

Deriving (MO)(I)CCCII Based Second-order Sinusoidal Oscillators with Non-interactive Tuning Laws using State Variable Method

Abhirup LAHIRI

36-B, J and K Pocket, Dilshad Garden, Delhi, India

lahiriabhirup@yahoo.com

Abstract. *The paper discusses systematic realization of second-order sinusoidal oscillators using multiple-output second-generation current controlled conveyor (MO-CCCII) and/or its inverting equivalent, namely the multiple-output inverting second-generation current controlled conveyor (MO-ICCCII) by state variable method. State variable method is a powerful technique and has been used extensively in the past to realize active RC oscillators using a variety of active building blocks (ABB). In this work, a non-interactive relationship between the condition of oscillation (CO) and the frequency of oscillation (FO) has been chosen priori and then state variable method is applied to derive the oscillators with grounded capacitors. All the resulting oscillator circuits, eight of them, are “resistor-less”, employ grounded capacitors and do not use more than three (MO)(I)CCCII. PSPICE simulation results of a possible CMOS implementation of the oscillators using 0.35 μ m TSMC CMOS technology parameters have validated their workability.*

Keywords

Sinusoidal oscillators, multiple-output second-generation current controlled conveyor (MO-CCCII), multiple-output second-generation inverting current controlled conveyor (MO-ICCCII), state variable method.

1. Introduction

There has been an increasing interest in the design of “resistor-less” and electronically tunable sinusoidal oscillators in the last two decades. Such circuits do not require external linear resistors, provide electronic tunability to the circuit parameters, namely the condition of oscillation (CO) and the frequency of oscillation (FO) and are also beneficial for compensating the process induced variations. Several realizations of “resistor-less” sinusoidal oscillators have been reported in the literature. This includes realizations using operational transconductance amplifiers (OTAs) [1]-[3],

second-generation current-controlled conveyor (CCCII) [4]-[5] and hybrid OTA-CCCII elements like current controlled current differencing transconductance amplifier (CCCDTA) [6]-[8]. It is also desirable to construct the circuit using commercially available active elements rather than hypothetical ones [9]. Single-resistance-controlled (SRC) type oscillators are preferred choice for variable frequency oscillators. The realizations in [1], [4]-[7] are based on the SRC type oscillators where each individual resistor is simulated by either a transconductance g_m element (in OTA-based oscillators) or by parasitic resistance R_x (in CCCII-based oscillators). It is well known that any second-order active RC oscillator circuit providing independent control to the CO and the FO would require at least three resistors and two capacitors (canonicity). Similarly, it can be argued that any second-order “resistor-less” active C oscillator providing non-interactive CO and FO controls would require two capacitors and three resistor simulating elements [9]. Several systematic methods have been devised over the years to derive such SRC type oscillators, e.g. the state variable method used by Senani et al in [9]-[11] and nodal admittance matrix (NAM) used by Soliman in [12]. Perhaps, the most researched SRC-type oscillators [13]-[15] (also in [9]-[12]) are the ones which are defined by the following tuning laws

$$CO: R_1 = R_2, \quad (1)$$

$$FO: f_o = \frac{1}{2\pi} \sqrt{\frac{1}{C_1 C_2 R_1 R_3}}. \quad (2)$$

In this paper, we deal with SRC-type oscillators which are governed by the following tuning laws

$$CO: C_1 R_1 = C_2 R_2, \quad (3)$$

$$FO: f_o = \frac{1}{2\pi} \sqrt{\frac{1}{C_1 C_2 R_1 R_3}}. \quad (4)$$

In the recent communication in [9], a variety of sinusoidal oscillators have been derived using SVA which can be summarized as follows:

- Circuits in Fig. 3(a), 3(b) are governed by tuning laws as in (1) and (2) provided here which have also been researched previously by Senani et al in [10] and [11].

- Circuits in Fig. 3(c), 3(d) are governed by non-independent CO and FO tuning (all the terms in the FO and also present in the CO) and which can be made independent only under the capacitor matching condition.
- Circuit in Fig. 3(e) is a minimum component oscillator (2gm-2C oscillator), but the tuning laws are interactive and independent tuning of the CO and the FO is not possible via transconductances.

