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SYSTEMATIC SYNTHESIS FOR ELECTRONIC-CONTROL LC OSCILLATORS USING SECOND ORDER CURRENT-CONTROLLED CONVEYOR

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Key words: *LC* oscillator, Electronic control, Current-controlled conveyor, Nodal admittance matrix expansion.

According to behavioral models of the second order current-controlled conveyor (CCCII) and using the nodal admittance matrix (NAM) expansion method and the adjoint network theorem, the family of double-mode quadrature LC oscillators employing CCCIIs is synthesized. It contains three different classes. The class A oscillator employs four CCCIIs as active element and possesses 32 different forms, the class B oscillator employs three CCCIIs and possesses 16 different forms, and the class C oscillator, employing four CCCIIs or two CCCIIs and one dual-output CCCII, possesses 32 different forms. In all, 80 LC oscillators using CCCIIs are obtained. By the aid of CCCII-based simulating grounded inductors, 80 LC oscillators are extended into 320 ones. Because of using grounded capacitors, the circuits can be easily integrated and the oscillation criterion, the oscillation frequency, and the 3 dB bandwidth can be independently, linearly, and electronically tuned by tuning bias currents of the CCCIIs. The Pspice simulation data match the hand analysis results, which match the synthesized circuits.

1. INTRODUCTION

It is well known that the traditional *LC* oscillator uses a discrete transistor as active element, a parallel *LC* resonant circuit as frequency-selective network/feedback network, but has the potential of being very high frequency oscillator and has been widely applied in chaos circuit design [1–5]. However, since it employs metal coil, adjust of oscillation frequency is complicated; since it does not employ advanced active devices, such as the current-controlled conveyor (CCCII), inverting CCCII (ICCCII), operational transconductance amplifier (OTA), current controlled current conveyor trans-conductance amplifier (CCCCTA), et al, as active element, the oscillation criterion cannot be electronically controlled. In fact, there are no CCCII -based *LC* oscillators to be cited in this paper. Therefore, this is a problem to be researched further.

On the other hand, the NAM expansion method has found wide applications since it was put forward [6–8]. Very recently, this method has been used in the synthesis of circuits employing CCCIIs [9], but the reported circuits include only gyrators rather than oscillators, especially *LC* oscillators. Because the CCCII has attracted considerable attention and a number of CCCII-based filters and oscillators have been reported [10–13], it is necessary that using the NAM expansion method synthesizes CCCII-based circuits except gyrators.

The primary objective of this paper is to utilize the NAM expansion approach to synthesize *LC* oscillators employing CCCIIs. First, according to an original *LC* oscillator, we derive its NAM stamp, from which the oscillators are classified as type A, type B, and type C. Next, making use of the NAM expansion method, the adjoint network theorem [14, 15], and behavioral models of the CCCII, three different classes of the double-mode quadrature *LC* oscillators are synthesized. The type A oscillator, utilizing four CCCIIs, has 32 different forms. The type B oscillator, utilizing three CCCIIs, has 16 different forms. The type C oscillator, utilizing four CCCIIs or two CCCIIs and one dual-output CCCII (DOCCCII), has 32 different forms. 80 *LC* oscillators are obtained in all. Moreover, the grounded inductors in the derived oscillators are substituted by CCCII-based simulating inductors to produce 320 different forms of the double-mode quadrature oscillators. Also, by the aid of bias currents of the CCCIIs, we can independently, linearly, and electronically tune oscillation criterion (OC), oscillation frequency (OF), and the 3 dB bandwidth (BW). Finally, the validity of the synthesized circuit is verified by means of the paper and pencil analysis and the computer simulation.

