Current-Mode CCII+ Based Oscillator Circuits using a Conventional and a Modified Wien-Bridge with All Capacitors Grounded

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Abstract: *The paper deals with a pair of current-mode sine-wave oscillator circuits. Both these circuits are implemented using positive second-generation current conveyors (CCII+). The principle of the first oscillator is based on a conventional Wien-bridge network. However, this implementation suffers from the use of a floating capacitor, which can be unacceptable in the case of on-chip integration. This drawback is solved in the second variant via a slight modification of the Wien-bridge network, which then allows the use of all capacitors grounded. The modified circuit version was manufactured by means of the so-called diamond transistors, which play the role of CCII+ active building blocks. The circuit behavior was analyzed theoretically, with particular emphasis on the identification of real effects and their elimination, and subsequently verified experimentally. The experimental results are included in the paper.*

Keywords

Diamond transistor, Wien-bridge oscillator, current-mode, current conveyor.

1. Introduction

Sinusoidal oscillator is a basic and one of the most common applications in analog signal processing. There is a wide range of conceptions and various approaches in terms of principle and utilized active elements. In the case of current-mode (CM) regime, a current conveyor (CC) is frequently employed as the active element. For example, several oscillators are implemented using first-generation current conveyors CCI in [1], [2]. Rather than CCIs, second-generation current conveyors, CCIIs, (usually of the positive type) are used in oscillator circuits [3]-[8]. The Wien-bridge oscillator (WBO) is a specific type. Two different WBOs using two CCIIs+ are reported in [6]. In addition to two CCIIs+, eight passive components are employed there. Another two papers, [7] and [8], are also focused on CCII+ based WBOs. All the WBOs published in [7] require a CCII+ with multiple Z-terminals for their operation (or for the explicit current output, if you like). Oscillator circuits in [8] employ CCII+ but the output signal is available only in the form of voltage. In addition, all the circuits published in [8] contain one floating capacitor. An economical version of the WBO employing one CFA (Current-Feedback Amplifier) and only four passive components with two capacitors grounded, is published in [9] and [10]. Owing to the CFA internal topology, these oscillators are of the CCII1 type, with buffered voltage outputs. Their key drawback consists in the impossibility of independent control of the oscillation frequency (OF) and oscillation condition (OC). The WBOs from [11] and [12] represent conventional oscillator structures employing one non-inverting voltage amplifier and the Wien bridge in a feedback loop. The amplifier is implemented by one CFA and two resistors. In spite of some important advantages of the CFA-based oscillators over their VFA (Voltage Feedback Amplifier) counterparts [11], these oscillators also provide only voltage outputs. CFA-based oscillators with explicit current outputs are proposed in [13] (Fig. 4). However, the load resistance affects the oscillator behavior.

The purpose of the present paper is to propose another two WBO realizations with explicit current outputs, each of them utilizing two CCIIs+ and six passive components (four resistors and two capacitors). The paper deals with two different explicit current-output WBO circuits. Both of them are based on the so-called diamond transistors (DT) [14]. DT is a part of the commercially available OPA860 integrated circuit. DT with emitter (E), base (B), and collector (C) terminals represents a CCII+ with terminals X, Y and Z (Fig. 1). DTs have successfully been employed in many applications such as more advanced active building blocks. DTs were successfully used, for example, in [15] to implement ZC-CG-CDBA (Z Copy–Controlled Gain-Current Differencing Buffered Amplifier) or in [16] to implement CIBTA (Current Inverting Buffered Transconductance Amplifier), ZC-CITA (Z Copy-Current Inverting Transconductance Amplifier) in [17], and FB-VDBA (Fully Balanced-Voltage Differencing

Buffered Amplifier) in [18]. A number of active elements and the methodology for constructing them using DTs were published in [19].

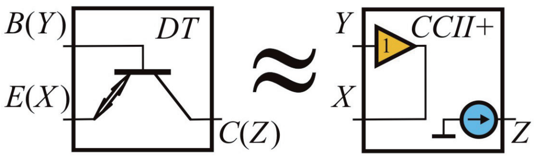

Fig. 1. Diamond transistor OPA860 as a CCII+.