To conclude, the oscillators in [9] either report circuits based on tuning laws of (1) and (2) or circuits involving non-independent CO and FO tuning via transconductance (without any component matching). Several oscillator realizations governed by (3) and (4) are also available in the literature, e.g. using unity gain cells in [15], recently proposed oscillators using current feedback amplifiers (CFOAs) in [18] and [19] and using CCCDTA in [7]. In this paper, we use state variable method of [10] to systematically derive “resistor-less” second-order oscillators using multiple-output second-generation current controlled current conveyor (MO-CCCI) and/or its inverting equivalent, namely the multiple-output inverting second-generation current controlled conveyor (MO-ICCCI), which are governed by (3) and (4). The parasitic x terminal resistances of CCCII or ICCCI simulate resistors (tunable via the bias current) and thereby avoiding any use of external resistors. The resulting circuits employ three (MO)(I)CCCI and two grounded capacitors. The realizations are compact and suitable to be implemented in standard bipolar or CMOS technology using the many available realizations of CCCII [20]-[21]. PSPICE simulation results using CMOS implementation of the circuit with $0.35\mu\text{m}$ TSMC CMOS technology parameters, has been included to verify its workability.

2. Oscillator Synthesis via State Variable Method

The method of synthesis of second-order sinusoidal oscillators has been dealt with extensively in [9]-[11] and [14] and here we restate the important steps. In general, a second-order oscillator can be characterized by means of the following matrix equation

$$\begin{bmatrix} \frac{dV_1}{dt} \\ \frac{dV_2}{dt} \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = A \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \quad (5)$$

where V_1 and V_2 are the state variables and are the voltages across the two capacitors C_1 and C_2 , respectively. From (5), the characteristic equation (CE) of the oscillator is given as

$$s^2 - s(a_{11} + a_{22}) + (a_{11}a_{22} - a_{12}a_{21}) = 0. \quad (6)$$

It is evident from (6), that the CO and the FO are given as

$$CO: \quad a_{11} = -a_{22}, \quad (7)$$

$$FO: \quad f_o = \frac{1}{2\pi} \sqrt{a_{11}a_{22} - a_{12}a_{21}}. \quad (8)$$

Now comparing (7) and (8) with the desired tuning laws as in (3) and (4) (with the resistors R_i being simulated by parasitic resistance R_{x_i} , where $i = 1, 2, 3$), we can derive different matrices A_k by appropriately choosing the parameters a_{ij} (where $i = 1, 2$). Working on different combinations, we derive the following matrices that confirm with the tuning laws in (3) and (4):

$$A_1 = \begin{bmatrix} \frac{1}{C_1 R_{x_1}} & -\frac{1}{C_1} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) \\ \frac{1}{C_2 R_{x_1}} & -\frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (9)$$

$$A_2 = \begin{bmatrix} \frac{1}{C_1 R_{x_1}} & \frac{1}{C_1} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) \\ -\frac{1}{C_2 R_{x_1}} & \frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (10)$$

$$A_3 = \begin{bmatrix} -\frac{1}{C_1 R_{x_1}} & -\frac{1}{C_1} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) \\ \frac{1}{C_2 R_{x_1}} & \frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (11)$$

$$A_4 = \begin{bmatrix} -\frac{1}{C_1 R_{x_1}} & \frac{1}{C_1} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) \\ -\frac{1}{C_2 R_{x_1}} & \frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (12)$$

$$A_5 = \begin{bmatrix} \frac{1}{C_1 R_{x_1}} & -\frac{1}{C_1 R_{x_1}} \\ \frac{1}{C_2} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) & -\frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (13)$$

$$A_6 = \begin{bmatrix} \frac{1}{C_1 R_{x_1}} & \frac{1}{C_1 R_{x_1}} \\ -\frac{1}{C_2} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) & -\frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (14)$$

$$A_7 = \begin{bmatrix} -\frac{1}{C_1 R_{x_1}} & -\frac{1}{C_1 R_{x_1}} \\ \frac{1}{C_2} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) & \frac{1}{C_2 R_{x_2}} \end{bmatrix}, \quad (15)$$

$$A_8 = \begin{bmatrix} -\frac{1}{C_1 R_{x_1}} & \frac{1}{C_1 R_{x_1}} \\ -\frac{1}{C_2} \left(\frac{1}{R_{x_2}} + \frac{1}{R_{x_3}} \right) & \frac{1}{C_2 R_{x_2}} \end{bmatrix} \quad (16)$$