2. BASIS CONCEPT OF *LC* OSCILLATORS

An ac equivalent circuit of the *LC* oscillator with an operational amplifier is shown in Fig. 1 [16]. A parallel *LC* resonant circuit is used to establish the oscillator frequency, and the feedback is provided by a conductance *G* and the *LC* resonant circuit. The conductance *G* is used to control 3 dB bandwidth of the loop. We assume that the operational amplifier is ideal. Routine analysis of the circuit gives the following equation:

$$V_1 = \frac{1/sC\,//\,sL}{1/G + 1/sC\,//\,sL} V_2 . \quad (1)$$

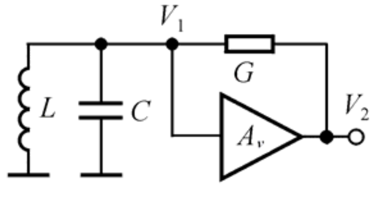

Fig. 1 – An original *LC* oscillator circuit.

It can be rearranged as:

$$(G + sC + 1/sL)V_1 - GV_2 = 0 . \quad (2)$$

Imposing $A_v = V_2/V_1 = G_3/G_4$, then

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$$-G_3V_1+G_4V_2=0\,. \tag{3}$$

From (2) and (3), the state equation is

$$\begin{bmatrix} G+sC+1/sL & -G \\ -G_3 & G_4 \end{bmatrix}\begin{bmatrix} V_1 \\ V_2 \end{bmatrix}=0\,. \tag{4}$$

The NAM matrix of the oscillator is then

$$Y=\begin{bmatrix} G+sC+1/sL & -G \\ -G_3 & G_4 \end{bmatrix}. \tag{5}$$

The characteristic equation of the oscillator is $|Y| = 0$, or

$$s^2+sG(1-G_3/G_4)/C+1/LC=0\,. \tag{6}$$

Therefore, the OC and the OF are

$$G_3 \ge G_4\,, \tag{7}$$

$$f_o=\frac{1}{2\pi\sqrt{LC}}\,. \tag{8}$$

Also, from (2) and (3), we can obtain the loop gain (LG) as follows

$$\mathrm{LG}=\frac{G_3}{G_4}\times\frac{sG/C}{s^2+sG/C+1/LC}\,. \tag{9}$$

This is a band-pass filter. Its BW, determining selectivity, is then

$$\mathrm{BW}=\frac{G}{C}=G\sqrt{\frac{L}{C}}\omega_o<\omega_o\,. \tag{10}$$

So long as G is taken small, the circuit has rich selectivity, producing less distortion. Therefore, the band-pass network should employ an active network so as to make the BW quite narrow. Then any distortion introduced by the amplifier can be filtered by the band-pass network.

Implicit in equations (7), (8) and (10) is that adjusting G_3 or G_4 can linearly turn the OC, and trimming L, if it is replaced by a simulating inductance, can linearly adjust the OF. Whereas the BW can be tuned quite narrow by adjusting G without affecting the OC and OF. This means that the oscillator can provide the attractive feature of electronically independent control of the OC, OF, and BW.

3. SYSTEMATIC SYNTHESIS OF *LC* OSCILLATORS

In accordance with the different stamps of the expanded NAM matrix, *LC* oscillators to be synthesized are classified into three different types and the NAM expansion method for three different types of the *LC* oscillators will be developed.

3.1. TYPE A OSCILLATOR

Configure 1. On the basis of the NAM expansion method, starting from (5), and taking into account type A oscillator with six nodes, the first step to expand is to add four blank rows and columns, and then use a first nullator to link columns 1 and 3 to move G_3 to the position 2, 3. The first current mirror is connected between rows 2 and 3 to move G_3 to the position 3, 3 with inverted sign. A second nullator is then connected columns 2 and 4 to move G_4 to the position 2, 4. A first norator is connected between rows 2 and 4 to move G_4 to the position 4, 4. A third nullator is then connected columns 2 and 5 to move G to the position 2, 5. A second current mirror is connected between rows 1 and 5 to move G to the position 5, 5 with inverted sign. A forth nullator is then connected columns 1 and 6 to move G to the position 1, 6. A second norator is connected between rows 1 and 6 to move G to the position 6, 6. The NAM matrix, including the added nullor-mirror elements represented by bracket notation, is displayed in (11).

