# DUAL-MODE MULTI-PHASE SINUSOIDAL OSCILLATOR WITH EQUAL AMPLITUDES

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***Abstract:***

*This paper presents a current-mode multiphase sinusoidal oscillator (MSO) employing current-controlled current differencing units (CCDUs). The proposed oscillator consists of only one CCDU and one grounded capacitor for each phase, and it can generate arbitrary 2n-phase current-output signals (n≥2) with equally spaced in phase and equal amplitude. Compared with the related reported current differencing unit (CDU) based configurations, the proposed structure has better performance in terms of the number of required passive resistors, the ability for electronic adjustment of the OF and OC, and the easy fabrication of integrated circuit (IC). Moreover, the described oscillator has low-sensitivity performance and high-output impedance. The theoretical results are confirmed by PSPICE simulation.*

## 1 Introduction

It is well known that multiphase sinusoidal oscillator (MSO) is widely applied in many fields, such as communication systems, instrumentation control, signal processing and measurement systems. In the past few decades, many MSOs based on a variety of voltage-mode and current-mode active building blocks have been reported.

Voltage-mode MSOs have been realized by employing operational transconductance amplifiers (OTAs) [1], current conveyors (CCIIs) [2-3], current feedback operational amplifiers (CFOAs) [4], current differencing buffered amplifiers (CDBAs) [5-6], current amplifiers (CA) [7] etc. Literature [1] provides *n* voltage output signals equal in magnitude and equally spaced in phase, which enjoys low sensitivity performance and wide range electronic tunability. However, it only realizes odd number of phases and uses too many floating passive elements. Although the MOSs in [2-3] encompassing CCIIs and grounded passive elements have been designed for simultaneously realizing odd and even phases, they exhibit some defects in the employment of no voltage-controlled/not electronically operated resistors. A MSO using CFOAs is introduced in Ref. [4], which only requires a small amount of external passive resistors. However, this method requires another current amplifier, and its oscillation condition and oscillation frequency can not be adjusted electronically. The multiphase oscillator described in Ref. [6] can generate 2n different phase sinusoidal voltage outputs by employing *n* lossy integrators and *n* inverters using CDBAs, and its oscillation condition and oscillation frequency can also be adjusted independently by changing the value of passive elements. However, this oscillator

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configuration employs a number of active and passive elements, which is not an ideal result for IC fabrication. In Ref. [7], a CA-based MSO is presented, and its structure is simple. However, the MSO requires two capacitors and two active blocks. From the above analysis, the existed MSOs using OTA, CCII, CFOA, CDBA and CA have the disadvantages of using large number of active and passive elements, most of them are without voltage-controlled frequency tunability.

In recent years, the current-mode MSO has received widespread attention. This operation mode includes the advantages of higher slew rate, larger dynamic range, greater linearity, lower power consumption and wider signaling bandwidth etc. Taking this fact into consideration, the current-mode MSO based on current followers has been presented [8]. This structure is composed of two current followers (CF), one floating resistor and one floating capacitor for each phase and can not be electronically adjusted. By contrast, the current controlled current conveyor (CCCII)-based topology [9] has some better performances in terms of the number of active elements, electronic tunability and output impedance. Unfortunately, this oscillator contains a great number of external passive capacitors. Additionally, its oscillation condition can be realized by tuning capacitance ratio of external capacitors, which is not easy to implement in its practical application.

In 2003, a novel active device, namely, current differencing trasconductance amplifier (CDTA) has been developed [10]. Compared to the former active building blocks, it seems to be a really current-mode active element since its input and output signals are types of current-mode. Consequently, CDTA is probably the most frequently used as an active element in recent design of current-mode analog signal processing [11-20]. From literature survey, it is also found that the MSO circuits exploiting CDTAs were reported in papers [17-20]. Attractive characteristic of MSOs using CDTAs is the possibility of low-input and high-output impedances and independent control of the oscillation frequency and the oscillation condition through external passive elements or bias currents. Regrettably, oscillators in [17] and [18] require an additional current amplifier employing two active devices. Especially, the integrator-based lossy oscillator [17] provides current signals of different amplitudes. The MSO system [18] using CDTA-based all-pass section makes use of two CDTAs in each all-pass section, which is more complicated. Since the topologies in [19-20] suffer from the shortcoming that one external resistor is required for realizing each all-pass section, it is not appropriate for monolithic integration. Moreover, from the point of view of on-chip integration, the complexity of the active device has a direct impact on the area and the power of the chip. Apparently, each CDTA or CCCDTA requires one CDU or CCDU and one OTA. Undoubtedly, this composition greatly increases the chip area and power consumption.