where R_{x_i} represents the x terminal parasitic resistance of the i^{th} CCCII or ICCCI and which simulates bias current tunable resistor. Thus, a total of eight different oscillator circuits are realized (corresponding to each A_k matrix). Matrices $A_1 - A_4$ are grouped into class A and matrices $A_5 - A_8$ are grouped into class B, since all the matrices in a particular class have the similar parameters which differ only in their signs, i.e. $|a_{ij}|$ is same for all matrices in a class (A or B). For class A, it is sufficient to give MO-CCCI based realization of any one of the matrix in the class and the rest can be derived by simple interchange of sign of the output current, i.e. by utilizing either $z+$ or $z-$ terminal of MO-CCCI. Class B offers some more interesting circuits which cannot be directly created using CCCII and the use of ICCCI is required. It is sufficient to give circuit realization of any one matrix from A_5/A_8 pair and from A_6/A_7 pair (and the circuit realization of the other matrix in a pair can be easily obtained by simple interchange of $z+$ or $z-$ terminal). It is interesting to note that all matrices within class B necessitate the use of floating resistor simulators R_{x_1} and R_{x_2} ; further, circuit realizations of pair A_6/A_7 pair would also necessitate the use of

ICCCII . Showing only the necessary and sufficient realizations, we choose matrix A_1 , A_5 and A_6 as example matrices and provide the corresponding MO-CCCII or MO-ICCCII based oscillator realization while employing only grounded capacitors. The resulting circuits corresponding to matrix A_1 , A_5 and A_6 are shown in Fig. 1, Fig. 2 and Fig. 3, respectively.

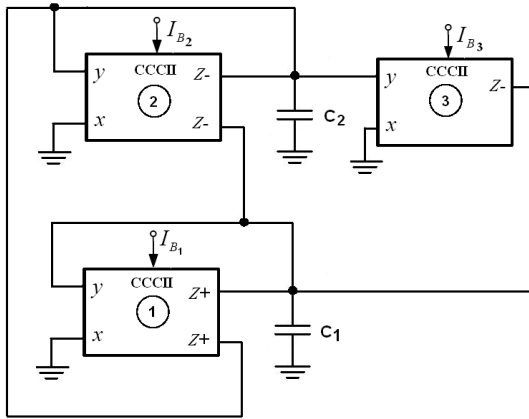


Fig. 1. Oscillator derived from A_1 .

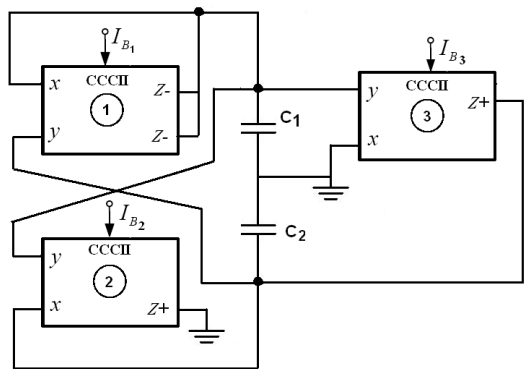


Fig. 2. Oscillator derived from A_5 .

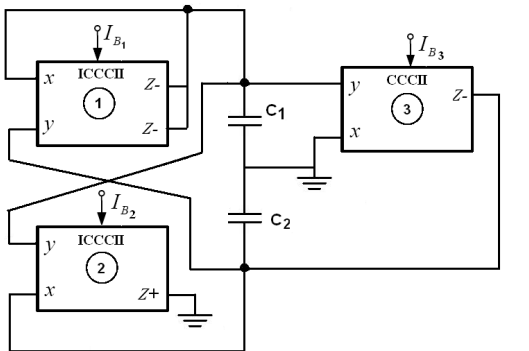


Fig. 3. Oscillator derived from A_6 .

3. Discussion on Parasitics Influence

Since the main purpose of the paper is to use state-variable method in realizing “resistor-less” second-order oscillators, we very briefly describe the effects of CCCII parasitics by taking one particular example.

