High-output-impedance current-mode multiphase sinusoidal oscillator employing current differencing transconductance amplifier-based allpass filters

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Online publication date: 05 July 2010

To cite this Article Jaikla, Winai, Siripruchyanun, Montree, Biolek, Dalibor and Biolkova, Viera(2010) 'High-output-impedance current-mode multiphase sinusoidal oscillator employing current differencing transconductance amplifier-based allpass filters', International Journal of Electronics, 97: 7, 811 — 826

To link to this Article DOI: 10.1080/00207211003733288

URL: http://dx.doi.org/10.1080/00207211003733288
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(Received 13 February 2009; final version received 26 February 2010)

This article presents a multiphase sinusoidal oscillator using current differencing transconductance amplifier (CDTA)-based allpass filters. The oscillation condition and oscillation frequency can be orthogonally controlled. The proposed circuit provides $2^n$ – phase signals ($n \geq 2$) that are equally spaced in phase and of equal amplitude. The circuit requires one CDTA, two resistors and one capacitor for each phase and no additional current amplifier. Owing to high-output impedances, the proposed circuit enables easy cascading in current-mode configurations. The effects of the nonidealities of the CDTA-allpass sections were also studied. The results of PSpice simulations are presented, demonstrating their consistency with theoretical assumptions.

Keywords: multiphase sinusoidal oscillator; CDTA; current-mode

1. Introduction

In electrical engineering applications, the multiphase sinusoidal oscillator (MSO) is treated as an important building block, widely used in many fields such as communications, signal processing, measurement systems and power electronics. Many techniques that implement MSOs have been presented by employing various high-performance active building blocks.

Voltage-mode MSOs based on current conveyors (CCIIs; Hou and Shen 1995; Wu, Liu, Hwang and Wu 1995a; Abuelma’atti and Al-Qahtani 1998a,c), current feedback operational amplifiers (CFOAs; Wu, Liu, Hwang and Wu 1995b), current differencing buffered amplifiers (CDBAs; Klahan, Tangsrirat and Surakampontorn 2004), and op-amps (Gift 1998, 2000) have been developed. The above circuits perform well, but they also suffer from several disadvantages. For example, let us mention the excessive use of passive elements, particularly external resistors, the lack

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ISSN 0020-7217 print/ISSN 1362-3060 online
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DOI: 10.1080/00207211003733288
http://www.informaworld.com
of electronic adjustability and the performance limitations at high frequencies due to finite gain-bandwidth of the op-amps. An operational transconductance amplifier (OTA)-based voltage-mode MSO (Khan, Ahmed and Minhaj 1992) enjoys electronic tunability and simple structure. Unfortunately, these OTA-based applications labour with a limited operating range and low voltage swing.

The current-mode technique has been more popular than the voltage-mode approach due to the specific requirements in low-voltage applications such as portability and battery-powered equipment. Presently, there is a growing interest in synthesising current-mode circuits because of their potential advantages such as larger dynamic range, higher signal bandwidth, greater linearity, simpler circuitry and lower power consumption (Toumazou, Lidgey and Haigh 1990). In particular, current-mode circuits with their high-output impedances are of great interest because of the ease with which they drive the loads and the way they can be cascaded without utilising any buffering devices (Abuelma’atti and Al-Zaher 1999; Cam, Toker, Cicekoglu and Kuntman 2000).

A current-mode MSO based on current followers has also been proposed (Abuelma’atti 1994). This topology requires two current followers, one floating resistor and one floating capacitor for each phase. However, it cannot be electronically controlled. The current controlled current conveyor (CCCII)-based MSOs (Abuelma’atti and Al-Qahtani 1998b; Loescharataramdee, Kiranon, Sangpisit and Yadum 2004) enjoy high-output impedances and electronic tunability. However, the first one requires a large number of external capacitors. In addition, the oscillation condition (OC) can be provided by tuning the capacitance ratio of external capacitors, which is not easy to implement. The latter circuit requires additional current amplifiers, which makes the circuit more complicated and increases its power consumption.

