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# Current-Mode Multiphase Sinusoidal Oscillator Using Current Differencing Transconductance Amplifiers

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**Abstract** This paper describes a simple current-controlled current-mode multiphase sinusoidal oscillator based on current differencing transconductance amplifiers (CDTAs) as active components. The proposed oscillator circuit, which employs only one CDTA and one grounded capacitor for each phase, can generate arbitrary $n$ current-output signals ($n$ being even or odd) equally spaced in phase, all at high output impedance terminals. The oscillation condition and the oscillation frequency can be controlled electronically and independently through the bias current of the CDTA. The oscillator has low-component count, low-sensitivity performance, and is highly suitable for monolithic implementation. PSPICE simulation results are given to confirm the operation of the proposed oscillator.

**Keywords** CDTA · Multiphase sinusoidal oscillator (MSO) · Current-mode circuit

## 1 Introduction

Multiphase sinusoidal oscillators (MSOs) have a wide range of applications in telecommunications, power electronics, and signal processing and measurement systems. Examples of commonly used MSO circuits are the vector control of single-phase-to-three-phase pulsewidth-modulation (PWM) converters [19], and a control scheme for a five-phase induction motor drive [20]. They can also be utilized for

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a decoupled dynamic control of a six-phase two-motor drive system [16]. Therefore, several techniques for the design of the MSO have been reported in the literature [1–4, 13–15, 23, 24]. In [2, 4, 15, 23], several MSO circuits have been proposed which use a second-generation current conveyor (CCII) as an active component. Most of the reported circuits suffer from the use of more passive components and a lack of electronic controllability. Also, they operate in voltage mode. The current feedback operational amplifier (CFOA)-based MSO circuit proposed in [24] exploits the internal pole of the device to operate at relatively high frequencies, but this approach requires access to the device compensation terminal. Recently, techniques to realize a voltage-mode MSO using operational amplifiers (op-amps) have been developed [13, 14]. However, the drawback of these circuits is the well-known limitations of the op-amps. Moreover, they utilize too many external passive components and a number of them float. In recent years, the current-mode approach to signal processing has offered elegant solutions for analog circuit problems [5, 22]. The main advantages of this operation mode are wide signaling bandwidth, high slew rate and low power consumption. Thus considering this fact, a current follower-based MSO structure operating in current mode has been proposed [1]. It employs two current followers, one floating resistor and one floating capacitor for each stage, and does not provide electronic tunability. More recently, a current-mode MSO circuit based on current-controlled conveyors (CCCIIs) has been reported [3]. This method utilizes the parasitic resistance ($R_x$) of the conveyor, making electronic tunability possible through the bias current. However, it still requires an excessive number of external passive capacitors, i.e., two grounded capacitors for each stage. Moreover, its oscillation condition is adjusted by tuning the ratio of external passive capacitors, which is not well controlled. In integrated circuits (ICs), controlling the circuit parameters electronically is much easier to realize than changing the values of passive components.

In 2003, a new current-mode active element with two current inputs and two kinds of current outputs, which is called a current differencing transconductance amplifier (CDTA), was introduced [8]. Owing to the current conveying property, the CDTA is one of the modifications of the CCII. This device is synthesized using the current differencing nature of the modified differential current conveyor (MDCC) [11] and a multiple-output transconductance amplifier to facilitate the implementation of current-mode analog signal processing. As a result, many applications and advantages in the design of various current-mode circuits using CDTAs as active elements have received considerable attention. They include, i.e., universal biquad filters [9], an $N$th-order lowpass filter circuit [6], a current-mode universal 2nd-order filter with one input and three outputs [10], and a current-mode KHN filter [18]. These applications have also proven that the CDTA is a versatile active building block for current-mode signal processing. Until now, only a current-mode quadrature oscillator circuit based on CDTAs has been recently reported [17]. However, it produces two sinusoidal output currents with only 90° phase difference. There is no CDTA-based current-mode sinusoidal oscillator that can provide multiphase sinusoidal output currents.