Consider the circuit shown in Fig. 1, derived from matrix A_1 . The parasitic capacitances C_{y2} , C_{z2} and C_{y3} appear in parallel with the external capacitor C_1 and similarly, the parasitic capacitances C_{y1} , C_{z1} , C_{z2} and C_{z3} appear in parallel with the external capacitor C_2 . Thus, it is required that the value of the external capacitors should be larger than the parasitics appearing in shunt to alleviate their effects. Also, parasitic output resistances appear between the high output impedance z terminals and ground and they appear in shunt with the external capacitors. Since it is required that this impedance remains capacitive in nature, the effects can be alleviated by considering the operating frequency $f_o \gg \frac{1}{2\pi C_i R_{zi}}$ (where $i = 1, 2$) and C_i represents the external capacitor and R_{zi} represents the net parasitic resistance in shunt with the external capacitor. As is evident, this directly limits the low frequency potential of the circuits. It should also be noted that the presence of parasitic capacitances in parallel with the external capacitors modify the CO (since both capacitor terms are directly involved) and thus depending on the values of external capacitors and parasitic capacitances appropriate start-up margin should be considered for the oscillator to start. As pointed earlier, choosing external capacitors much larger than parasitic capacitances is a good practise. The non-ideal analysis and effects of parasitics is not considered for each and every circuit; however, it is believed that the aforementioned parasitic effects can be generally applied to any circuit and can bring about important results regarding the choice of external passive components or frequency potential.

4. Simulation Results

The proposed oscillator circuits have been simulated in PSPICE using a possible MOSFET implementation of CCCII, as shown in Fig. 4 (depicting a dual complimentary output CCCII). The oscillator circuit shown in Fig. 2 is taken as the design example. The circuit is implemented using $0.35 \mu\text{m}$ TSMC CMOS technology [22] with ± 2.5 voltage supplies. From Fig. 4, it can easily be derived that the parasitic resistance at terminal x is related to the bias current I_B according to the following equation

$$R_x = \sqrt{\frac{1}{8kI_B}} \quad \text{where}$$

$$k = \mu_p C_{ox} \left(\frac{W}{L}\right)_{1,2} = \mu_n C_{ox} \left(\frac{W}{L}\right)_{3,4}. \quad (17)$$

The aspect ratios of the transistors in for CCCII are indicated in Tab. 1. The circuit is designed with capacitor values of $C_1 = C_2 = 1 \text{ nF}$. The bias currents are taken to be

$I_{B1} = I_{B3} = 150 \mu A$ and $I_{B2} = 149 \mu A$ for the start-up of oscillations (so that the loop gain is greater than unity and the oscillations build). Note, that no external auxiliary amplitude control circuitry is used to stabilize the amplitude. The amplitude is inherently limited due to the non-linearity of the active device. The start-up of oscillations and the steady-state waveforms for the voltages across the capacitors are shown in Fig. 5 and 6, respectively. It should also be noted that external voltage buffers are needed to use the voltages across the capacitors without loading the circuit, but, such an overhead cost is inevitable in any current conveyor based oscillator (since they do not offer any buffered terminals for voltage outputs). The observed frequency of 499 kHz is in close correspondence with the theoretical value of 510.43 kHz. The magnitudes of the harmonics obtained using Fourier analysis are shown in Tab. 2, with the total harmonic distortion (THD) of 0.054 % (in terms of power) and 2.42 % (in terms of voltage). The variation of FO with bias current I_{B3} is shown in Fig. 7, indicating the independent electronic tuning of the oscillator. Note that value of THD for the performed experiment may not be acceptable

for some applications in which case auxiliary amplitude control should be used. A very simple way (a very commonly used technique) of regulating the amplitude is to increase I_{B2} with increase in amplitude and thereby reducing the positive feedback action. This can be done by sensing the growing amplitude by means of a peak-detector circuit and then increasing the bias voltage from which I_{B2} is derived. This results in reduced amplitude of oscillation and thus reduced effect of the device non-linearity; yielding a better THD performance. Similar methods have been used in recent communications in [23]-[24].

MOSFET	W/L ($\mu m/\mu m$)
M1-M2	45/0.35
M3-M4	15/0.35
M5-M6, M11-M12, M14, M16-M17	30/0.35
M7-M8, M9-M10, M13, M15, M18-M19	10/0.35

Tab. 1. Transistor aspect ratios for CCCII.