Recently, the current differencing transconductance amplifier (CDTA) (Biolek 2003) has been introduced. It seems to be a versatile component for the realisation of a class of analogue signal processing circuits, not just analogue frequency filters (Biolek 2003; Biolek, Hancioglu and Keskin 2008). It is really a current-mode element whose input and output signals are currents. In addition, the output currents of the CDTA can be electronically adjusted.

CDTA-based current-mode MSOs have been recently presented (Tangsrirat and Tanjaroen 2008; Tangsrirat, Tanjaroen and Pukkalanun 2009). The first MSO employs CDTA-based lossy integrators, whereas the MSO in the second article contains CDTA-based allpass sections. Both circuits exhibit good performance in terms of electronic tunability, high-output impedances and independent control of the oscillation frequency (OF) and the OC. However, both of them require an additional current amplifier, which is implemented by two CDTAs. Moreover, the output waveforms of the MSO, utilising the CDTA-based lossy integrators, are of different amplitudes. An MSO employing CDTA-based allpass sections requires two CDTAs in each allpass section, and the circuitry becomes more extensive. Consequently, it occupies a larger chip area for VLSI design and its practical implementation for off-the-shelf design becomes difficult. In addition, its power consumption is also increased.

The aim of this article is to propose an alternative topology for current-mode MSO based on CDTAs. The features of the proposed circuit are the following:

1. The possibility of generating multi-phase harmonic signals for both an even and an odd number of equally spaced phases.
(2) High-impedance current outputs.
(3) Electronic adjustment of the OC.
(4) Independent control of the OF and the OC.
(5) Use of identical circuit configurations for each section in the MSO topology, which is appropriate for mass fabrication.
(6) Equality of amplitudes of each phase due to utilising identical sections.
(7) Requirement for only one CDTA as the active element for each phase without any additional current amplifiers.

Note that the first six features mentioned above, which are suggested for the proper operation of current-mode multiphase oscillators, are fulfilled simultaneously by the proposed topology, and, in addition, that no hitherto published oscillator from the reference list fulfills the feature (7).

The article is organised as follows. Section 2 contains a review of the CDTA, describes the CDTA-based allpass section, and proposes a topology of the MSO, utilising the above all-pass filters in cascade connection. Section 3 analyses in depth the impact of non-ideal effects on the OF and the OC. Section 4 summarises the results of PSpice simulations, demonstrating the features of the proposed oscillator topology. Several suggestions on how to proceed with further improvement of the circuit performance are included in section 5.

2. Principles of operation

2.1. CDTA

Since the proposed circuit is based on the CDTA, a brief review of the CDTA is given in this section.

In general, the CDTA may contain an arbitrary number of $x$ terminals, providing currents $I_x$ in both directions, which are linearly dependent on the voltage of the $z$ terminal. This terminal is driven by an internal current source, which is controlled by the difference of input currents $I_p$ and $I_n$, flowing into low-impedance terminals $p$ and $n$ (Biolek 2003; Keskin and Biolek 2006).

As an example, the symbol and the equivalent circuit of the CDTA with a pair of $x+$ and $x-$ terminals are illustrated in Figure 1(a) and (b), respectively.

The operation of the ideal CDTA (Figure 1(a,b)) is represented by the following hybrid matrix:

$$
\begin{bmatrix}
V_p \\
V_n \\
I_z \\
I_x
\end{bmatrix} =
\begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
1 & -1 & 0 & 0 \\
0 & 0 & 0 & g_m
\end{bmatrix}
\begin{bmatrix}
I_p \\
I_n \\
V_x \\
V_z
\end{bmatrix}. 
$$

(1)

For a CDTA implemented with bipolar technology, the transconductance $g_m$ can be expressed as follows (Tangsrirat et al. 2009):

$$
g_m = \frac{I_B}{2V_T}.
$$

(2)

Here, $V_T$ and $I_B$ are the thermal voltage and the input bias current, respectively.
2.2. Implementation of CDTA-based allpass sections