In view of the above disadvantages in previous studies, a current-mode MSO based on current differencing unit has been proposed in [21]. This oscillator enjoys the prominent advantage in the complexity of the active device, the chip area and power. However, drawbacks of this circuit are the following: (a) one passive floating resistor is required for each phase; (b) electronic tunability is not possible; (c) capacitors are not grounded; and (d) different amplitudes of the generated waveforms.

In this paper, we propose a new current-controlled current-mode MSO circuit by using all-pass filters (APF) employing fundamental current-controlled current differencing units. The novel structure offers the following attractive characteristics: (1) only one grounded capacitor is required for realizing each all-pass section without requiring resistors; (2) all passive elements are grounded; (3) electronic tunability can be realized through adjusting the bias current of CCDU; (4) the complexity of the employed active block is lower than those of the already reported blocks; and (5) the amplitudes of the output signals are equal. According to the above analysis, a comprehensive comparison of the presented MSO circuit and previously reported works [17-21] based on the above mentioned features is given in Table 1.

## 2 Current-controlled current differencing unit (CCDU)

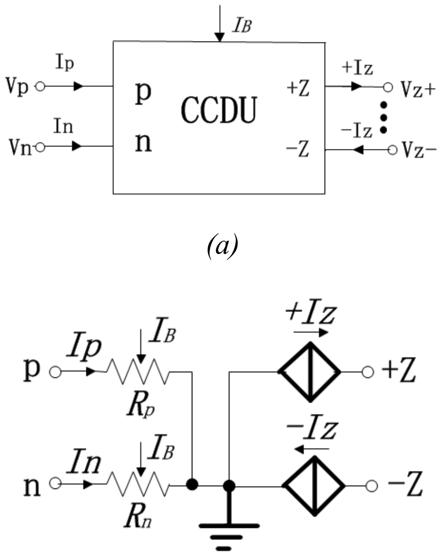
The circuit symbol and equivalent circuit of the CCDU are shown in Fig. 1.

The terminal relations of the CCDU can be described by the following matrix equation [22]:

*Table 1. Comparison with the reported MSOs in [16-20]*

| Topology | Number of active elements | Number of resistors for each phase | Passive components grounded | Electronic tunability of the OF and OC | Equality of Amplitudes of each phase |
|---|---|---|---|---|---|
| [16] | n+2 | 0 | Yes | Yes | No |
| [17] | 2n+2 | 0 | No | Yes | Yes |
| [18] | n | 2 | No | No | Yes |
| [19] | n | 1 | Yes | Yes | Yes |
| [20] | n | 1 | No | No | No |
| This work | n+1 | 0 | Yes | Yes | Yes |

$$\begin{bmatrix} V_p \\ V_n \\ I_z \end{bmatrix} = \begin{bmatrix} R_p & 0 & 0 \\ 0 & R_n & 0 \\ 1 & -1 & 0 \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ 0 \end{bmatrix} \quad (1)$$




(b)

*Figure 1. CCDU: (a) circuit symbol; (b) equivalent circuit.*

According to Equation (1) and the equivalent circuit in Fig. 1(b), the current flowing out of the terminal +*z* (+$I_z$) is a difference between the input currents through the terminals *p* and *n* ($I_p$-$I_n$). The proposed device provides current $I_z$ in both directions, but they are equal in magnitude. Although there are several techniques to realize CCDU, one possible low voltage CMOS implementation is shown in Fig. 2 [22]. The circuit realization mainly comprises a mixed translinear loops (M1-M6). The mixed loops are DC bias by $I_B$ using current mirrors (M7-M10) and (M17-M19). The output +*z* terminal generates the current difference of *p* and *n* terminals, and it is realized using transistors (M11-M16) and (M20-M27). The internal current mirror using resistors (M28-M33) provides a copy of the current flowing out of the +*z* terminal to the -*z* terminals. Assuming that the transistors M1-M6 and M26-M27 operate in saturation region, the input parasitic resistances of *p* and *n* terminals can be characterized with the following equations:

$$R_p = \frac{1}{g_{m3} + g_{m6}} \quad (2)$$

$$R_n = \frac{1}{g_{m2} + g_{m5}} \quad (3)$$

where $g_{mi}$ is the transconductance of transistor $M_i$. If transistors M1-M6 and M26-M27 are identical, the parasitic resistances of *p* and *n* can be approximately obtained by

$$R_p \cong R_n \cong \frac{1}{\sqrt{8\mu C_{ox}(W/L)I_B}} \quad (4)$$

where $\mu$ is the carrier mobility, and $C_{ox}$ is the gate oxide capacitance per unit area, *W* and *L* are the effective channel width and length, respectively. From Equation (4), the parasitic resistances $R_p$ and $R_n$ can be controlled by adjusting the bias current $I_B$ of the CCDU. This property makes it different from conventional CDUs. In addition, another advantage of this device is its direct current-to-current conveying characteristic between input and output terminals. Hence, it is a really current-mode active building block.

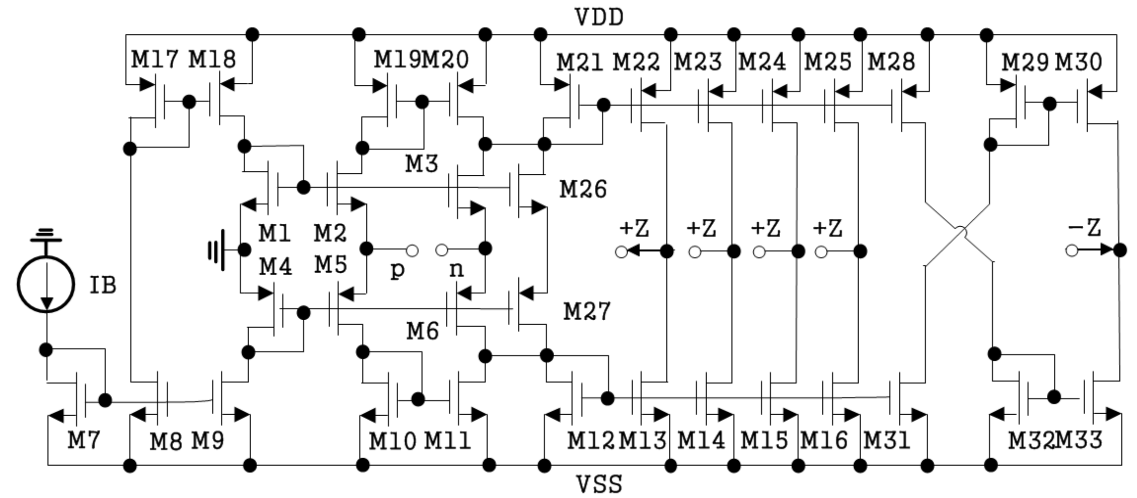



*Figure 2. A possible CMOS implementation for CCDU.*

## 3 Circuit description

In general, a MSO can be realized by choosing various topologies, such as lossy integrators and high-pass filters, etc. In this work, we present a MSO composed of first-order all-pass filters (APF). This structure is based on a single-phase oscillator whose output is copied onto the other APF providing the required phase-shifts of the output signals. The excellence of this topology is direct current-to-current conversions between independent APF blocks. In this topology, each all-pass filter is based on one current-controlled current differencing unit. Additionally, to get different output amplitudes at identical frequency, a current amplifier is needed in this case. The current-mode APF section and current inverting amplifier employing CCDU are analyzed as follows.