Implicit in (11) is that the expanded matrix contains four different pairs of pathological elements, one grounded capacitor and one grounded inductance between node 1 and ground, and four grounded admittances, namely G_3, G_4 and two G.

Shown in Fig. 2 is the nullor-mirror equivalent circuit model for (11). Making use of the nullor-mirror descriptions for CCCII [16] and bearing Fig. 2 in mind, an equivalent CCCII-based implementation can be achieved, as exhibited in Fig. 3.

$$Y=\begin{bmatrix} sC+1/sL & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & G_3 & 0 & 0 & 0 \\ 0 & 0 & 0 & G_4 & 0 & 0 \\ 0 & 0 & 0 & 0 & G & 0 \\ 0 & 0 & 0 & 0 & 0 & G \end{bmatrix}. \tag{11}$$

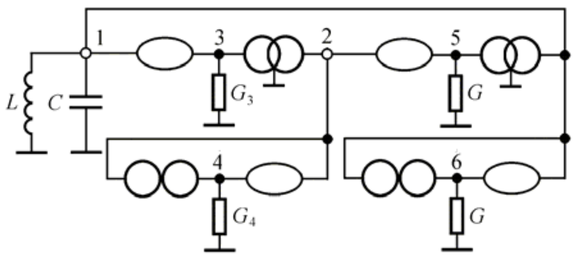

Fig. 2 –Equivalent circuit model constructed by (11).

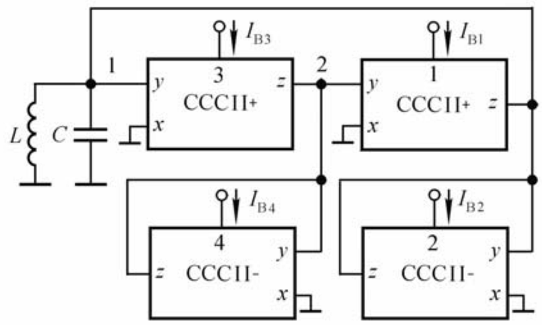

Fig. 3 – CCCII implementation for Fig.2.

Because expanding the matrix should have 16 possible combinations of the added nullor-mirror elements, the equivalent circuit models have 16 forms and CCCII-based implementations have 16 forms too. Figures 2 and 3 are one of them, respectively.

Configure 2. According to the adjoint network theorem, by replacing the nullator by a norator and the current mirror by a voltage mirror, vice versa, the circuit in Fig. 2 is transformed to the circuit in Fig. 4.

The corresponding equivalent CCCII-based realization is shown in Fig. 5. Notice that in Figs. 3 and 5: $I_{B1} = I_{B2} = I_B$, $G = 2I_B/V_T$, $G_3 = 2I_{B3}/V_T$, $G_4 = 2I_{B4}/V_T$.

Similarly, applying the adjoint network theorem, the other 15 models relative to the circuit in Fig. 2 can also be tuned into the corresponding adjoint circuits.

We observe that the type-A oscillator circuits, employing four CCCIIs, possess 32 different forms.

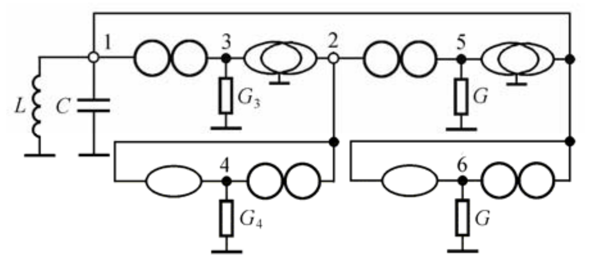

Fig. 4 – Adjoint circuit model from Fig. 2.