2. Proposed Circuits

Fig. 2 shows a proposed CM oscillator using two CCIIs+ and a conventional Wien bridge network. Note that DT_1 in the role of CCII+ together with the voltage buffer, operating between the base-emitter terminals of DT_2, acts as a conventional CFA. From this point of view, resistors R_2 and R_4 adjust the gain of the non-inverting amplifier, which is a standard block of the WBO. However, due to the double use of two diamond transistors, not only voltage but also current output is available.

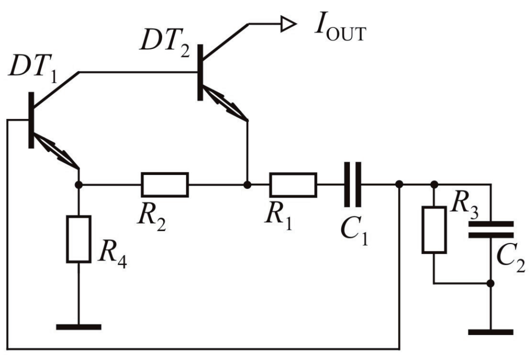

Fig. 2. CCII+ based Wien-bridge oscillator.

A potential drawback of this circuit consists in the floating capacitor C_1, which can cause a problem in the case of integration on a chip. It can be overcome with the help of an idea from [13] and [20], where a derivation of Wien-bridge equivalent forms is described. A practical solution is achieved by interchanging complete impedances between two branches, namely a serial combination of R_1 and C_1 with R_4. The resulting circuit shown in Fig. 3 is potentially suitable for on-chip integration since all capacitors are grounded, concurrently enabling explicit current output. Note that this oscillator resembles a topology published in [13] (in Fig. 4 (b) therein): In [13], CFA is used instead of DT_1 and DT_2. However, the output current is flowing from the z-terminal of the CFA to an external load, labeled as R_{load} in [13]. In Fig. 3, it would be represented by a connection of grounded resistor R_{load} to the node between the collector of DT_1 and the base of DT_2. In this way, the value of R_{load} would modify the amplifier gain and the oscillation condition. By contrast to this case, the topology in Fig. 3 does not suffer from this drawback; since the current output is available at the high-impedance collector node of DT_2.

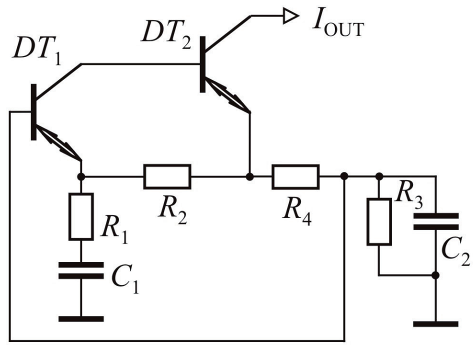

Fig. 3. Modified CCII+ based Wien-bridge oscillator with all capacitors grounded.

The characteristic equation (CE) is the same for both variants in Figs 2 and 3:

CE: $$s^2 C_1 C_2 R_1 R_3 R_4 + s(C_1 R_1 R_4 + C_2 R_3 R_4 - C_1 R_2 R_3) + R_4 = 0 . \quad (1)$$

The oscillation conditions (OC) and the oscillation frequency (OF) are as follows:

OC: $$\frac{R_2}{R_4} \geq \frac{R_1}{R_3} + \frac{C_2}{C_1} , \quad (2)$$

OF: $$\omega_0 = \frac{1}{\sqrt{C_1 C_2 R_1 R_3}} . \quad (3)$$

Under the conditions $C_1 = C_2 = C$ and $R_1 = R_3 = R$, OC and OF can be simplified to the forms:

OC: $$\frac{R_2}{R_4} \geq 2 , \quad (4)$$

OF: $$\omega_0 = \frac{1}{RC} . \quad (5)$$

3. Analysis of Real Effects

In reality, both OC and OF differ from their ideal forms (2) and (3) due to non-ideal effects of the active devices used or due to imperfections of technological processes in the case of on-chip implementation.