### 3.1 CCDU-based current-mode first-order all-pass section

A resistor-less current-mode first-order all-pass filter (APF) topology, with positive and negative outputs, employing only one CCDU and one grounded capacitor is shown in Fig. 3. Routine analysis using the defined equation in (1), we can get the current transfer function of the circuit in Fig. 3. It is indicated by the following expression:

$$\frac{I_{out}}{I_{in}} = \frac{\dfrac{1}{R_{pi}C} - S}{\dfrac{1}{R_{pi}C} + S} \tag{5}$$

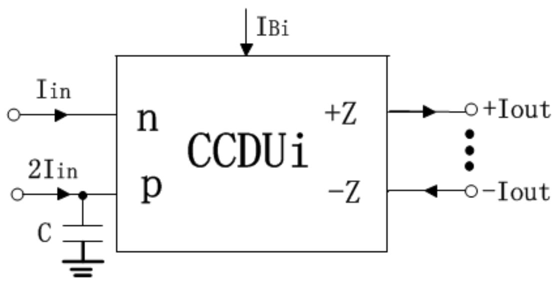



*Figure 3. CCDU-based current-mode first-order all-pass section.*

Here, $R_{pi}$ is the input resistance of the CCDUi at the $p$ terminal. We can see from Fig. 3 that the CCDU-based circuit does not require any external passive resistor. The phase shift *($\varphi$)* of this section can be written as

$$\varphi = -2\tan^{-1}\left(\omega R_{pi} C\right) \tag{6}$$

In the ideal case, the $\varphi$ with the $\omega$ changes from 0° to -180° and from zero (DC) to infinity. It is obvious, from Equation (6), that the shifted phase value can be tuned electronically by adjusting the bias current $I_{Bi}$.

### 3.2 CCDU-based current-controlled current inverting amplifier

A current tunable current inverting amplifier using only one CCDU and one grounded resistor has been illustrated in Fig. 4. In this case, the current gain *(K)* of this circuit is:

$$K = \frac{I_{out}}{I_{in}} = -\frac{R}{R_{na}} \tag{7}$$

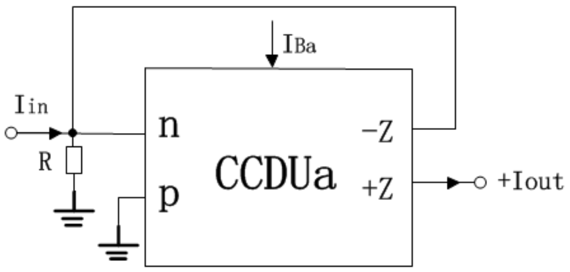



*Figure 4. CCDU-based current-controlled current inverting amplifier.*

where $R_{na}$ is the input resistance of CCDU at an $n$ terminal, note from Eq. (7) that the amplifier gain $K$ can be controlled electronically by tuning the bias current $I_{Ba}$.

### 3.3 Proposed CCDU-based equal amplitude current-mode MSO circuit

The multiphase sinusoidal oscillator (MSO) based on the CCDU is shown in Fig. 5. We can note that the system includes $n$ cascaded CCDU-based all-pass sections of Fig. 3. The output current $I_{on}$ of the nth stage is fed back to the input of the first stage through the current inverting amplifier of Fig. 4. Relative to the already reported MSO, this structure reduces the complexity of oscillator topology, considerably. Moreover, it is suitable for integration point of view [23] for the absence of the floating passive elements. Assuming that $R_{P1}=R_{P2}=\ldots=R_{Pn}=R_P$, $(I_{B1}=I_{B2}=\ldots=I_{Bn}=I_B)$ and $C_1=C_2=\ldots=C_n=C$, the loop gain $L(S)$ of the proposed structure in Fig. 5 can be calculated as:

$$L(S) = K\left(\frac{1/R_pC - S}{1/R_pC + S}\right)^n = -\frac{R}{R_{na}}\left(\frac{1/R_pC - S}{1/R_pC + S}\right)^n \tag{8}$$

According to the Barkhausen criterion, the condition for the proposed MSO circuit of Fig. 5 to produce and sustain sinusoidal oscillation of frequency ($\omega=2\pi f$) is that

$$L(j\omega) = -\frac{R}{R_{na}}\left(\frac{1/R_pC - j\omega}{1/R_pC + j\omega}\right)^n = 1 \tag{9}$$

$$\phi = \frac{\pi}{n} \tag{10}$$

where $n\geq 2$. Eq. (10) shows that there are $n$ outputs $I_{oi}$(i=1, 2, …, n) of each shifted in phase by 180°/n available from the topology. Thus, the oscillation condition (OC) and oscillation frequency (OF) can be found from (9) as:

$$\text{OC:} \quad \frac{R}{R_{na}} = 1 \tag{11}$$

$$\text{OF:} \quad \omega = \left(\frac{1}{CR_p}\right)\tan\left(\frac{\pi}{2n}\right) \tag{12}$$

