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Wien oscillators using current conveyors

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Abstract

Two new Wien type oscillators using the current conveyor as the active building block are given. Both circuits have the advantage that the condition of oscillation can be adjusted by varying a single resistor without affecting the frequency of oscillation. One of the oscillator circuits has the advantage of using grounded capacitors and is obtained from a generalized configuration derived from the first oscillator circuit. PSpice simulations and experimental results are included. © 1999 Elsevier Science Ltd. All rights reserved.

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1. Introduction

A variety of sinusoidal oscillator circuits using the operational amplifier (op amp) as the active element are available in the literature [1]. It is well known that the finite gain-bandwidth product of the op amp affects both the condition and the frequency of oscillation [1]. The first oscillators using the second generation current conveyor (CCII) have been introduced in the literature since almost twenty years [2]. One of the oscillators given in Ref. [2] is based on using two opposite polarity voltage integrators and the second oscillator is based on an RC phase shift network with the CCII employed as a voltage to current converter. The oscillator circuits given in Refs. [3, 4] are based on the application of the CCII in realizing a grounded inductor and a grounded frequency dependent negative resistance, respectively. Several authors have proposed circuits for sinusoidal oscillators using a single CCII [5–9]. The oscillators

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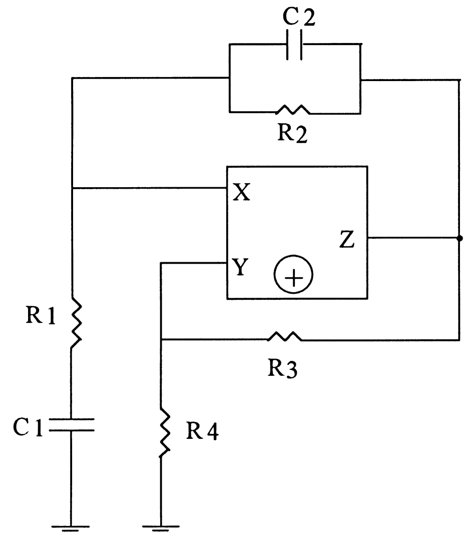

Fig. 1. A voltage mode oscillator derived from the well-known bandpass circuit [12].

reported in Refs. [5–7] employ three capacitors. The oscillator reported in Refs. [8, 9] has two capacitors, one of them however is floating.

A Wien type oscillator using the CCII has been introduced in Ref. [10] and is generated from the conventional Wien oscillator using the nullor concept. The oscillator given in Ref. [10] employs the CCII as a negative impedance converter (NIC). Recently Wien type oscillators using the CCII and based on replacing the voltage controlled voltage source (VCVS) in the classical Wien oscillator by a transconductance circuit were given in Ref. [11], none of these oscillators employs grounded capacitors.

In this paper two new oscillator circuits using the current conveyor are proposed, one of them employs grounded capacitors. The first oscillator is generated from the bandpass filter circuit given in Ref. [12] and the second oscillator is obtained from a generalized configuration derived from the first oscillator circuit. PSpice simulations and experimental measurements indicating the performance of both oscillators are given.

2. The proposed oscillators

Fig. 1 represents the oscillator circuit which is obtained from one of the bandpass filters given in Ref. [12] after setting the input port short circuit. The characteristic equation of this oscillator circuit is given by:

$$s^2C_1C_2R_1R_2 + s\left[C_1R_1 + C_2R_2 + \frac{C_1R_2}{2R_3}(R_1 - R_4)\right] + 1 + \frac{R_2}{2R_3} = 0 \quad (1)$$

From Eq. (1), the condition of oscillation and the radian frequency of oscillation are given, respectively, by:

$$R_4 = R_1 + 2R_3\left[\frac{R_1}{R_2} + \frac{C_2}{C_1}\right] \tag{2}$$

and

$$\omega_0 = \sqrt{\frac{1 + (R_2/2R_3)}{C_1 C_2 R_1 R_2}} \tag{3}$$

Thus it is seen that the grounded resistor R_4 controls the condition of oscillation without affecting ω_0. One set of design equations is obtained by taking:

$$C_1 = C_2 = C \quad \text{and} \quad R_1 = R_2 = R_3 = R \tag{4}$$

Thus

$$R_4 = 5R \quad \text{and} \quad \omega_0 = \sqrt{\frac{3}{2}}\frac{1}{CR} \tag{5}$$

The frequency of oscillation can be adjusted by varying the two equal capacitors without affecting the condition of oscillation.

PSpice simulations for the oscillator circuit of Fig. 1 have been performed using the AD844A/AD biased with ± 9 V supplies and taking $R_1 = R_2 = R_3 = 10$ kΩ, $C_1 = C_2 = 0.1$ nF and R_4 was taken 53 kΩ (to start oscillations). Fig. 2(a) represents the oscillator waveform obtained, from which it is seen that the actual oscillation frequency foa = 180 kHz, thus $\Delta f_0/f_0 = -7.656\%$.