As mentioned above, the proposed MSO structure is based on identical first-order allpass sections. A prospective CDTA-based implementation is shown in Figure 2 (Keskin and Biolek 2006). It starts from the well-known allpass topology which utilises low-impedance $p$ and $n$ inputs of current differencing unit (CDU)-based active elements such as CDBA or CDTA (Biolek, Senani, Biolkova and Kolka 2008). A capacitor/resistor pair is connected between the input terminal and the $n/p$ terminal of the CDU. Then the difference current flowing out of the $z$ terminal is of allpass character. For the CDBA allpass section, the corresponding voltage drop at the $z$-terminal resistance is transformed into the output voltage via a unity-gain buffer (Toker, Ogozuz, Cicekoglu and Acar 2000). For the CDTA-based filter in Figure 2, this voltage is transformed back into current through the internal transconductance $g_m$. The current transfer function can be written as follows:

$$\frac{I_o(s)}{I_{in}(s)} = K \frac{a - s}{a + s}, \quad K = g_m R_K, \quad a = \frac{1}{RC}.$$  \hspace{1cm} (3)

Here $K$ and $a$ denote the filter current gain and natural frequency, respectively.

Note that due to both polarities of the $x$-currents, the circuit in Figure 2 can represent the non-inverting and inverting allpass sections simultaneously. The (non-)inverting property can be also modified by interconnecting the $C$ and $R$ elements (Keskin and Biolek 2006).

It is particularly important that the section gain can be adjusted by the product $g_m R_K$ independently of the natural frequency. This feature will be further used for an easy setting of the OC independently of the frequency of oscillation.

Figure 1. CDTA (a) schematic symbol, (b) equivalent circuit.

Figure 2. CDTA-based allpass filter.
Note again that the CDTA can have more than the two current outputs indicated in Figure 2, which is also important for the easy implementation of the multiphase oscillator.

### 2.3. Implementation of MSO

Recently, three different block diagrams of multiphase oscillators with \( n \) identical first-order cells have been published: (1) all cells are inverting, \( n > 2 \), odd (Abuelma’atti and Al-Qahtani 1998a; Gift 1998, 2000), (2) first \((n-1)\) cells are inverting and the \( n \)-th cell is inverting/non-inverting if \( n \) is odd/even (Souliotis and Psychalinos 2009), (3) first \((n-1)\) cells are noninverting and the last cell is inverting, arbitrary \( n > 1 \) (Abuelma’atti and Al-Qahtani 1998a; Gift 2000). The latter version was selected for implementing the CDTA-based oscillator whose block diagram is shown in Figure 3. With the utilisation of the features of CDTA-based first-order allpass sections, this structure provides a universal solution for an arbitrary number of phases greater than 1.

Since each allpass section provides output currents in both directions with a mutual phase shift of \( \pi \) radians, one should generally distinguish between the number of sections \( (n) \) and the number of phases generated \( (N) \).

The allpass sections are designed with identical natural frequencies; thus, each of them provides an identical phase delay of the signal passing through the cascade. The phase delay, which is in the interval \((0, \pi)\) rads, is denoted as \( \Delta \phi \). The total phase shift \( \phi_S \) within the feedback loop is

\[
\phi_S = n \Delta \phi + \pi. \tag{4}
\]

The total loop phase shift of \( 2\pi \) radians for proper oscillation yields

\[
\Delta \phi = \frac{\pi}{n}. \tag{5}
\]

The corresponding constellation diagrams for \( n = 2, 3, 4 \) and 5 are shown in Table 1. The minimum number \( n \) of allpass cells is 2. The number \( N \) of generated signals with equally spaced initial phases is either \( n \) or \( 2n \) depending on which of the available outputs \( I_o, I'_o \) are used. For example a three-phase oscillator can be easily designed from three allpass cells \( (n = 3) \) and selecting outputs \( I_{o1}, I_{o3}, \text{ and } I'_{o2} \). But utilising all six outputs will yield a six-phase oscillator.