As mentioned above, the second oscillator circuit was implemented using the OPA860 DTs for experimental purposes. Major real influences are described in the DT datasheet [14]. For simplicity, consider that the circuit will operate at a constant temperature and within the bandwidth of DTs used and thus the frequency and temperature dependence may not be analyzed. The parasitic capacitances and resistances of the collector and base terminals, C_c, R_c, and C_b, R_b, and the parasitic non-zero resistance of emitter terminal R_e are the major non-idealities. A simple modeling of actual DT effects is shown in Fig. 4. Typical parameters according to [14] are R_c= 54 kΩ, C_c= 2 pF, R_b= 445 kΩ, and C_b= 2.1 pF. Emitter input resistance R_e depends on a bias current I_{ADJ}, which must be set via an external resistor R_{ADJ} (see [14]). A typical R_e value is between 9 Ω and 11 Ω. The resistor R_{ADJ}= 330 Ω was used during all experiments, giving R_e resistances of about 10 Ω.

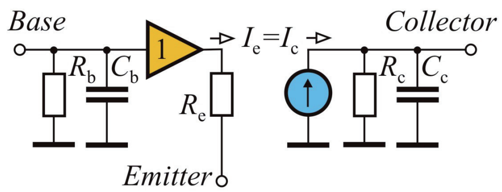

Fig. 4. A simple model of major DT non-idealities.

With all the above real effects taken into consideration, the resulting model of the oscillator from Fig. 3 is shown in Fig. 5, where R_Z is formed by a parallel combination of R_{c1} and R_{b2}, and capacitor C_Z represents a parallel connection of C_{c1} and C_{b2}. The same situation is in the case of resistor $R_3' = R_3 \| R_{b1}$ and capacitor $C_2' = C_2 \| C_{b1}$. A proper choice of component values can cause that parasitic parameters R_{b1} and C_{b1} do not affect R_3 and C_2 in any way (when R_{b1} is much higher than R_3 and C_{b1} is much lower than C_2). Their influence can then be neglected. We assume this condition and thus we use R_3 and C_2 instead of R_3' and C_2' in the following analysis.

The schematic in Fig. 5 serves as a starting point of the non-ideal analysis. Taking all the above real effects into consideration, the corresponding CE has a more general form:

CE: $$a_3 s^3 + a_2 s^2 + a_1 s + a_0 = 0 . \qquad (6)$$

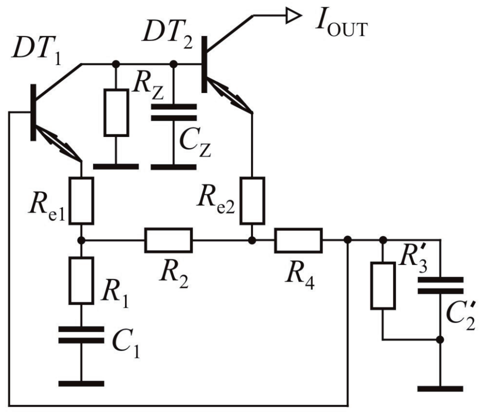

Fig. 5. Modified WBO version with major real influences taken into consideration.

The coefficients a_0 to a_3 of the CE are rather complex, composed of the summation of several terms. Some of them have a significant impact on the total coefficient value, but some of them are insubstantial and can be neglected. If they are neglected, the simplified forms of the coefficients are as follows:

$$a_0 = R_2R_4 + R_4R_Z + R_2R_3 ,$$
$$a_1 = C_ZR_Z(R_2R_4 + R_2R_3 + R_4R_{e1} + R_4R_{e2} + R_2R_{e2}) + \\ + C_1(R_1R_4R_Z - R_2R_3R_Z + R_1R_2R_4 + R_1R_2R_3 + R_2R_3R_{e1} + \\ + R_2R_4R_{e1} + R_1R_2R_{e2}) + C_2R_3(R_4R_Z + R_2R_4 + R_2R_{e2}),$$
$$a_2 = C_1C_2R_1R_3R_4(R_2 + R_Z) + C_2C_ZR_2R_3R_4R_Z + \\ + C_1C_ZR_2R_Z(R_1R_3 + R_1R_4 + R_4R_{e1}),$$
$$a_3 = C_1C_2C_ZR_2R_3R_Z(R_1R_4 + R_4R_{e2} + R_1R_{e1}) .$$

The OC and OF for the real case are

OC: $$a_1 a_2 \le a_0 a_3 , \qquad (7)$$

OF: $$f_0' = \frac{\omega_0'}{2\pi} = \frac{1}{2\pi}\sqrt{\frac{a_0}{a_2}} \qquad (8)$$

where f_0' (ω_0') differs from f_0 (ω_0) due to real influences.

