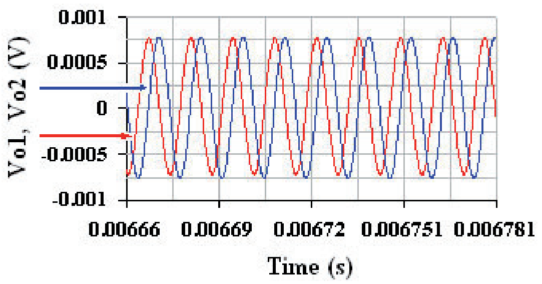

Fig.4. The steady-state waveforms for proposed circuit when $C_1 = 1.08$ nF, $C_2 = 1$ nF, $I_B = 25$ µA, R_L=100 Ω.

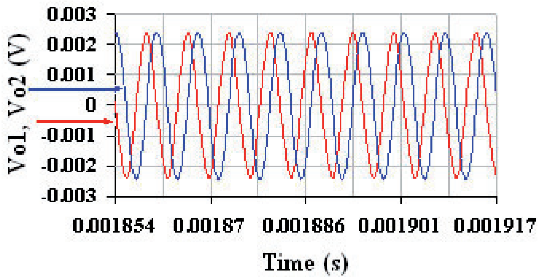

Fig.5. The steady-state waveforms for proposed circuit when $C_1 = C_2 = 1$ nF, $I_B = 50$ µA, $R_L = 100$ Ω.

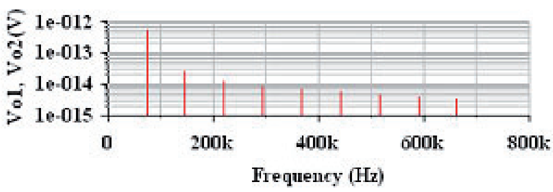

Fig. 6. Simulation result of the output spectrum when $C_1 = 1.08$ nF, C2 = 1 nF, IB = 25 µA, RL = 100 Ω.

quadrature outputs is shown in Fig. 5. Theoretically, for the ideal case, the design value for oscillation frequency is 153.111 kHz. Using the pointer in EWB 5.0, it has been observed that the actual value for oscillation frequency is 143.950 kHz, so the deviation for oscillation frequency is (143.95 – 153.111) / 153.111= -5.98 %. The error is mainly due to parasitic impedances appearing in z, g, and x ter-minals. From Fig. 3, we can obtain $G_1 = G_z + G_g + G_x$, $G_2 = G_z + G_g$, of them, G_z denotes output conductance of current differencing circuit of MCDTA, G_g denotes input conductance of the tranconductance amplifier of MCDTA, G_x denotes output conductance of output current mirror of MCDTA. Using the frequency analysis in EWB 5.0, we receive $G_1 = 0.0426$ ms, $G_2 = 0.0416$ ms, and the corresponding parasitic capacitance $C_{p1} = 0.0644$ nF, $C_{p2} = 0.056$ nF. Substituting these data into (12) gives that the relative error is -6.1 %. It is noted that the results of circuit simulations are in agreement with theory.

Fig. 6 shows the simulated output spectrum, where the total harmonic distortion is about 3.16562 %. The dis-tortion is due to the fact that the nonlinear elements (such as diode, thermal resistor etc) are not used to stabilize the amplitude of oscillation. In fact, the oscillations grow exponentially in amplitude until the linear dynamic range of two OTAs in MCDTA is exceeded. When two OTAs approach their saturation regions, respectively, the oscil-lator can sustain an output signal with small distortion.

4. Conclusions

In this paper, a new MCDTA is presented. It has two parameters controlled electronically, the I_{B1} and I_{B2}. A MCDTA-based current-mode quadrature sinusoidal os-cillator with high output impedance was realized. It uses relatively few components and grounded capacitors. Its oscillation frequency can be tuned by varying the bias current of MCDTA. Although the oscillation condition for proposed oscillator cannot be adjusted electronically, this topology can be introduced as an economical oscillator. The proposed circuit structure is expected to be useful for applications in communications, instrumentation and meas-urement systems, especially at a high frequency range.

Appendix

There is a mechanism by which I_{B1} and I_{B2} can be varied together. Fig. 7 shows a simple implementation. This is a multi-transistor current mirror. The relationship between each load current and the reference current, as-suming all transistors are matched and Early voltage $V_A = \infty$, is

$$I_{B1} = I_{B2} = I_R = \frac{V_C - V_{EE} - V_{BE}}{R}.$$

It is apparent that if the circuit parameters R, V_{EE}, and V_{BE} are given, the load currents I_{B1} and I_{B2} can be varied simultaneously by varying an external control voltage V_C.

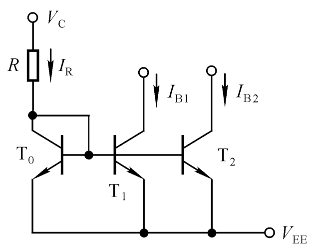

Fig. 7. Multi-transistor current mirror.

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