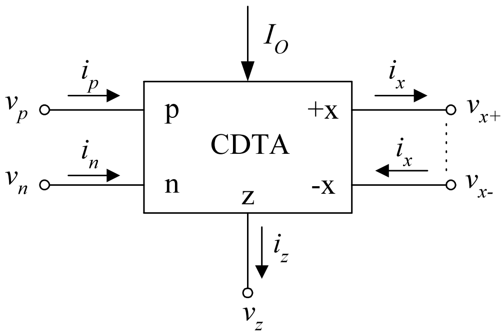
In view of the above limitations in previous works, we propose a current-controlled current-mode MSO circuit using CDTAs and only grounded capacitors. The proposed oscillator circuit can produce $n$ output current signals ($n$ being even or odd) of identical frequency and equally spaced in phase, all at high output impedance terminals. The circuit provides the attractive feature of independent electronic control of the oscillation frequency ($\omega_o$) and the oscillation condition by varying the bias current of the CDTA. The oscillator also exhibits low active and passive sensitivities and is very suitable for integration. PSPICE simulation results are given demonstrating close agreement with the predicted theory.

## 2 Current Differencing Transconductance Amplifier (CDTA)

The circuit representation and the equivalent circuit of the CDTA are shown in Fig. 1. The terminal relation of the CDTA can be characterized by the following set of equations [8]:

$$v_p = v_n = 0, \qquad i_z = i_p - i_n \quad \text{and} \quad i_x = g_m v_z = g_m Z_z i_z \tag{1}$$

where $p$ and $n$ are input terminals, $z$ and $\pm x$ are output terminals, $g_m$ is the transconductance gain, and $Z_z$ is an external impedance connected at the terminal $z$. According to the above equation and the equivalent circuit of Fig. 1b, the current flowing



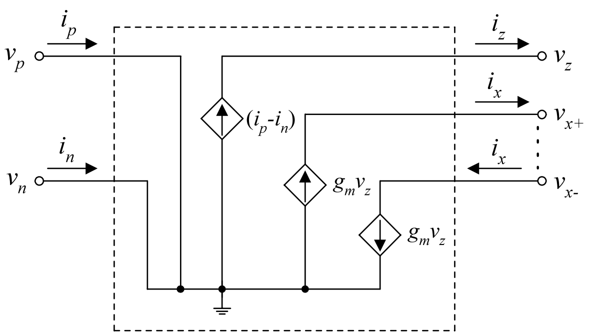




**Fig. 1** CDTA: **a** circuit representation, **b** equivalent circuit

out of the terminal $z$ ($i_z$) is a difference between the currents through the terminals $p$ and $n$ ($i_p - i_n$). The voltage drop at the terminal $z$ is transferred to a current at the terminal $x$ ($i_x$) by a transconductance gain ($g_m$), which is electronically controllable by an external bias current ($I_{\mathrm{O}}$). These currents, which are copied to a general number of output current terminals $x$, are equal in magnitude but flow in opposite directions.

Although there are several techniques to realize the CDTA, one possible bipolar realization is shown in Fig. 2 [21]. It mainly comprises a current differencing circuit formed by two current followers $Q_1$–$Q_9$, a basic current mirror $Q_{10}$–$Q_{11}$, and a multiple-output transconductance amplifier $Q_{12}$–$Q_{42}$. In this case, the transconductance gain $g_m$ of the CDTA is directly proportional to the external bias current $I_{\mathrm{O}}$, which can be written as

$$g_m = \frac{I_{\mathrm{O}}}{2V_{\mathrm{T}}} \tag{2}$$

where $V_{\mathrm{T}} \cong 26$ mV at 27°C is the thermal voltage.

## 3 Circuit Descriptions

In this section, two basic building blocks used for realizing the proposed current-mode MSO scheme are discussed in detail. The current-mode lossy integrator and current amplifier using CDTAs are described as follows.

### 3.1 CDTA-Based Current-Mode Lossy Integrator

Figure 3 shows the CDTA-based current-mode lossy integrator using a grounded capacitor. The basic current transfer function of this circuit can be given by

$$\frac{i_{\mathrm{o}}}{i_{\mathrm{in}}} = \frac{1}{1 + s(\frac{C}{g_m})}. \tag{3}$$

This building block will be used as a core circuit in a generalized current-mode MSO structure.

### 3.2 CDTA-Based Current Amplifier

The second basic building block, shown in Fig. 4, functions as a current amplifier. It is formed of only two CDTAs without the need for an external passive component. In this case, the current gain ($K$) of this circuit can be expressed as

$$\frac{i_{\mathrm{o}}}{i_{\mathrm{in}}} = K = \frac{g_{ma}}{g_{mb}}. \tag{4}$$

From (4), it is worth mentioning here that the amplifier gain $K$ can be tuned linearly and electronically by adjusting the transconductance ratio ($g_{ma}/g_{mb}$).