Fig. 2(b) represents V_{C2} versus V_{C1} showing a simple limit cycle of period one, indicating a single harmonic frequency. The results of the total harmonic distortion analysis are summarized in Table 1.

It is clear that this oscillator circuit is more generally represented as shown in Fig. 3(a). For this generalized oscillator the characteristic equation is given by:

$$\frac{Z_2}{Z_1} = 1 + \frac{2Z_3}{Z_4} \tag{6}$$

In order to generate an oscillator circuit with grounded capacitors from the generalized configuration shown in Fig. 3(a), Z_3 and Z_4 must be taken as resistors and in this case, the two circuits shown in Fig. 3(b) and (c) are obtained. For the circuit of Fig. 3(b), the characteristic equation is given by:

$$s^2 C_1 C_2 R_1 R_2 + s\left[C_1 R_1 + C_2 R_2 - \frac{C_1 R_2}{1 + (2R_3)/(R_4)}\right] + 1 = 0 \tag{7}$$

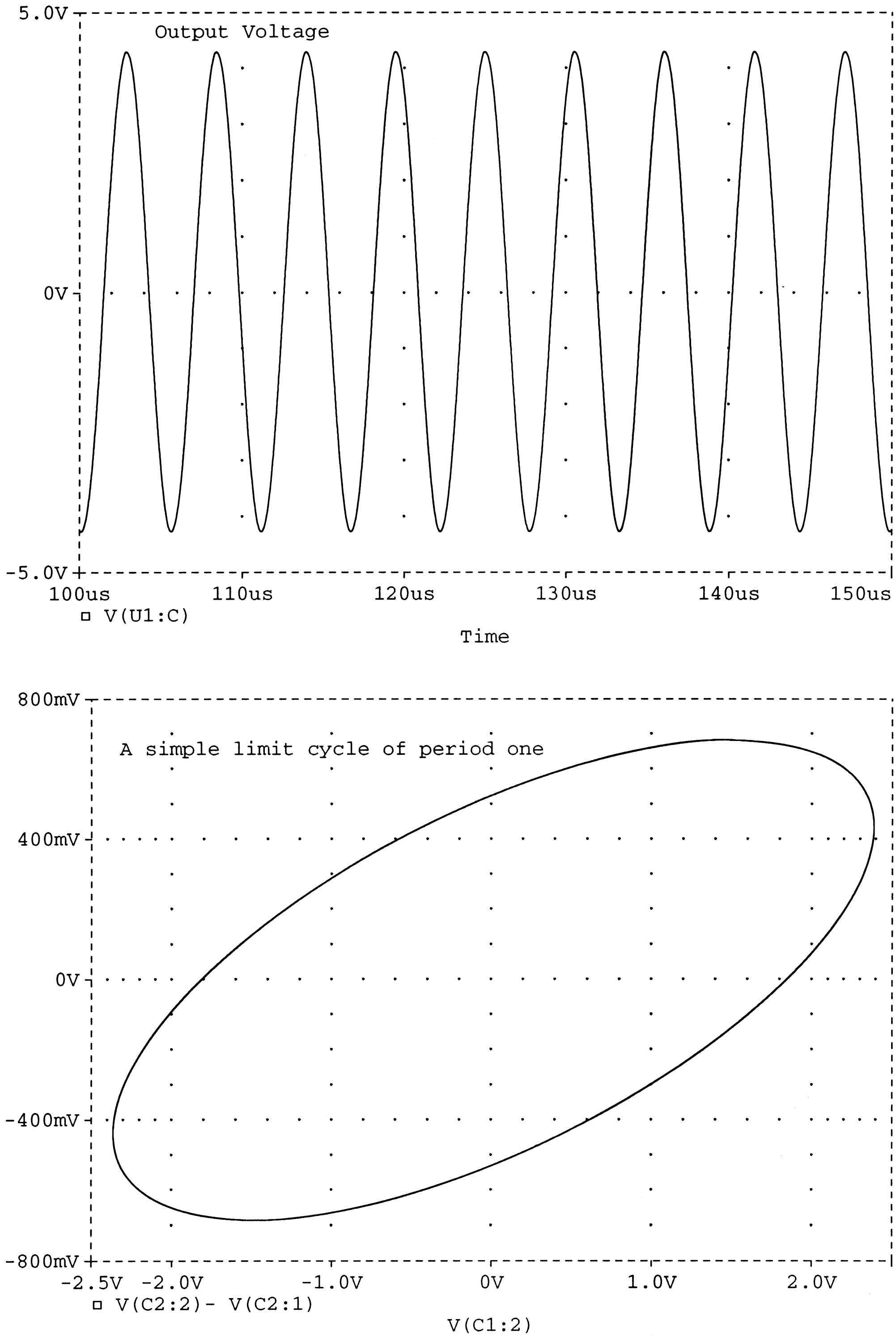

Fig. 2. (a) The output voltage waveform for the oscillator circuit of Fig. 1. (b) V_{C2} versus V_{C1} for the oscillator circuit of Fig. 1.