To derive formulae for the OF and OC, the system loop gain can be written as follows (Gift 2000):

\[
L(s) = -K_1 K_2 \cdots K_n \left( \frac{a-s}{a+s} \right)^n. \tag{6}
\]

![Figure 3](image.png)

Figure 3. Block diagram of multiphase oscillator with even or odd number of phases.
At the OF $\omega_{osc} = 2\pi f_{osc}$, the Barkhausen criterion must be fulfilled such that

$$L(j\omega_{osc}) = -\left(\frac{K_{1}a - j\omega_{osc}}{a + j\omega_{osc}}\right)^{n} = 1,$$

(7)

where $K_{1} = K_{2} = \ldots K_{n} = K$, then $K_{1}K_{2} \ldots K_{n} = K^{n}$. From Equation (6), the magnitude and the phase of the system loop gain are as follows:

$$|L(j\omega_{osc})| = 1,$$

(8)

and

$$\angle L(j\omega_{osc}) = 2k\pi, \quad \text{abs}(k) = 0, 1, 2, \ldots .$$

(9)

Combining Equations (7)–(9), the OC and the OF are given by the formulae

$$\text{OC}: \quad K = 1,$$

(10)

and

$$\text{OF}: \quad \omega_{osc} = a \tan\left(\frac{\pi}{2n}\right).$$

(11)

It can be seen from Equations (10) and (11) that the OC can be controlled independently of the OF by the gain $K$, while the OF can be changed by the natural frequency $a$.

The resulting current-mode CDTA-based MSO is shown in Figure 4.

According to Equations (10) and (11), the OC and the OF are as follows:

$$\text{OC}: \quad g_{m}R_{Ki} = 1, \quad i = 1, 2, \ldots n$$

(12)

and

$$\text{OF}: \quad \omega_{osc} = \frac{1}{RC} \tan\left(\frac{\pi}{2n}\right).$$

(13)
Substituting \( g_m = \frac{I_B}{2V_T} \) into Equation (12), the OC becomes

\[
\text{OC: } \frac{I_B}{2V_T} R_{K_i} = 1, \quad i = 1, 2, \ldots n.
\] (14)

3. Analysis of non-ideal case

For the non-ideal case, the allpass section in Figure 2 is modelled by the circuit in Figure 5 (Jaikla, Siripruchyanun, Bajer and Biolek 2008). The \( z \)-terminal current of the CDTA is now described in the form

\[
I_z = \alpha_p I_p - \alpha_n I_n
\] (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 the parasitic resistances of \( p \) and \( n \) terminals.

For simplicity, let us assume that all the allpass sections in Figure 4 are identical. As can be seen from this figure, the output current of section no. \( k \) drives the input of section no. \( k + 1 \). Figure 5 respects this fact such that the output current \( I_k \) of section no. \( k \) is flowing into the ‘\( C-R_n-R-R_p \)’ impedance of the subsequent section. In conjunction with the parasitic impedance of the \( x \) terminal (\( C_x, R_x \)), this impedance acts as a frequency-dependent current divider. The frequency response is also affected by the parasitic impedance of the \( z \) terminal (\( C_z, R_z \)).
Analysis of the model in Figure 5 leads to the following transfer function of the nonideal allpass section:

$$H = \frac{I_k}{I_{k-1}} = \frac{I_z}{I_{k-1}} \frac{I_x I_k}{I_z} = H_1 H_2 H_3,$$

where

$$H_1 = \frac{I_z}{I_{k-1}} = \alpha_p \frac{1 - sRC\left(\frac{\beta_p}{\beta_p} \left(1 + \frac{R_p}{R_c}ight) - \frac{R_c}{R_p}\right)}{1 + sRC\left(1 + \frac{R_p}{R_c} + \frac{R_n}{R_c}\right)}$$

and

$$H_2 = \frac{I_x}{I_z} = \frac{g_m R_k}{1 + \frac{R_c}{R_c} + sR_k C_z},$$

$$H_3 = \frac{I_k}{I_x} = \frac{1 + sRC\left(1 + \frac{R_p}{R_c} + \frac{R_n}{R_c}\right)}{1 + \frac{R_p}{R_c} + sR_k C_z} + s^2 R_n (R + R_p) CC_x.$$  