Further, by substituting Eq. (4) into Eqs. (11) and (12), yields the following the oscillation condition and the oscillation frequency:

$$\text{OC:} \quad R = \frac{1}{\sqrt{8\mu C_{ox}(W/L)I_{Ba}}} \tag{13}$$

$$\text{OF:} \quad \omega = \left(\frac{\sqrt{8\mu C_{ox}(W/L)I_B}}{C}\right)\tan\left(\frac{\pi}{2n}\right) \tag{14}$$

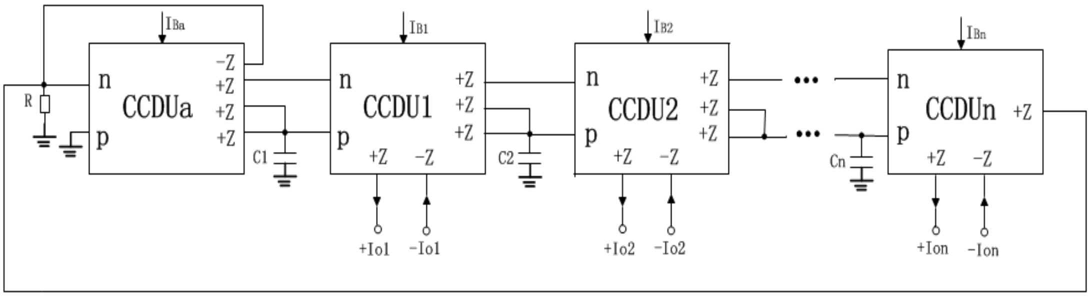



*Figure 5. Proposed current-mode MSO circuit.*

It can be seen from Eqs. (13) and (14) that the ω can be tuned electronically by the bias current $I_B$ without affecting the condition of oscillation, which can also be changed through varying the bias current $I_{Ba}$ without influencing the $\omega$. Therefore, the circuit provides attractive feature of independent current control of the frequency and the condition of oscillations. Additionally, employing an inverted version of the output current of the CCDU, the 2n phase outputs ($-I_{o1}$, $-I_{o2}$ … $-I_{on}$) are also obtained from the same topology.

## 4 Analysis of non-ideal case

For the non-ideal case, the CCDU in Fig. 1 (a) is modeled by the circuit in Fig. 6 [17]. As is seen, there are parasitic resistance $R_z$ and parasitic capacitances $C_z$ at terminal $z$. Now, the non-ideal $z$ terminal current of the CCDU is described in the form

$$I_z = \alpha_p I_p - \alpha_n I_n \tag{15}$$

where the parameters $\alpha_p$ and $\alpha_n$ are current transfer values, which can deviate from their ideal values (one) depending on the internal circuit construction. $R_p$ and $R_n$ are parasitic resistances of $p$ and $n$ terminals. Therefore, taking into account the non-ideal CCDU characteristics and its parasitic elements, the modified current transfer functions of Fig. 3 and Fig. 4 can be rewritten as:

$$\frac{I_{out}}{I_{in}} = \alpha_n \frac{1 - s\left(\dfrac{\alpha_n C R_p}{\alpha_p}\right)}{\dfrac{\alpha_n}{\alpha_p} + s\left(\dfrac{\alpha_n C R_p}{\alpha_p}\right)} \tag{16}$$

$$\frac{I_{out}}{I_{in}} = -\frac{\alpha_n R}{R_{na}(1 + sC_z) + R\left(1 - \alpha_n + \dfrac{R_{na}}{R_z}\right)} \tag{17}$$

Re-analysis of the proposed MSO circuit in Fig. 5 uses Equations (16) and (17) with $C_z \ll C$, $R_{na} \ll R_z$, the modified $\omega$ and oscillation condition can be given, respectively, by the following relations:

$$\text{OC:} \quad R = \frac{1}{(2\alpha_n - 1)\sqrt{8\mu C_{ox}(W/L) I_{Ba}}} \tag{18}$$

$$\text{OF:} \quad \omega = \left(\frac{\alpha_p \sqrt{8\mu C_{ox}(W/L) I_B}}{\alpha_n C}\right) \tan\left(\frac{\pi}{2n}\right) \tag{19}$$