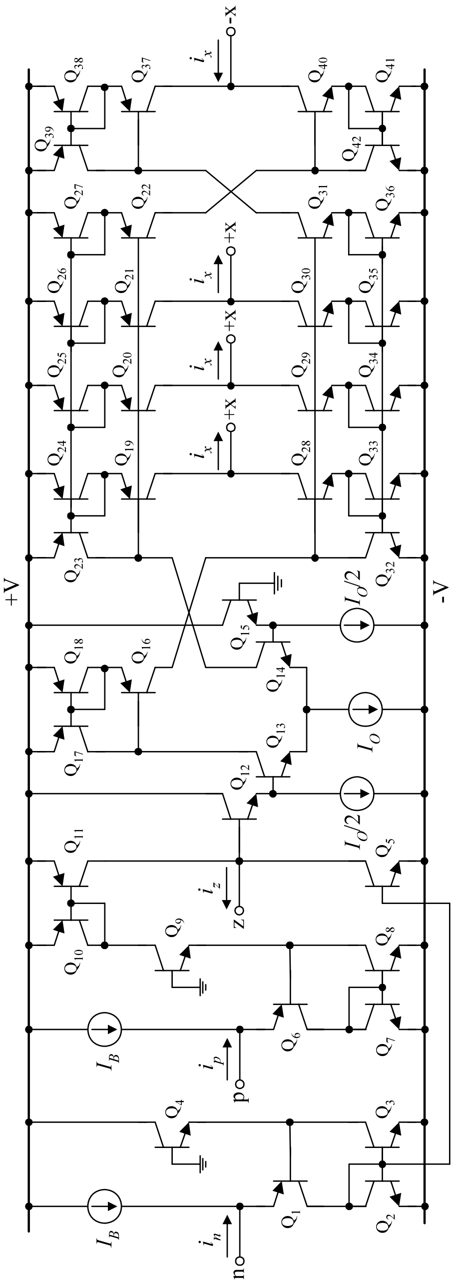



**Fig. 2** Possible bipolar realization of the CDTA

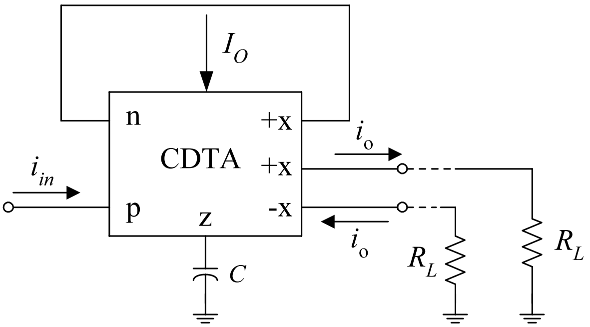



**Fig. 3** CDTA-based current-mode lossy integrator

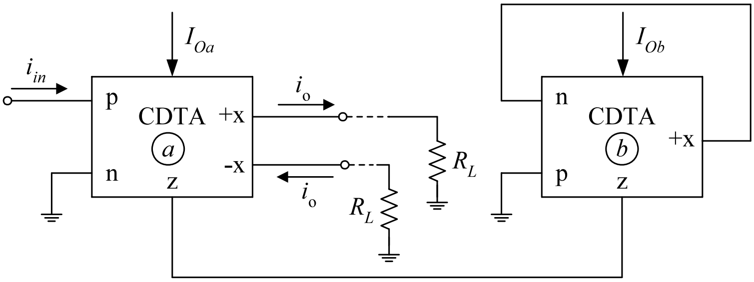



**Fig. 4** CDTA-based current amplifier

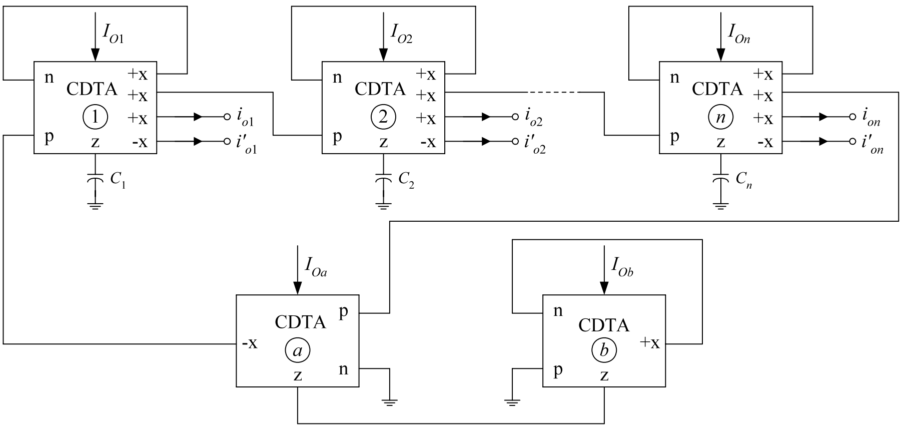



**Fig. 5** Proposed CDTA-based MSO circuit using grounded capacitors

### 3.3 Proposed CDTA-Based Current-Mode MSO Circuit

The realization of the proposed current-mode MSO structure using CDTAs is shown in Fig. 5. The circuit consists of $n$ cascaded CDTA-based current-mode lossy integrators of Fig. 3 with the output $i_{on}$ of the $n$th stage being fed back to the input of the first stage through the current amplifier of Fig. 4. Note that the current amplifier performing a feedback path has a gain of $-K$. From the circuit configuration, we see that it requires only CDTAs and grounded capacitors and is suitable from an integration point of view [7]. Assuming that $g_{m1} = g_{m2} = \cdots = g_{mn} = g_m$ and $C_1 = C_2 = \cdots = C_n = C$, the open loop gain $L(s)$ of the proposed structure in Fig. 5 can be given by

$$L(s) = -K\left[\frac{1}{1+s(\frac{C}{g_m})}\right]^n = -\left(\frac{g_{ma}}{g_{mb}}\right)\left[\frac{1}{1+s(\frac{C}{g_m})}\right]^n. \tag{5}$$