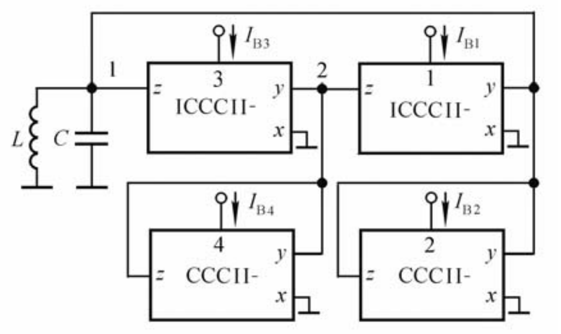

Fig. 5 – CCCII implementation for Fig.4.

3.2. TYPE B OSCILLATOR

Configure 1. The synthesis process of type B oscillator proceeds in the same way as the previous synthesis process. Beginning from (5) and applying all possible combinations of the added nullor-mirror elements will produce the following eight different forms of the expanded matrixes, whose one form is exhibited in (12). The corresponding nullor-mirror equivalent circuit model is depicted in Fig. 6, whereas the CCCII-based implementation for Fig. 6 is depicted in Fig. 7.

$$Y=\begin{bmatrix} G+sC+1/sL & 0 & 0 & 0 & -G \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & G_4 & 0 & 0 \\ 0 & 0 & 0 & G_3 & 0 \\ -G & 0 & 0 & 0 & G \end{bmatrix}. \quad (12)$$

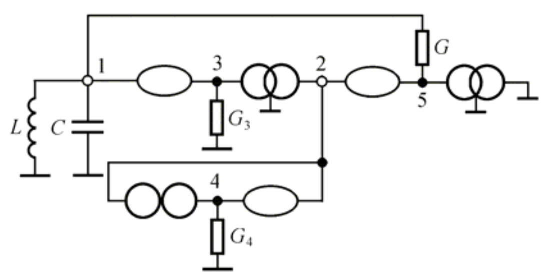

Fig. 6 –Equivalent circuit model constructed by (12).

Implicit in (12) is that the expanded matrix contains three different pairs of pathological elements, one grounded capacitor and one grounded inductance between node 1 and ground, one floating admittance between nodes 1 and 4, and two grounded admittances, namely G_3, G_4.

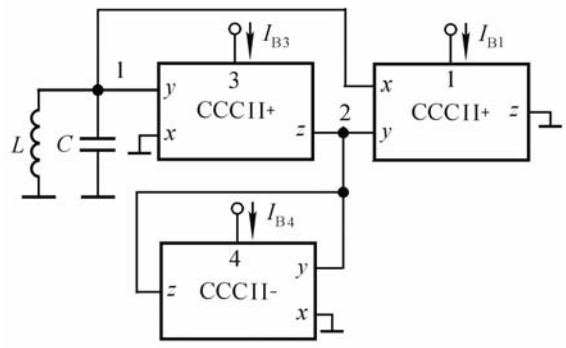

Fig. 7 – CCCII implementation for Fig.6.

Configure 2. According to the adjoint network theorem, eight models relative to the circuit in Fig.6 can also be tuned into the corresponding adjoint circuits and eight CCCII-based implementations can also be obtained. Fig. 8 and Fig. 9 are one of them, respectively.

Notice that in Figs. 7 and 9: $G = 2I_{B1}/V_T$, $G_3 = 2I_{B3}/V_T$, $G_4 = 2I_{B4}/V_T$.

We see that the type B oscillator circuits, employing three CCCIIs, possess 16 different forms.

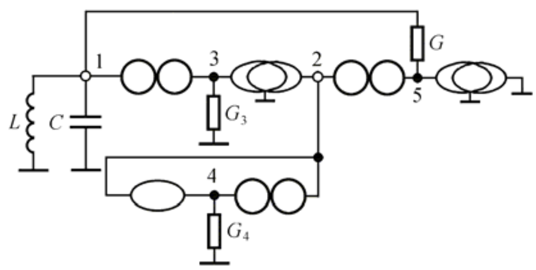

Fig. 8 – Adjoint circuit model from Fig. 6.