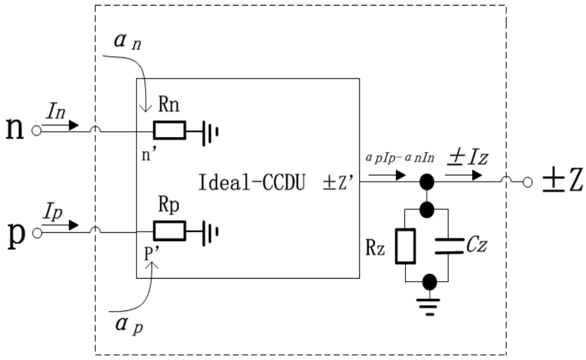



*Figure 6. Model of non-ideal CCDU.*

It should be mentioned that the oscillation condition is mainly affected by the current transfer error $\alpha_n$ of the CCDU. However, the effect of these errors can be easily accommodated by changing the bias current ($I_{Ba}$). Also, $\omega$ is slightly deviated from the ideal case by the factor of $\alpha_p/\alpha_n$, however, we can slightly adjust the bias current $I_B$ to compensate this effect. By following the well-known definition for the sensitivity of a parameter $F$ to an element variation $x_i$, recall the following equation:

$$S_{xi}^{F} = \frac{xi}{F}\frac{\partial F}{\partial xi} \tag{20}$$

The $\omega$-sensitivity analysis with respect to the parameters of the active and passive element used can be derived as:

$$S_{\alpha_p}^{\omega} = -S_{\alpha_n}^{\omega} = -S_{R_p}^{\omega} = 1 \tag{21}$$

$$S_{C}^{\omega} = -1 \tag{22}$$

Thus, the proposed topology enjoys low sensitivity values in active and passive element variations.

## 5 Simulation results

The behavior of the MSO topology in Fig. 5 has been studied through simulations performed by employing PSPICE software. The CCDU in Fig. 2 was simulated using the TSMC 0.25 um CMOS technology [24]. The dimensions of the CMOS transistors are listed in Table 2. Low voltage low power circuit design has been very popular in recent years [25], in our design, the CCDU circuit is supplied with symmetrical voltages of VDD=-VSS = 1.5 V. By PSPICE simulations, the variations of resistance ($R_p$) with the bias current $I_B$ are shown in Fig. 7. It is noted that the resistance $R_p$ can be controlled via changing the bias current $I_B$ of the CCDU within a wide current range.

In order to prove the performances of the proposed circuit, a six-phase (n = 3) oscillator has been designed. Here, external bias currents are $I_{B1} = I_{B2} = I_{B2} = I_{B2} =$ 40 μA, ($R_p$ = 2.62 kΩ), $I_{Ba}$ = 35 μA ($R_{na}$ = 3.6 kΩ) and the grounded capacitors $C_1 = C_2 = C_3 = C$ =52 pF and resistor $R$=3.63 kΩ, respectively. The frequency response of the resistance $R_p$ is given in Fig. 8, while the bias current $I_B$ is 40 μA. The simulation result also shows that the resistance $R_p$ is stable in a large frequency range.

*Table 2. Dimensions of the CMOS transistors*

| COMS TRANSISTORS | W(um)/L(um) |
|---|---|
| M1-M3,M26 | 2/0.5 |
| M4-M6,M27 | 7/0.5 |
| M7-M16,M31-M33 | 5/0.5 |
| M17-M25,M28-M30 | 15/0.5 |