To sustain sinusoidal oscillations at frequency $\omega_o = 2\pi f_o$, the Barkhausen criteria must be satisfied such that

$$L(j\omega_o) = -\frac{(\frac{g_{ma}}{g_{mb}})}{[1+s(\frac{C}{g_m})]^n} = 1 \tag{6}$$

and

$$\phi = \frac{\pi}{n} = \tan^{-1}\left(\frac{\omega_o C}{g_m}\right) \tag{7}$$

where $\phi$ is the phase shift of each current integrator. It should be noted that (6) will have a solution only if the value of $n$ is odd ($n \geq 3$). From (7), there are $n$ outputs $i_{oi}$ ($i = 1, 2, \ldots, n$), each shifted in phase by $180°/n$, available from the scheme. By rearranging (6) and equating the imaginary and real parts to zero, the condition for oscillation and the frequency of oscillation ($\omega_o$) for values of $n = 3, 5, 7, \ldots$ can respectively be given as

$$\frac{g_{ma}}{g_{mb}} = \left[1+\left(\frac{\omega_o C}{g_m}\right)^2\right]^{\frac{n}{2}} \tag{8}$$

and

$$\omega_o = \left(\frac{g_m}{C}\right)\tan\left(\frac{\pi}{n}\right). \tag{9}$$

By substituting (2) and (9) into (8), the oscillation condition can now be rewritten as

$$\frac{I_{Oa}}{I_{Ob}} = \left[1+\tan^2\left(\frac{\pi}{n}\right)\right]^{\frac{n}{2}} \tag{10}$$

where $I_{Oa}$ and $I_{Ob}$ are the external DC bias currents of CDTA$_a$ and CDTA$_b$, respectively. Also, substituting (2) into (9) gives

$$\omega_o = \left(\frac{I_O}{2V_T C}\right)\tan\left(\frac{\pi}{n}\right) \tag{11}$$