Table 1
Circuit of Fig. 1

Harmonic number	Frequency (Hz)	Fourier component	Normalized component	Phase (DEG)	Normalized phase (DEG)
1	1.800E + 05	4.283E + 00	1.000E + 00	4.513E + 01	0.00E + 00
2	3.600E + 05	1.305E−02	3.048E−03	−1.446E + 02	−1.897E + 02
3	5.400E + 05	5.471E−03	1.277E−03	1.465E + 02	1.013E + 02
Dc component	1.996E−02				
Total harmonic distortion	3.837E−01% (−48.32 dB)				

For this oscillator circuit, the recommended design is to take:

$$C_1 R_1 = C_2 R_2 \tag{8}$$

In this case the condition of oscillation is given by:

$$\frac{2R_3}{R_4} = \frac{R_2}{2R_1} - 1 \tag{9}$$

and the radian frequency of oscillation is given by:

$$\omega_0 = \frac{1}{C_1 R_1} \tag{10}$$

Fig. 4(a) represents the oscillator waveform obtained from the oscillator circuit shown in Fig. 3(b) with $R_1 = R_3 = 10$ kΩ, $R_2 = 40$ kΩ, $C_1 = 0.4$ nF, $C_2 = 0.1$ nF and R_4 was taken as 21 kΩ (to start oscillations). The actual oscillation frequency was measured at 39.08 kHz, therefore $\Delta f_0/f_0 = -1.782\%$.

Fig. 4(b) represents V_{C2} versus V_{C1} showing a simple limit cycle. Harmonic distortion analysis results are summarized in Table 2.

It is worth noting that in the special case of setting R_3 as a short circuit and R_4 as an open circuit, the CCII operates as a current negative impedance converter and the oscillator simplifies to the minimum component oscillator given in Refs. [10, 11, 13]. In this case however the condition of oscillation and the frequency of oscillation are both dependent on the four passive circuit components. It should also be noted that the PSpice simulations of the circuit of Fig. 3(c) indicates a latch-up mode of operation.

3. Amplitude control and stability of the limit cycle

It is a common practice to describe the operation of sinusoidal oscillators by means of linear models. This is usually justified in view of the very low distortion levels observed in practical circuits (less than 1% THD). However, it is well known that an entirely linear system cannot maintain stable oscillations. Any practical oscillator must contain a nonlinear

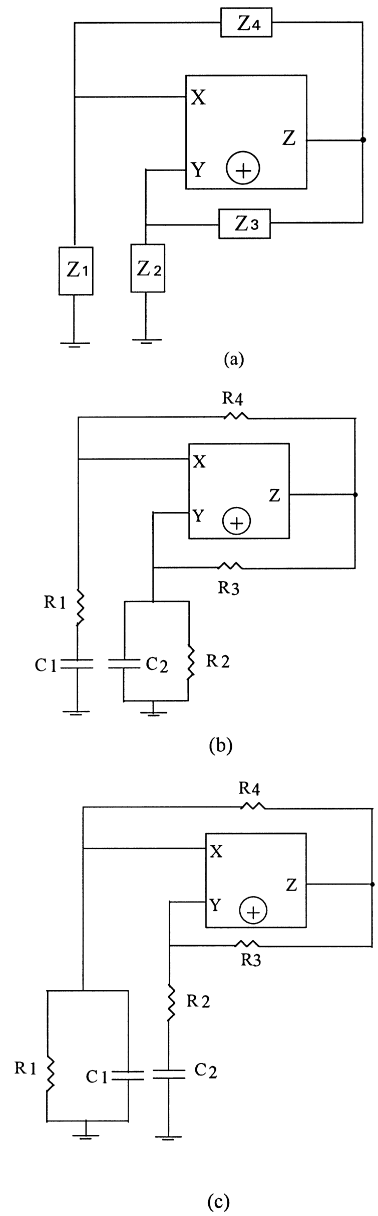

Fig. 3. (a) A generalized voltage mode oscillator using a single CCII based on the circuit of Fig. 1. (b) The new grounded capacitor voltage mode oscillator. (c) The unstable grounded capacitor circuit.

device or mechanism to control amplitude. For precision oscillators, external circuitry containing Zener diodes and a voltage controlled resistor is usually added to control the amplitude. However, most general purpose oscillators rely on the output voltage saturation of the active device as a nonlinear mechanism for amplitude control. Such oscillators are termed self-limiting oscillators [14]. This mechanism (also known as soft limiting) is a simple