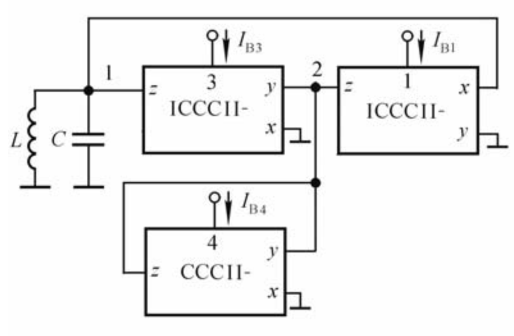

Fig. 9 – CCCII implementation for Fig. 8.

3.3. TYPE C OSCILLATOR

Configure 1. Proceeding in similar fashion, we expand the NAM matrix (5) via all possible combinations of added nullor-mirror elements, producing 16 alternative expanded matrixes, one of them is depicted in (13).

Implicit in (13) is that the expanded matrix contains four different pairs of pathological elements, one grounded capacitor and one grounded inductance between node 1 and ground, one floating admittance between nodes 5 and 6, and two grounded admittances, namely G_3, G_4.

Similarly, the equivalent circuit model and CCCII-based implementation are given respectively by Fig. 10 and Fig. 11, which are one of 16 counterparts, respectively. Notice that in Fig. 11: $I_{B1} = I_{B2} = I_B$,

$$G = 1/(R_{x1} + R_{x2}) = I_B/V_T,$$

$$G_3 = 2I_{B3}/V_T,\ G_4 = 2I_{B4}/V_T.$$

$$Y = \begin{bmatrix} sC+1/sL & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & G_3 & 0 & 0 & 0 \\ 0 & 0 & 0 & G_4 & 0 & 0 \\ 0 & 0 & 0 & 0 & G & -G \\ 0 & 0 & 0 & 0 & -G & G \end{bmatrix}. \qquad (3)$$

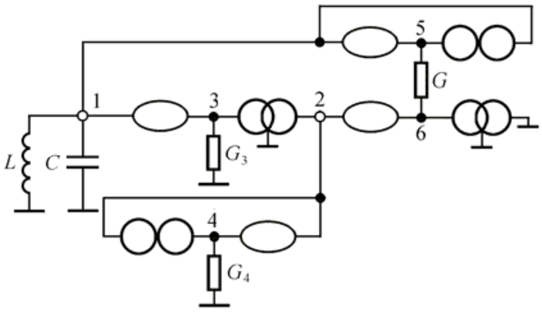

Fig. 10 – Equivalent circuit model constructed by (13).

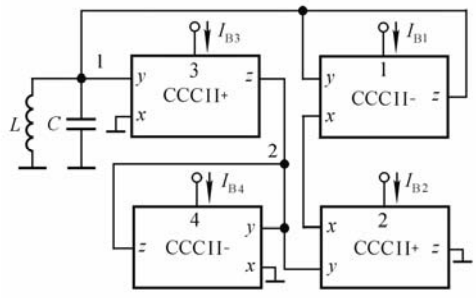

Fig. 11 – CCCII implementation for Fig.10.

Configure 2. Applying again the adjoint network theorem, 16 models relative to the circuit in Fig. 10 can also be tuned into the corresponding adjoint circuits and 16 CCCII-based implementations can also be obtained. Fig. 12 and Fig.13 are one of them, respectively. Notice that in Fig. 13: $G = 2I_{B1}/V_T$, $G_3 = 2I_{B3}/V_T$, $G_4 = 2I_{B4}/V_T$.