Note that the model of allpass section in Figure 5 represents a fourth-order linear system on the assumption that the parameters $\alpha_p$, $\alpha_n$ and $g_m$ are frequency independent. Otherwise, this order is further increased. The basic phase shift between the output and the input currents, described by transfer function $H_1$, is affected by parasitic parameters $\alpha_p$, $\alpha_n$ and $R_p$, $R_n$. The additional current-to-current conversions, described by transfer functions $H_2$ and $H_3$, include additional phase errors, which altogether modify the OF. As the results from Equations (17) and (19) show, a zero-pole cancellation appears in transfer functions $H_1$ and $H_3$, simplifying the resulting transfer function to the following form:

$$H = \frac{1 - sRC\left(\frac{\beta_p}{\beta_p} \left(1 + \frac{R_p}{R_c}\right) - \frac{R_c}{R_p}\right)}{1 + \frac{R_p}{R_c} + sRC\left(1 + \frac{R_p}{R_c} + \frac{R_n}{R_c}\right)\left(1 + \frac{R_p}{R_c}\right) + s^2 R_n (R + R_p) CC_x} \times \frac{\alpha_p g_m R_k}{1 + \frac{R_c}{R_c} + sR_k C_z}.$$  

Note that after omitting all the terms modelling the parasitic impedances of $z$ and $x$ terminals, Equation (20) is simplified to transfer function $H_1$, multiplied by the current gain $g_m R_k$.

The analytical analysis of the modification of OF and OC for this nonideal case is rather difficult. That is why the following procedure was used:

The circuit in Figure 5 was simulated in PSpice on the basis of behavioural modelling in order to identify how the parameters in Equation (20) affect the frequency response. The numerical values of CDTA parasitics were adopted from section 4, including one-pole modelling of frequency dependancies of parameters $\alpha_p$, $\alpha_n$ and $g_m$, concretely

$$\frac{\alpha_p}{1 + \frac{\omega_p}{\omega_p}}, \frac{\alpha_n}{1 + \frac{\omega_n}{\omega_n}}, \frac{g_m}{1 + \frac{\omega_m}{\omega_m}}.$$  

(21)
The allpass section was designed with the aim to have a three- or six-phase sinusoidal oscillator ($n = 3$, $N = 3$ or 6) with an OF of 400 kHz. This means that each allpass section should perform a phase shift of $60^\circ$ of the passing signal, altogether $180^\circ$ for three sections. The remaining phase shift ($180^\circ$) is accomplished via the inverting feature of the last section. The parameters of the passive components were set as follows: $R = 2.29 \text{k}\Omega$, $C = 100 \text{ pF}$ and $R_k = 460 \text{ }\Omega$. According to Equation (13), the corresponding theoretical value of OF is approximately 401 kHz. With $g_m = 2.346 \text{ mA/V}$ (see section 4), the loop gain in Equation (12) is greater than one, which theoretically guarantees the self-starting of oscillations:

$$g_m R_k = 1.079 > 1$$

(22)

The simulated frequency responses are shown in Figure 6 for CDTA parasitic capacitances $C_z = 3 \text{ pF}$, $C_x = 2.28 \text{ pF}$ (see section 4).

The magnitude frequency responses show the following: the current transfer from the input to the $z$ terminal $H_1 = I_z/I_k$ is smaller than one, largely due to $\alpha_p < 1$. This gain drop is compensated by choosing a $g_m R_k$ product higher than 1 according to Equation (22). That is why the DC gain $H_1 H_2 = I_z/I_k$ is greater than 1. The total gain $H = H_1 H_2 H_3 = I_z/I_k$ decreases due to the attenuation by current divider ‘$R_x, C_x, R, R_p, R_n, C$’ but still remains greater than 1 within the frequency interval of potential OF, enabling the starting-up oscillations. The approximate starting-up

![Figure 6](image-url)
oscillation condition (S-U OC) for such a non-ideal case, i.e. $H(0) = 1$, can be
derived from Equation (20):

$$\text{S-UOC : } g_mR_k \geq \frac{1}{\alpha_p} \left( 1 + \frac{R_k}{R_z} \right) \left( 1 + \frac{R}{R_x} + \frac{R_p}{R_x} \right)$$  \hspace{2cm} (23)$$

One can check that Equation (23) is fulfilled for our case.