From (10) and (11), it is clearly seen that $\omega_o$ can be tuned electronically and linearly by adjusting the bias current $I_O$ without affecting the condition of oscillation, which can also be tuned electronically by adjusting the bias current ratio $I_{Oa}/I_{Ob}$ without influencing $\omega_o$. This means that the circuit provides the attractive feature of independent linear current control of the frequency and the condition of oscillation. In addition, by the use of an inverted version of the output current of the CDTA, the even-phase ($2n = 6, 10, 14, \ldots$) output currents ($i'_{o1}, i'_{o2}, \ldots, i'_{on}$) are also obtained from the same structure. Therefore, at $\omega_o$, the proposed configuration of Fig. 5 can realize an $n$th-order ($n$ = odd/even integer) MSO. Table 1 summarizes the oscillation condition and $\omega_o$ for realizing an $n$-phase current-mode sinusoidal oscillator. From Table 1, it is easy to see that the active and passive sensitivities of $\omega_o$ are expressed as

$$S_{I_o}^{\omega_o} = -S_C^{\omega_o} = 1. \tag{12}$$

Thus, the proposed MSO configuration has low active and passive sensitivities.

### 3.4 Effects of CDTA Non-idealities

In this section, the effects of CDTA non-idealities on the oscillator characteristics will be investigated. Figure 6 shows the simplified equivalent circuit that will be used to represent the behavior of the non-ideal CDTA. As is seen, there are parasitic resistances ($R_p$ and $R_n$) at terminals $p$ and $n$, and parasitic resistances and capacitances ($R_z$, $C_z$ and $R_x$, $C_x$) from terminals $z$ and $x$ to the ground. In the same figure, $\alpha_p = 1 - \varepsilon_p$, $|\varepsilon_p| \ll 1$ is the current transfer error from the $p$ to $z$ terminals, $\alpha_n = 1 - \varepsilon_n$, $|\varepsilon_n| \ll 1$ is the current transfer error from the $n$ to $z$ terminals, and $\beta$

**Table 1** Frequency and condition of oscillation of the proposed MSO of Fig. 5

| Number of phases ($n$) | Condition of oscillation | Frequency of oscillation ($\omega_o$) |
|---|---|---|
| 3 | $I_{Oa} = 8I_{Ob}$ | $\omega_o \cong 33.308(I_O/C)$ |
| 4 | $I_{Oa} = 4I_{Ob}$ | $\omega_o \cong 19.230(I_O/C)$ |
| 5 | $I_{Oa} \cong 2.885I_{Ob}$ | $\omega_o \cong 13.972(I_O/C)$ |
| 6 | $I_{Oa} \cong 2.370I_{Ob}$ | $\omega_o \cong 11.103(I_O/C)$ |
| 7 | $I_{Oa} \cong 2.075I_{Ob}$ | $\omega_o \cong 9.261(I_O/C)$ |
| 8 | $I_{Oa} \cong 1.883I_{Ob}$ | $\omega_o \cong 7.965(I_O/C)$ |
| 9 | $I_{Oa} \cong 1.750I_{Ob}$ | $\omega_o \cong 7(I_O/C)$ |

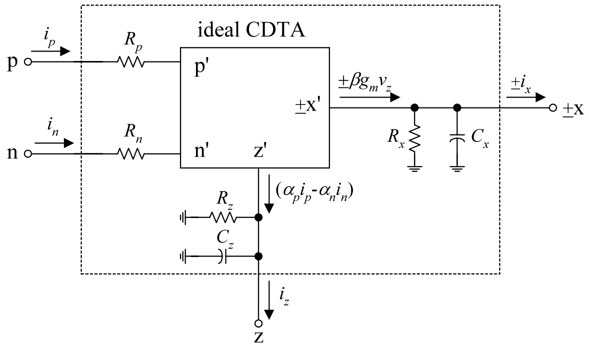



**Fig. 6** Simplified equivalent circuit of the non-ideal CDTA

is the transconductance inaccuracy factor from the $z$ to $x$ terminals. Therefore, taking into account the non-ideal CDTA characteristics and its parasitic elements, the modified current transfer functions of Figs. 3 and 4 can respectively be rewritten as

$$\frac{i_{\mathrm{o}}}{i_{\mathrm{in}}} = \frac{(\frac{\alpha_p}{\alpha_n})}{(1 + \frac{1}{\alpha_n \beta g_m R_z} + \frac{sC}{\alpha_n \beta g_m})} \tag{13}$$

and

$$\frac{i_{\mathrm{o}}}{i_{\mathrm{in}}} = K = \frac{(\frac{\alpha_p}{\alpha_n})(\frac{g_{ma}}{g_{mb}})}{(1 + \frac{2sC_z}{\alpha_n \beta g_{mb}})(1 + sR_x C_x)}. \tag{14}$$