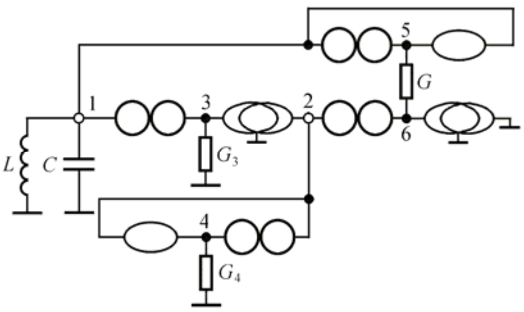

Fig. 12 –Adjoint circuit model from Fig. 10.

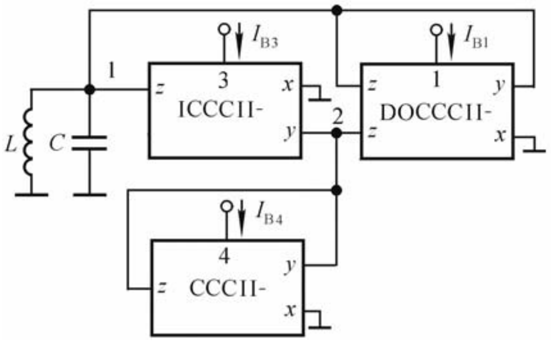

Fig. 13 – CCCII implementation for Fig. 12.

We observe that the type-C oscillator circuits employing either four CCCIIs (two of them have floating x teminals) or two CCCIIs and one DOCCCII, possess 32 different forms.

It can be readily observed that the LC oscillators using CCCIIs possess three classes and have 80 different forms.

4. SYSTEMATIC SYNTHESIS OF *LC* OSCILLATORS WITH ONLY GROUNDED CAPACITORS

In order for the derived 80 oscillators to contain no grounded inductance, we can substitute grounded inductors in the circuits by CCCII-based simulating grounded inductors. The literature [9] has reported CCCII-based simulating grounded inductors, four of which, using two CCCIIs, are the most simple because they employs two CCCⅡs and one grounded admittance, and do not require any matched conditions. Combining the 80 oscillators with four simulating grounded inductors in the literature [9] will produce 320 different forms of the LC oscillators that employ least amount of active and passive components, one of which is shown in Fig. 14, which is constructed by Fig. 7 and one of the four simulating grounded inductors, its inductance is given by

$$L_{eq} = \frac{V_T^2 C_L}{4I_{B2}I_{B5}}. \qquad (14)$$

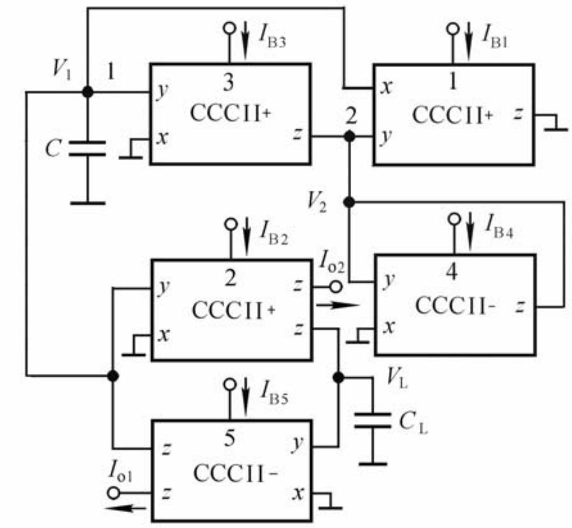

Fig. 14 – One form of 320 CCCII-based oscillators with only grounded capacitors.

The remaining implementations are omitted to limit length of the paper.