Tedious derivations lead to the oscillation condition

OC: $$\frac{R_2}{R_4} \ge \frac{R_1}{R_3} + \frac{C_2}{C_1} + \delta \qquad (9)$$

where δ is an error term

$$\delta = \beta + \gamma + \sigma , \qquad (10)$$

and

$$\beta = \frac{R_2}{R_Z}\left[\alpha + \frac{R_{e1}}{R_4} + \frac{R_1R_{e2}}{R_3R_4} + \frac{C_2}{C_1}\left(1 + \frac{R_{e2}}{R_4}\right)\right], \qquad (11)$$

$$\gamma = \frac{C_Z}{C_1}\left[\frac{R_2}{R_3} + \left(\frac{R_{e2}}{R_3} + 1\right)\left(\frac{R_2}{R_4} + 1\right) + \frac{R_{e1}}{R_3}\right], \qquad (12)$$

$$\sigma = \frac{C_Z}{C_1}\frac{\left(\frac{R_2}{R_Z} + 1 + \frac{R_2R_3}{R_ZR_4}\right)\left(\frac{R_1}{R_3} + \frac{R_{e2}}{R_3} + \frac{R_1R_{e1}}{R_3R_4}\right)}{\frac{R_1}{R_Z} + \frac{R_1}{R_2} + \frac{C_Z}{C_2}\alpha + \frac{C_Z}{C_1}}, \qquad (13)$$

$$\alpha = \frac{R_1}{R_3} + \frac{R_1}{R_4} + \frac{R_{e1}}{R_3} . \qquad (14)$$

From (9) to (14), it is possible to reveal the actual impact of parasitic effects on the OC. The β factor can be eliminated by selecting a sufficiently low R_2/R_Z ratio whereas the influence of factors γ and σ is negligible when the C_Z/C_1 ratio is small enough.

The oscillation frequency can then be expressed as

$$f_0' = \frac{f_0}{\sqrt{1+\varepsilon}} \qquad (15)$$

where

$$\varepsilon = \frac{\frac{C_Z}{C_2}\frac{R_4}{R_3}\left(1 + \frac{R_{e1}}{R_1} + \frac{R_3}{R_4} + \frac{C_2}{C_1}\frac{R_3}{R_1}\right) - \frac{R_3}{R_Z}}{\frac{R_3}{R_Z} + \frac{R_4}{R_Z} + \frac{R_4}{R_2}} . \qquad (16)$$

An error term ε, of zero value in the ideal case, is a small positive number in the real case, causing a decrease in the oscillation frequency, with R_Z and C_Z being the most dominant real effects. Their influence on the OF can be minimized by a proper choice of other component values.

4. Experimental Results

The modified WBO version in Fig. 3 was selected for experimental verification. Two OPA860 DTs were employed for circuit implementation. The following values of passive components were used: $C_1 = C_2 = 1$ nF, $R_1 = R_3 = 100\ \Omega$, $R_2 = 4.3$ kΩ, $R_4 = 1$ kΩ, $R_{ADJ} = 330\ \Omega$.

According to (5), the corresponding theoretical f_0 value is 1.592 MHz.

4.1 Amplitude Stabilization

In order to perform a relevant measurement, the manufactured WBO was equipped with an automatic amplitude stabilization circuit as shown in Fig. 6. Such a circuit was simply implemented using one photoresistor-based optocoupler, one operational amplifier, and a few resistors. The photoresistor was connected in parallel to resistor R_2, which influences the oscillation condition. The LED diode inside the optocoupler was excited by the generated signal after it had been amplified by a conventional inverting amplifier (see Fig. 6). A growing amplitude causes an increase in the LED light emission, causing a reduced resistance of the photoresistor and thus damping the amplitude. According to (4), the value of R_2 should be twice that of R_4 in the ideal case. The fixed value of R_2 was intentionally chosen more than twice as much (4.3x as mentioned above), since the oscillation condition is fulfilled by R_2 in parallel with the photoresistor.

In the ideal case, under the conditions $C_1 = C_2$ and $R_1 = R_3$, the R_2/R_4 ratio should be 2, as mentioned above. However, in reality, their ratio must be rather different. The required value can be determined from the OC, taking the real effects into consideration. According to (9), the required R_2/R_4 ratio is 2.11. In other words, the resulting resistance of the parallel connection of R_2 and photoresistor should be 2110 Ω with $R_4 = 1$ kΩ. This value exactly corresponds with a simulation in the SNAP program [21], with the CE coefficients analyzed without any simplification.