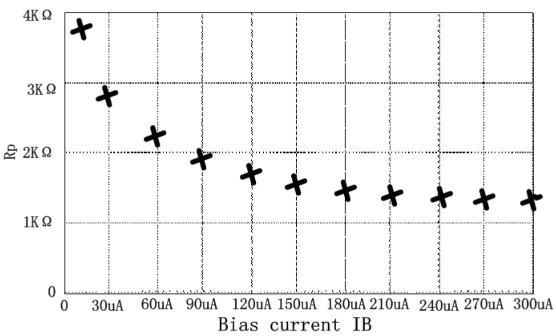



*Figure 7. Variability of the resistance $R_p$ versus bias current $I_B$.*

The simulated output waveforms and frequency spectrums of the proposed MSO are shown in Fig. 9 and Fig. 10. Figure 9 shows the current output signals of each stage in the designed six-phase oscillator. Figure 9 shows clearly that the amplitudes of all output signals are equal. The oscillation frequency is measured as 672.24 kHz, which is close to the theoretically calculated value of 674.77 kHz. From the simulation results it is noted that the phase difference of $I_{o2}$, $I_{o3}$, -$I_{o1}$, -$I_{o2}$ and –$I_{o3}$ compared with $I_{o1}$ are 61°, 119°,180°, 242° and 300°, are very close to the theoretical values.

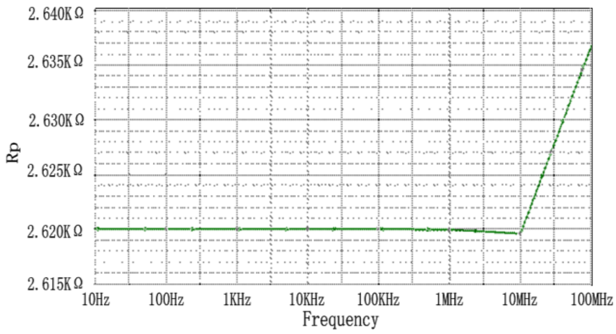


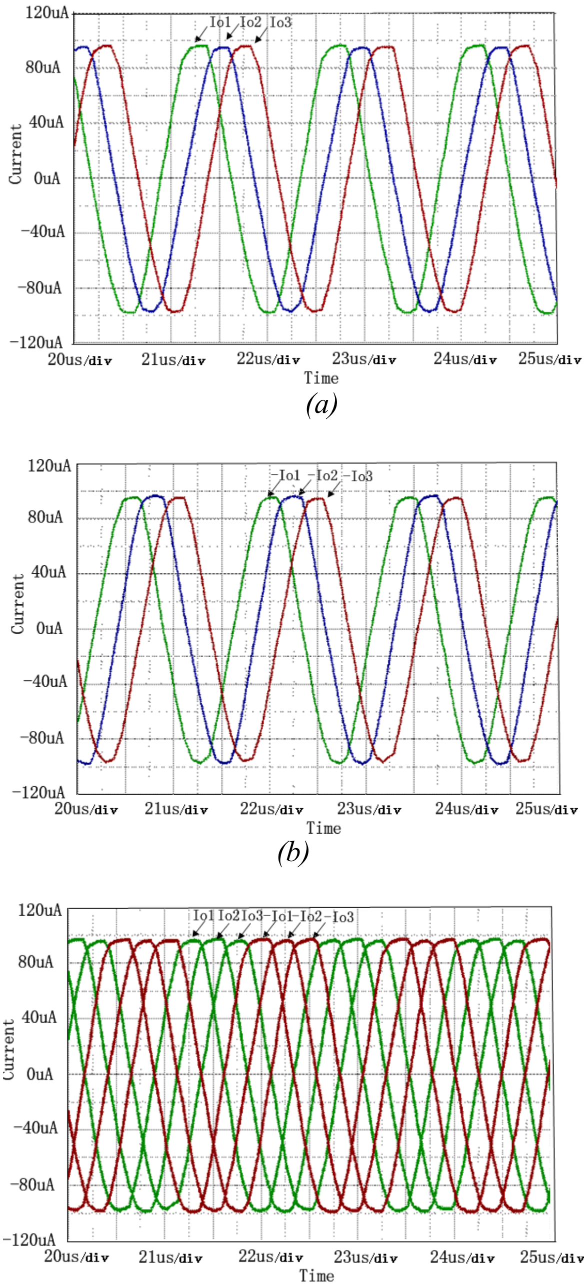

*Figure 8. Variation of $R_p$ versus frequency.*

The variability of the oscillation frequency ($f_o$) as a function of the bias current ($I_B$) is shown in Fig. 11. It is easily noted that the circuit exhibits a large tuning range. Although there is a deviation between the theoretical values and the simulation values in high bias current value region, decreasing this error can be offset by simply modifying the value of external capacitor.