5. CIRCUIT ANALYSIS

As an example of the *LC* oscillator's analysis, we consider only the circuit in Fig. 14. Breaking the loop at the terminal *y* of $CCCII_3$ and injecting a test signal V_t. As this signal propagates around the loop, it comes back as return signal V_r. The LG = V_r / V_t is the same as (9). The BW, after using $G = 2\,I_{B1}/V_T$, is

$$\mathrm{BW} = \frac{2I_{B1}}{V_T C}. \qquad (15)$$

The characteristic equation of the oscillator is 1-LG = 0, which is the same as (6). The OC and the OF, from (7)–(8), (14), and $G_3 = 2\,I_{B3}/V_T$, $G_4 = 2\,I_{B4}/V_T$, are

$$I_{B3} \geq I_{B4}, \qquad (16)$$

$$f_o = \frac{I_B}{\pi V_T C}. \qquad (17)$$

Here, $C = C_L$, $I_B = I_{B2} = I_{B5}$. The oscillator is tuned as follows: (a) adjust I_{B1} to ensure lower BW; (b) adjust I_{B3} or I_{B4} to satisfy the OC; (c) adjust I_B to vary the OF for the desired value of f_o. It is intriguing that the OC and the OF could be independently tuned by adjusting bias current I_B and I_{B3} or I_{B4} and that the BW could be sustained quite narrow by adjusting I_{B1}.

For sinusoidal steady state, we can write, by inspection of Fig. 14,

$$V_L = I_{o2}/sC_L, \quad I_{o2} = G_2 V_1, \quad I_{o1} = -G_5 V_L. \qquad (18)$$

Combining the above equations and considering (16–17), the following transfer functions can be calculated as

$$\frac{V_1}{V_L} = \mathrm{j}, \quad \frac{I_{o1}}{I_{o2}} = \mathrm{j}. \qquad (19)$$

Equation (19) states that the oscillator can provide not only two quadrature current outputs with equal amplitude but also two quadrature voltage outputs with equal amplitude. Then double-mode quadrature oscillators employed CCCIIs and two grounded capacitors are obtained. The paper and pencil analysis has verified the synthesized circuits.

The analysis for other circuits is omitted, but the results have been tabulated, as shown in Table 1. It can be seen that the synthesized quadrature oscillators employ only grounded capacitors. It can also be seen that the class B oscillator employs only three CCCIIs and des not require any matched conditions, and its three parameters, OC, OF and BW, can be linearly, independently, and electronically tuned by trimming bias currents of the CCCIIs. Therefore, the class B oscillator is the best.

6. COMPUTER VERIFICATION

A Pspice simulation was performed using the circuit in Fig. 14, whose sub-circuit, the CCCII, was created by using the transistor model of PR200N and NR200N [12]. When $C = C_L = 1$ nF, $I_{B1} = 20.4$ µA, $I_{B3} = 100$ µA $\geq I_{B4} = 81.6$ µA, $I_B = 81.6$ µA, the design value for f_o, from (17), is 1 MHz, the design value for BW, from (15), is 0.25 MHz, $V_1/V_L = \mathrm{j}$, and $I_{o1}/I_{o2} = \mathrm{j}$.

Shown in Fig.15–18 are the simulation results, which gives $f_o = 971$ kHz, BW = 0.310 MHz, $V_1/V_L = \mathrm{j}$, and $I_{o1}/I_{o2} = \mathrm{j}$.

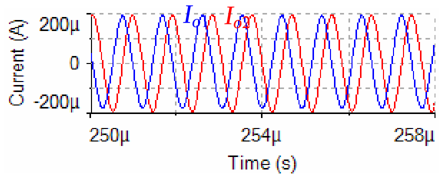

Fig. 15 – Expanded view of the current outputs for the design value of 1 MHz.

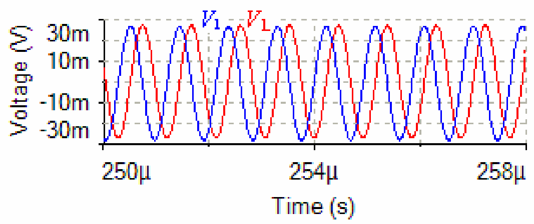

Fig. 16 – Expanded view of the voltage outputs for the design value of 1 MHz.