	Freq (Hz)	Mag	Norm. Mag
1	0.0 (offset)	-28.905m	-
2	499 k	328.007 m	1.000
3	998 k	7.388 m	22.524 m
3	1.497 Meg	1.302 m	3.971 m
4	1.996 Meg	668.851 u	2.039 m
5	2.495 Meg	399.092 u	1.217 m
6	2.994 Meg	295.163 u	899.867 u
7	3.493 Meg	616.168 u	1.880 m
9	3.992 Meg	348.837 u	1.067 m
10	4.491 Meg	202.488 u	617.327 u

Tab. 2. Fourier analysis.

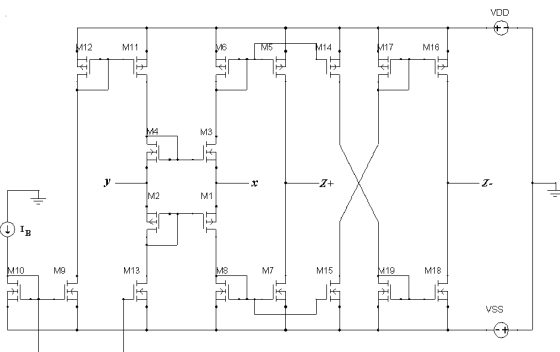


Fig. 4. A possible CMOS implementation of complementary output CCCII.

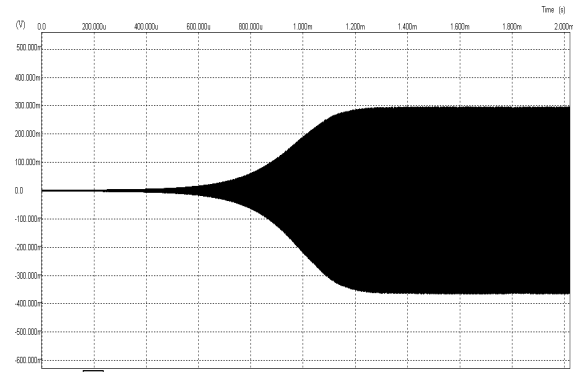


Fig. 5. The start-up of oscillations.

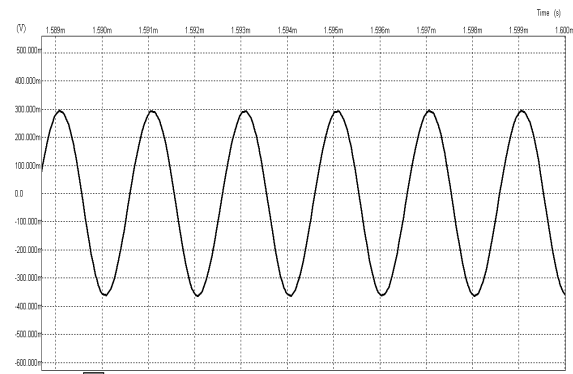


Fig. 6. The steady-state oscillation waveform.

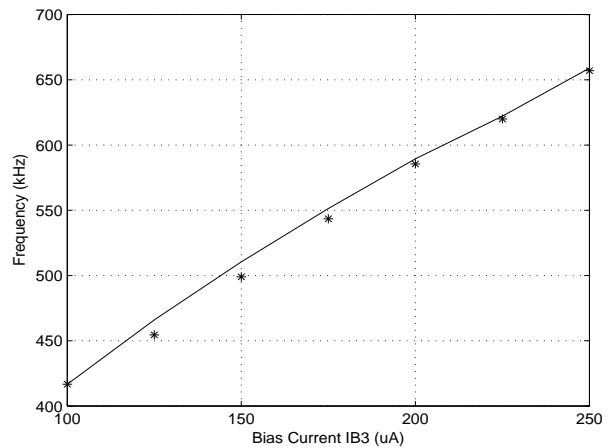


Fig. 7. The variation of FO with I_{B3} .

5. Concluding Remarks

State variable method is a powerful technique in realizing sinusoidal oscillators with priori known tuning laws for the CO and the FO. This paper exhibits the usefulness of state variable method in realizing MO-CCCI based “resistor-less” sinusoidal oscillators with non-interactive (independently controllable) tuning laws for the CO and the FO. Eight different matrices are systematically developed from the information of the targeted CO and FO and correspondingly eight different oscillator circuits are derived, all of which employ grounded capacitors and no more than three MO-CCCIIs and therefore leading to canonic realizations. The method is direct, can be further extended to other oscillator topologies governed by different tuning laws and helps in systematic evolution of oscillators.

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