Re-analysis of the proposed MSO circuit in Fig. 5 using (13) and (14) with $C \gg C_z, C_x$ yields the oscillation condition and $\omega_o$ as described by the following relations:

$$\left(\frac{g_{ma}}{g_{mb}}\right)\left(\frac{\alpha_p}{\alpha_n}\right)^{n+1} = \left[\left(1 + \frac{1}{\alpha_n \beta g_m R_z}\right)^2 + \tan^2\left(\frac{\pi}{n}\right)\right]^{\frac{n}{2}} \tag{15}$$

and

$$\omega_{\mathrm{o}} = \left(\frac{\alpha_n \beta g_m + \frac{1}{R_z}}{C}\right) \tan\left(\frac{\pi}{n}\right). \tag{16}$$

If $\alpha_n \beta g_m$ is sufficiently higher than $1/R_z$, then (15) and (16) can be approximated to

$$\left(\frac{g_{ma}}{g_{mb}}\right)\left(\frac{\alpha_p}{\alpha_n}\right)^{n+1} \cong \left[1 + \tan^2\left(\frac{\pi}{n}\right)\right]^{\frac{n}{2}} \tag{17}$$

and

$$\omega_{\mathrm{o}} \cong \left(\frac{\alpha_n \beta g_m}{C}\right) \tan\left(\frac{\pi}{n}\right). \tag{18}$$

From the above expressions, it should be mentioned here that the oscillation condition is mainly affected by the current transfer errors ($\alpha_p$ and $\alpha_n$) of the CDTAs. However, the effect of these errors can be easily accommodated by tuning the ratio of $g_{ma}/g_{mb}$. Also, one can see that $\omega_{\mathrm{o}}$ is slightly deviated from the ideal case by the factor of $\alpha_n \beta$. Therefore, to compensate this effect, we can slightly adjust the $g_m$-value.

From (18), the active $\omega_{\mathrm{o}}$-sensitivity values in this case can be expressed as

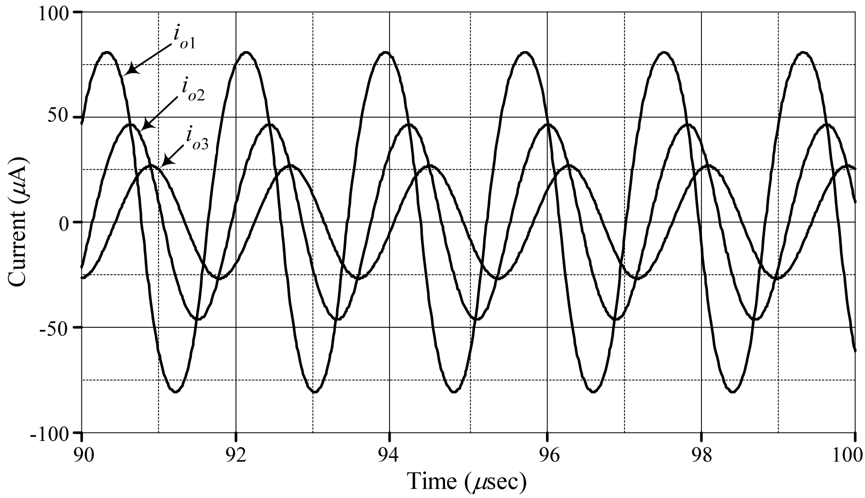
$$S_{\alpha_p}^{\omega_{\mathrm{o}}} = 0 \tag{19}$$