Measuring the frequency at which the phase responses $H_1, H_1H_2$, and $H_1H_2H_3$
cross the level of $-60^\circ$, the following values are obtained:

$$388.2 \text{ kHz}, \ 381.5 \text{ kHz}, \ 378.0 \text{ kHz}$$

This sequence demonstrates how the additional phase shifts of individual sub-
blocks decrease the OF from its theoretical value.

The above analyses also show the negligible influence of the cutoff frequencies of
$\alpha_p$ and $\alpha_n$ coefficients on the OF because of their location high above the operating
frequency of the oscillator (see section 4). Because of this they will not be considered
hereafter.

Owing to the complexity of an analytical analysis of the OF under real influences
modelled by Equations (17)–(21), the following procedure will be used: In the first
step, the real effects within the current differencing unit will be studied with the aim
of finding a closed-form formula for oscillating frequency, without considering the
real properties of the remaining blocks. Subsequently, some correcting terms will be
added to this result, taking into account the non-ideal behaviour of the remaining
circuitry.

For the transparency of the subsequent analysis, the transfer function $H_1$ in
Equation (17), describing the operation of the current differencing unit, can be
rewritten in the form

$$H_1 = \frac{I_z}{I_{k-1}} = \frac{\alpha_p \left( 1 - \frac{s\tau_1}{1 + s\tau_2} \right)}{1 + \frac{s\tau_2}{1 + s\tau_2}}, \quad \tau_1 = RC \left[ \frac{\alpha_n}{\alpha_p} \left( 1 + \frac{R_p}{R} \right) - \frac{R_n}{R} \right], \quad \tau_2 = RC \left( 1 + \frac{R_p}{R} + \frac{R_n}{R} \right),$$  \hspace{2cm} (24)$$

where the time constants $\tau_1$ and $\tau_2$ are generally different when considering the above
non-idealities.

Assuming the real value of $\alpha_p > 0$, the corresponding phase frequency response is as follows:

$$\phi_1(\omega) = \angle H_1(j\omega) = -\tan^{-1} \frac{\omega (\tau_1 + \tau_2)}{1 - \omega^2 \tau_1 \tau_2}, \quad \omega^2 \tau_1 \tau_2 < 1.$$  \hspace{2cm} (25)$$

The phase OC $\Phi_1(\omega_{osc}') = -\pi / n$ leads to the formula for OF $\omega_{osc}'$ for the non-ideal
case:

$$\omega_{osc}' = \frac{\tau_1 + \tau_2}{2\tau_1 \tau_2 \tan \frac{\pi}{n}} \left[ \sqrt{1 + 4 \left( \frac{\tau_1 \tau_2}{(\tau_1 + \tau_2)^2} \tan \frac{\pi}{n} \right)} - 1 \right].$$  \hspace{2cm} (26)$$

For the ideal case, $\tau_1 = \tau_2 = 1/RC$, Equation (26) changes to Equation (13).
Tedious rearrangements of Equation (26) lead to the following result:

\[
\frac{\omega_{\text{osc}}'}{\omega_{\text{osc}}} = l \left[ 1 + \sqrt{1 - m^2 \sin^2 \frac{\pi}{n} - 1} \right],
\]

(27)

where

\[
l = \frac{(1 + \frac{R_p}{R})(1 + \frac{z_n}{z_p})}{2\frac{z_n}{z_p}(1 + \frac{R_p}{R}) - \frac{R_p}{R}(1 + \frac{R_p}{R} + \frac{R_n}{R})}, \quad m = \frac{\tau_1 - \tau_2}{\tau_1 + \tau_2} = \frac{\frac{z_n}{z_p} - 1 - \frac{2R_p}{R_p + R_n}}{\frac{z_n}{z_p} + 1}. \]

(28)

For the real parameters of CDTA in section 4 and \( R = 2.29 \) k\( \Omega \), the quantity \( m \) becomes \( m = -0.001085 \), and the second term inside the square brackets of Equation (27) can be neglected with an error smaller than \( 9 \times 10^{-6} \). Then the ratio of both frequencies is \( l = 0.968 \). The theoretical OF 401 kHz should be decreased to ca. 388 kHz. To compensate for this modification, Equation (28) shows that we should design the \( R \) value to be as large as possible in comparison with the \( R_p \) and \( R_n \) values, and also to design the CDTA input stage such that the symmetry \( x_p = x_n \) is preserved.