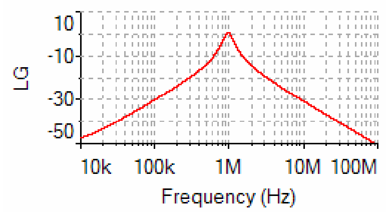

Fig. 17 – Frequency responses of the feedback loop for the design value of 1 MHz and $I_{B1} = 20.4$ µA.

Table 1

Properties for the three different types of synthesized oscillators

Class	No. of oscillators	No. of active devices	OC	OF	Independent control for OC and OF	BW	Initial conditions
A	32	Four CCCIIs	$I_{B3} \geq I_{B4}$	$I_B/\pi V_T C$	Yes	$2I_{B1}/V_T C$	$I_{B1} = I_{B2} = I_B$
B	16	Three CCCIIs	$I_{B3} \geq I_{B4}$	$I_B/\pi V_T C$	Yes	$2I_{B1}/V_T C$	No
C – Config.1	32	Four CCCIIs	$I_{B3} \geq I_{B4}$	$I_B/\pi V_T C$	Yes	$2I_{B1}/V_T C$	$I_{B1} = I_{B2} = I_B$, $G = I_B/V_T$
C – Config.2		Two CCCIIs and one DOCCCII					No

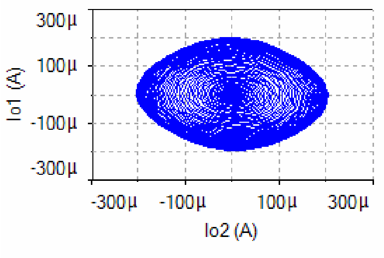

Fig. 18 – Lissajous figure formed by I_{o1} and I_{o2}.

The total harmonic distortions for I_{o2} and I_{o1} are 0.778 % and 1.32 %, respectively. Figure 19 shows only the simulated output spectrum for I_{o2}.

To explain the controllability of f_o by adjusting I_B, I_{B1}, I_{B3} and I_{B4} are kept as before. When I_B is tuned from 81.6 µA to 816 µA, the design value for f_o is changed from 1 MHz to 10 MHz. In Fig. 20 are given the transient responses of I_{o1}, where the simulation result of f_o is 7.90 MHz when $I_B = 816$ µA. The reason resulting in the error is mainly due to the effects of parasitic admittances from CCCIIs, but is not analyzed here to limit length of the paper.

The results of circuit simulations are out of question in agreement with theory.

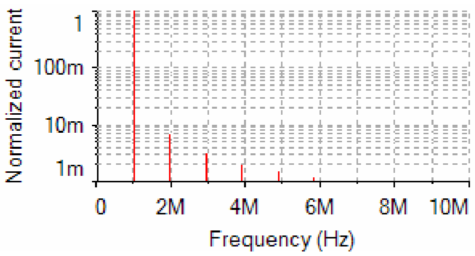

Fig. 19 – The output spectrum of I_{o2} for the design value of 1MHz.

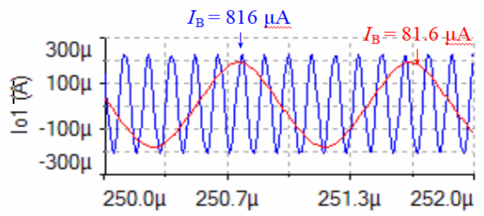

Fig. 20 – Controllability of f_o observed on transient responses.

7. CONCLUSIONS

In this paper, by the aid of the NAM expansion approach and the adjoint network theorem, we get 80 *LC* oscillators. By the aid of CCCII-based simulating inductors, 80 *LC* oscillators are extended into 320 ones. Needless to say, the simulated grounded inductance in these oscillators can be no better than capacitances, CCCIIs in their simulation. However, the synthesized double-mode quadrature oscillators enjoy many advantages, such as independent, linear, and electronical control of the OC, OF, and BW, use of grounded capacitors, use of least amount active device, no externally connected resistors, and so on. The results of hand analysis and simulation have verified the synthesis method involved.

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