The optocoupler is a low-speed device, responding to the mean value of the signal rather than to its instantaneous values. Due to this feature, the circuitry for amplitude control can be made up in such a simple way. However, for this reason it is impossible to use the same stabilization circuit for very low frequency oscillators when the optocoupler follows the instantaneous rather than the mean values.

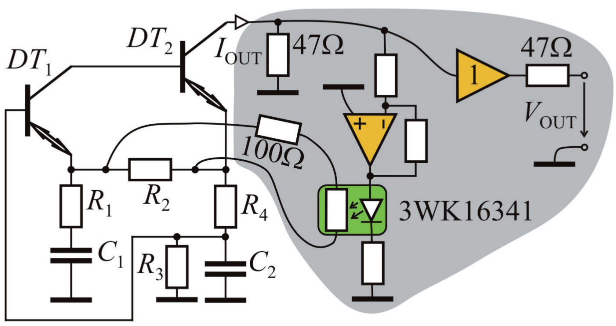

Fig. 6. Amplitude stabilization circuit and signal measuring.

4.2 Results Measured

The WBO output terminal (collector of DT_2) was enhanced by an additional external resistor of 47 Ω (Fig. 6). The current signal causes a voltage drop on this resistor, which can be subsequently used for measuring as well as for amplitude control.

The voltage signal was connected via the voltage buffer and additional 47Ω matching resistor to an oscilloscope/ spectrum analyzer.

Fig. 7 shows the waveform generated. The vertical axis is scaled such that 100 mV represents approximately 2 mA of the output current. The oscillation frequency measured was 1.433 MHz, which is about 10 % lower than the expected theoretical value 1.592 MHz. The THD measured was slightly lower than 0.25 %.

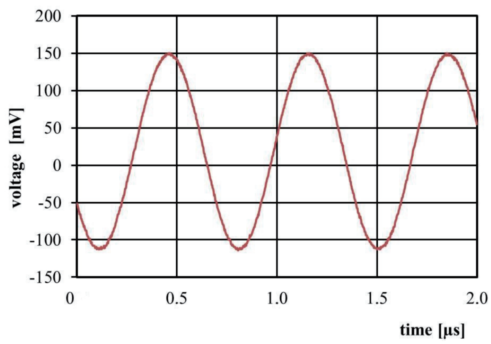

Fig. 7. The steady-state waveform of the voltage measured.

An analysis of the real influences revealed two main reasons for the difference between the measured and the theoretical oscillation frequency. It is a total parasitic impedance of the node where the collector of DT_1 is connected to the base of DT_2. According to [14], the concrete values are $C_Z = 4.1$ pF ($C_Z = C_{c1}||C_{b2}$) and R_Z only 48 kΩ ($R_Z = R_{c1}||R_{b2}$). The value of the OF computed from (15) is 1.466 MHz, which is less than a 2.5% deviation from the frequency measured. It proves that all the major real effects are properly modeled via (15).

The THD measured is on a relatively acceptable level. However, the actual THD value primarily depends on the linearity of CCIIs+, on the functionality of the amplitude stabilization circuit, and also on the R_3/R_1 and C_2/C_1 ratios as shown in [22]. Since the diamond transistor is not a linear device, the so-called degeneration resistor should be used in order to increase the linearity [14], [15]. That is why the actual THD value achievable with an on-chip WBO will be dependent on the above factors.

5. Conclusion

Two different versions of Wien-bridge type oscillators employing CCIIs+ and providing explicit current outputs were proposed in the paper. A method for overcoming the drawback of floating capacitors was shown and tested experimentally on the manufactured circuit. Positive second-generation current conveyors CCIIs+ were implemented by means of the OPA860 diamond transistors. Experimental results described in section 4 show a 10% decrease of measured oscillation frequency compared to its ideal theoretical value. A detailed analysis of real effects describes all the major influences and provides a formula for computing an approximate oscillation frequency, with a maximum error of less than 2.5 %. A relatively low collector terminal resistance of the diamond transistor as well as its parasitic capacitance were indicated as the major factors causing the error in the oscillation frequency. To eliminate this error, a CCII+ with high output resistance and low parasitic capacitances should be used.

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