As shown in section 4 and also as indicated above, the OF is also affected by the parasitic impedances of the \( z \) and \( x \) terminals of the CDTA as well as by the frequency dependence of the transconductance \( g_m \). These effects were not taken into account in Equations (27) and (28). They include an additional phase shift to the loop gain of the oscillator and cause a modification of the OF, i.e. the frequency at which the resulting phase of one allpass section crosses the \(-\pi/n\) level. These additional phase shifts, much smaller than the phase shift generated by transfer function (17), can be evaluated from transfer function (20). Then they can be converted into additional frequency shifts via the slope of phase response \( \frac{\partial \Phi_1}{\partial \omega} \) or, alternatively, from the slope of the inverse function \( \omega = \omega (\Phi_1) \) at \( \Phi_1 = -\pi/n \).

It can be derived from Equations (25) and (26) that the above slope is as follows:

\[
\omega_{\Phi_1}' = \frac{\partial \omega}{\partial \Phi_1} = \frac{\omega}{\sin \Phi_1 \sqrt{1 - m^2 \sin^2 \Phi_1}} \approx \frac{\omega}{\sin \Phi_1} \text{ for } m^2 \ll 1.
\]

(29)

Then the additional modification of the OF \( \Delta \omega_{\text{osc}} \) due to a phase shift \( \Delta \phi \) caused by a non-ideal effect is approximately

\[
\Delta \omega_{\text{osc}}' \approx \frac{\omega_{\text{osc}}'}{\sin \frac{\pi}{n}} \Delta \phi \quad \text{or} \quad \frac{\Delta \omega_{\text{osc}}'}{\omega_{\text{osc}}'} \approx \frac{\Delta \phi}{\sin \frac{\pi}{n}}.
\]

(30)

Equation (30) indicates that the sensitivity of the OF to additional parasitic phase shifts in the feedback loop increases with increasing the number of phases \( 2n \).

The phase shifts, generated by parasitic elements \( R_z, C_z, R_x, C_x \), and by the frequency dependence of \( g_m \) (see Equation 21), can be computed from Equations (18) and (19), and evaluated for the values of parameters from section 4 and for \( C_z = 3 \) pF, \( C_x = 2.28 \) pF:

\[
\Delta \phi_{R \pm C_z} = -\tan^{-1} \frac{\omega_{\text{osc}}' C_z}{\frac{1}{R_z} + \frac{1}{R_x}} \approx -3.341 \times 10^{-3} \text{ rads},
\]

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\[
\Delta \phi_{RxCx} = \tan^{-1} \left[ \frac{\omega'_{osc} RC}{1 + \frac{R_p + R_n}{R}} \right]
\]
\[
- \tan^{-1} \frac{\omega'_{osc} RC}{1 - \omega^2 \frac{R_p + R_z}{R} + \left( 1 + \frac{R_z}{R} \right) \left( \frac{C_x + R_n}{R} \right)} \approx -9.767 \times 10^{-3} \text{ rads}, \quad \frac{\omega^2 \frac{R_p + R_n}{R} Cx_x}{\omega_{osc} Cx_x} < 1 \quad (31)
\]
\[
\Delta \phi_{gm} = -\tan^{-1} \frac{\omega'_{osc}}{\omega_{gm}} \approx -1.213 \times 10^{-2} \text{ rads}.
\]

Applying Equation (30), the corresponding modifications of the OF are as follows:

\[
\Delta f_{osc,RzCz} \approx -1.498 \text{ kHz}, \quad \Delta f_{osc,RxCx} \approx -4.379 \text{ kHz}, \quad \Delta f_{osc,gm} \approx -5.439 \text{ kHz},
\]

thus the real OF will be

\[
f_{osc} + \Delta f_{osc,RzCz} + \Delta f_{osc,RxCx} + \Delta f_{osc,gm} \approx 377 \text{ kHz}.
\]

These conclusions are in conformity with the above results of PSpice analysis.