and

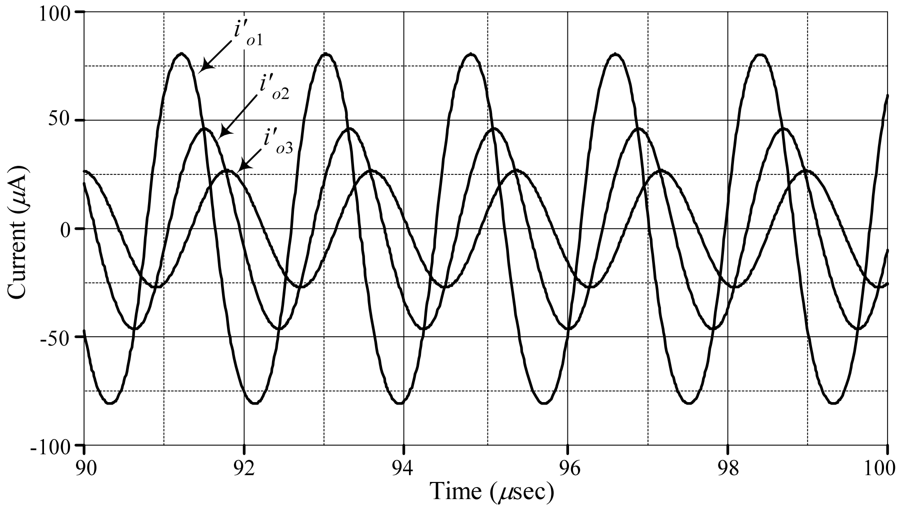
$$S_{\alpha_n}^{\omega_{\mathrm{o}}} = S_{\beta}^{\omega_{\mathrm{o}}} = 1. \tag{20}$$

## 4 Simulation Results

To confirm the given theoretical analysis, the proposed current-mode MSO circuit in Fig. 5 has been simulated with the PSPICE program. The CDTAs were realized by the schematic bipolar implementation given in Fig. 2 with the transistor model




(a)




(b)

**Fig. 7** Simulated output waveforms of the proposed current-mode MSO of Fig. 5: **a** $i_{o1}$, $i_{o2}$, $i_{o3}$, **b** $i'_{o1}$, $i'_{o2}$, $i'_{o3}$

parameters of PR100N (PNP) and NP100N (NPN) of the bipolar arrays ALA400 from AT&T [12]. The supply voltages are $+V = -V = 3$ V and the values of the bias currents are $I_B = 200$ µA.

As an example for $n = 3$, the output waveforms and frequency spectra obtained from the proposed MSO structure of Fig. 5 are respectively shown in Figs. 7 and 8