To test the accuracy of output waveforms of the proposed MSO, the total harmonic circuit distortion (THD) analysis is presented in Fig. 12. It is easy to see that the THD of this oscillator is less than 2.2 % when frequency is less than 1 MHz.

## 6 Conclusion

A new current-mode MSO based on first-order all-pass filter employing simple CCDU active building block is presented and analyzed in this paper. The proposed MSO offers lower passive and active component count compared with that of the corresponding already published structure. Moreover, the oscillation condition and oscillation frequency can be electronically adjusted through varying the bias currents of CCDU. Meanwhile, this circuit is suitable for monolithic integrated circuit (IC) due to using simple active element (CCDU) and including only grounded passive elements. The proposed circuit provides high output impedance and low active and passive sensitivities. It is verified from PSPICE simulation that the results agree well with the theoretical analysis. Thus, it is expected that the proposed topology could be employed in current-mode analog signal processing.




(a)

(b)

(c)

*Figure 9. Simulated output waveforms of the proposed current-mode MSO of Fig. 4.:(a) $I_{o1}$, $I_{o2}$, $I_{o3}$; (b) $-I_{o1}$, $-I_{o2}$, $-I_{o3}$; (c) $I_{o1}$, $I_{o2}$, $I_{o3}$, $-I_{o1}$, $-I_{o2}$, $-I_{o3}$ .*

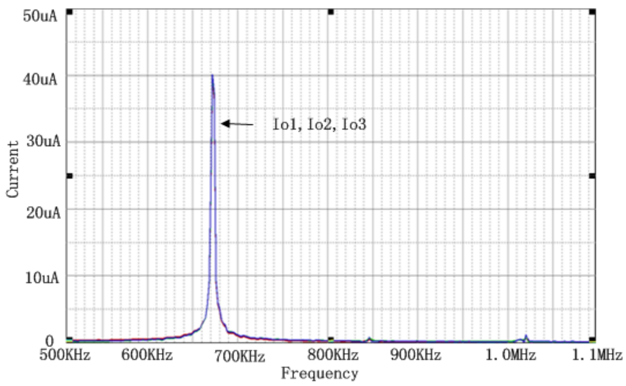



*Figure 10. Frequency spectrums of $I_{o1}$, $I_{o2}$ and $I_{o3}$ .*

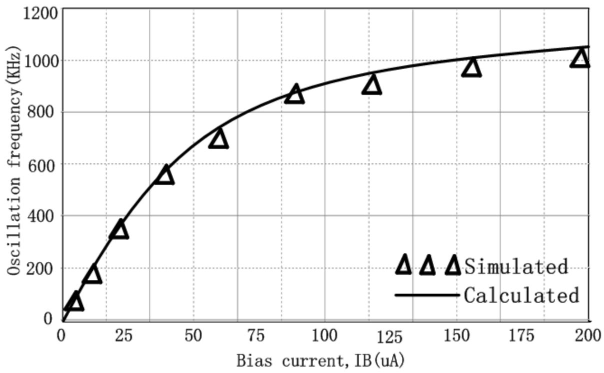



*Figure 11. Variations of the frequency ($f_o$) versus bias currents $I_B$ .*

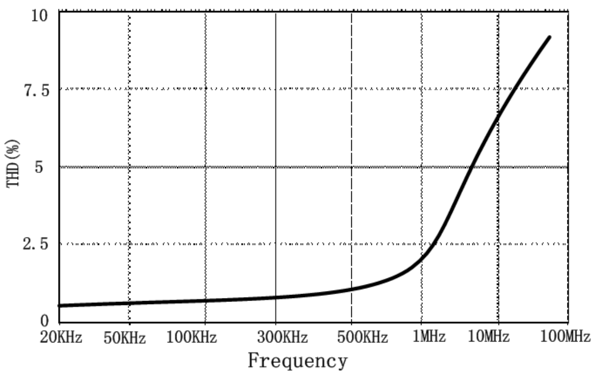



*Figure 12. THD versus oscillation frequency.*

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