4. Simulation results

To confirm the performance of the proposed MSO, a PSpice simulation was carried out. Figure 7 shows a CDTA schematic used for the simulation. For simplicity, an auxiliary circuit for offset compensation is omitted here. The PNP and NPN transistors employed in the proposed circuit were modelled by parameters of the PR200N and NR200N bipolar transistors of ALA400 transistor array from AT&T (Frey 1993). CDTA was biased with ±2.5 V supply voltages. For \( I_A = 300 \mu A \) and \( I_B = 120 \mu A \), the following basic small-signal parameters can be obtained:

- Input resistances of \( p \) and \( n \) terminals \( R_p = R_n = 51 \Omega \).
- Resistance and capacitance of \( z \) terminal \( R_z = 59 \text{ k}\Omega, \quad C_z = 3 \text{ pF} \).
- Resistance and capacitance of \( x_+, x_- \) terminals \( R_{x+} = 580 \text{ k}\Omega, \quad R_{x-} = 646 \text{ k}\Omega, \quad C_{x+} = C_{x-} = 2.28 \text{ pF} \).

---

Figure 7. Internal construction of the CDTA.
• Current gains from \( p \) to \( z \) terminal and from \( n \) to \( z \) terminal \( \alpha_p = 0.96 \) and \( \alpha_n = 0.98 \), with \(-3\)dB cutoff frequencies of 102 MHz and 121 MHz, respectively.

• Transconductance of the OTA amplifier \( g_m = 2.346 \) mA/V with a \(-3\)dB cutoff frequency of 32 MHz at \( x_+ \) terminal and 39 MHz at \( x_- \) terminal.

As also shown in section 3, the oscillator providing three- or six-phase \((n = 3, N = 3 \text{ or } 6)\) 400 kHz sinusoidal signals was designed on the basis of Figure 4. The values of the elements and other circuit parameters were set as follows: \( I_B = 120\mu A, \) \( R_k = 0.46 \) k\( \Omega, \) \( R = 2.29 \) k\( \Omega \) and \( C = 100 \) pF. The simulated output waveforms are shown in Figure 8(a) and (b), respectively. The OF of 375 kHz is less than the theoretical value, but near the values which were predicted in section 3. Figure 9 shows the simulated output spectrum, where the total harmonic distortion (THD) is about

![Figure 8. Simulated output waveforms, \( n = 3, N = 3 \) (a), 6 (b).](image)

![Figure 9. Example of simulated spectrum of generated waveform \( I_{o3} \).](image)
1.032%. Figure 10 provides the plots of the simulated and theoretical oscillation frequencies versus the resistance $R$, with $C$ values of 0.1 nF, 1 nF and 10 nF. It proves that the simulation results are in good agreement with the theoretical formula (11).

5. Conclusion

In this study, a current-mode MSO using CDTAs as active elements is presented. The circuit is based on the all-pass filter employing only a single CDTA in each subsection. It enjoys several advantages, such as high-output-impedances, independent control of OF and OC and only one CDTA for each phase; no additional current amplifier is requested. The control of OF can be achieved electronically by substituting resistors $R$ with floating resistance simulators, whose resistances can be electronically adjusted.

The proposed oscillator structure can be further optimised for a concrete case via variously interchanging the $(p, n)$ inputs and $(x, x^-)$ outputs of individual CDTAs. Moreover, the allpass cell can also be designed more economically, applying local negative feedback, which results in omitting the resistor $R_K$ (Keskin and Biolek 2006). However, this method cannot be applied to all the all-pass cells simultaneously due to the subsequent loss of the possibility of controlling the OC, or that condition must be then accomplished by another method. Anyway, such modifications work against other features of the suggested oscillator structures, i.e. the use of unified all-pass sections without the necessity for an additional amplifier in the feedback loop.

The above features demonstrate the suitability of the prospective implementation of the proposed MSO as a monolithic chip for use in modern control and communication systems.

Acknowledgements

This study was supported by the Czech Grant Agency under grant no. 102/09/1628, and by the research programmes of BUT nos. MSM0021630503 and MSM0021630513, and UD Brno no. MO FVT0000403, Czech Republic.
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