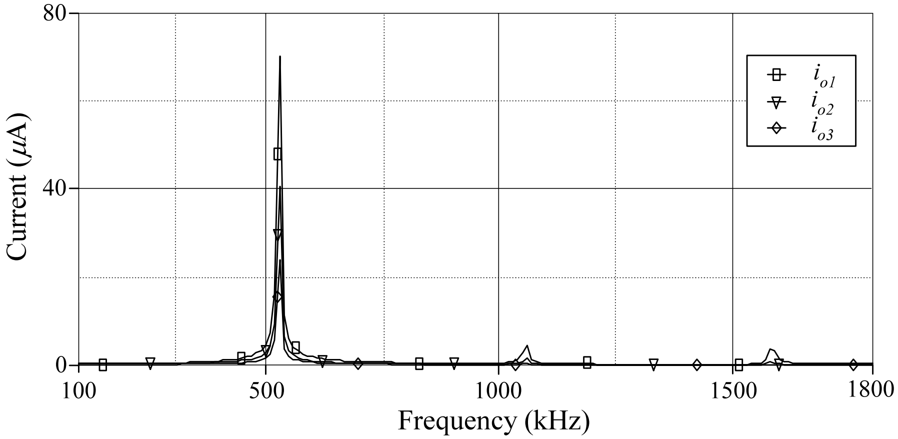



**Fig. 8** Simulated frequency spectra of $i_{\mathrm{o}1}$, $i_{\mathrm{o}2}$ and $i_{\mathrm{o}3}$

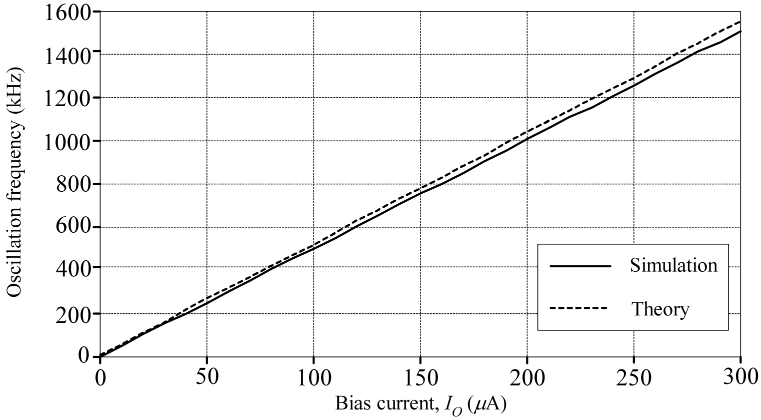



**Fig. 9** Variations of the oscillation frequency $(f_{\mathrm{o}})$ with the bias currents $I_{\mathrm{O}}$

with $I_{\mathrm{O}} = I_{\mathrm{O}i}$ $(i = 1, 2, 3) = 100$ μA and $C = 1$ nF. The simulated oscillation frequency was found to be 500 kHz, while the calculated frequency using (11) is about 530 kHz. Also from the simulations, the phase difference of $i_{\mathrm{o}2}$, $i_{\mathrm{o}3}$, $i'_{\mathrm{o}1}$, $i'_{\mathrm{o}2}$ and $i'_{\mathrm{o}3}$ comparing with $i_{\mathrm{o}1}$ were measured as: 61°, 116°, 179°, 238° and 296°, respectively. The total harmonic distortion (THD) in the output waveforms $i_{\mathrm{o}1}$, $i_{\mathrm{o}2}$ and $i_{\mathrm{o}3}$ were approximated to 5.6%, 2.3% and 1.3%, respectively. From both figures, it can be seen that the amplitudes of the $n$ output signals are not the same. For applications requiring equal amplitude outputs, some current amplifier circuits are needed.

The tunability of $\omega_{\mathrm{o}}$ through $I_{\mathrm{O}}$ without influencing the oscillation condition for the predicted theory is illustrated in Fig. 9. It is obvious that the simulated oscillation frequency values are in good agreement with the theoretical values within a wide frequency range. The simulation results of the oscillation frequency for three different values of $C$ (i.e. 1 nF, 10 nF and 100 nF) are summarized in Table 2, which shows the

**Table 2** Oscillation frequency ($f_O$) of the proposed MSO of Fig. 5 for various values of $C$

| $I_O$ (μA) | $f_O$ (kHz) | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|
| | $C = 1$ nF | | | $C = 10$ nF | | | $C = 100$ nF | | |
| | Theory | Simulation | Error (%) | Theory | Simulation | Error (%) | Theory | Simulation | Error (%) |
| 10 | 53 | 50 | 5.66 | 5.3 | 5 | 5.66 | 0.53 | 0.50 | 5.66 |
| 50 | 265 | 250 | 5.66 | 26.5 | 25 | 5.66 | 2.65 | 2.50 | 5.66 |
| 100 | 530 | 500 | 5.66 | 53 | 50 | 5.66 | 5.3 | 5 | 5.66 |
| 150 | 795 | 750 | 5.66 | 79.5 | 70 | 5.66 | 7.95 | 7.50 | 5.66 |
| 200 | 1060 | 1000 | 5.66 | 106 | 95 | 10.37 | 10.60 | 10 | 5.66 |
| 250 | 1325 | 1200 | 9.43 | 132.5 | 115 | 13.20 | 13.25 | 11.50 | 13.20 |
| 300 | 1590 | 1500 | 5.66 | 159 | 140 | 11.94 | 15.90 | 14 | 11.94 |
| 350 | 1855 | 1740 | 6.19 | 185.5 | 170 | 8.35 | 18.55 | 16.50 | 11.05 |
| 400 | 2120 | 1980 | 6.60 | 212 | 190 | 10.37 | 21.20 | 19 | 10.37 |
| 450 | 2385 | 2200 | 7.75 | 238.5 | 205 | 14.04 | 23.85 | 20.50 | 14.04 |
| 500 | 2650 | 2400 | 9.43 | 265 | 230 | 13.20 | 26.50 | 23 | 13.20 |

linearity over a wide range of the bias current $I_O$. From Table 2, it can be observed that the frequency error is less than 15% up to about 2.65 MHz. However, reducing this error can be easily done by simply controlling the bias current of the CDTAs.

## 5 Conclusion

A simple structure has been proposed to realize a current-mode MSO using CDTAs as active elements. The proposed MSO circuit can produce an even number or an odd number of equally spaced in-phase output currents. This circuit offers the following advantages.

(1) Use of only grounded capacitors.
(2) Use of the minimum number of active and passive components.
(3) Independent linear current control of the frequency and the condition of oscillation through the transconductance gain of the CDTA.
(4) Low active and passive sensitivities.
(5) High output impedances.

PSPICE simulation results which agree with the theoretical analysis have been obtained.

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