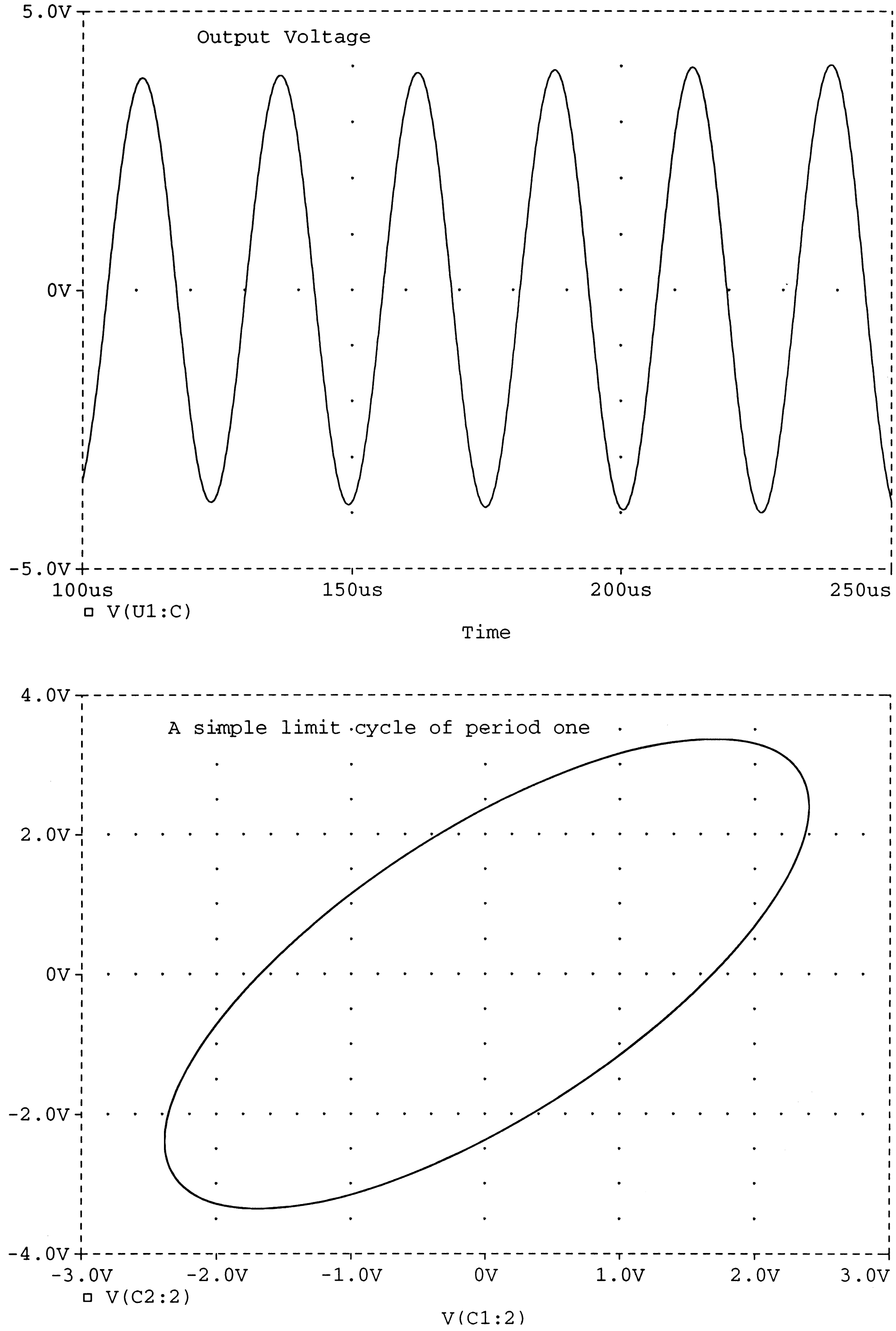

Fig. 4. (a) The output voltage waveform for the oscillator circuit of Fig. 3(b). (b) V_{C2} versus V_{C1} for the oscillator circuit of Fig. 3(b).

Table 2
Circuit of Fig. 3(b)

Harmonic number	Frequency (Hz)	Fourier component	Normalized component	Phase (DEG)	Normalized phase (DEG)
1	3.908E + 04	4.278E + 00	1.000E + 00	1.721E + 02	0.00E + 00
2	7.817E + 04	1.754E − 02	4.101E − 03	−4.910E + 00	−1.770E + 02
3	1.172E + 05	1.175E − 02	2.746E − 03	2.880E + 01	−1.433E + 02
Dc component	6.005E − 03				
Total harmonic distortion	5.4296E − 01% (−45.31 dB)				

nonlinear gain with a montonically increasing or decreasing describing function that depends only on the amplitude and not on the frequency. A typical describing function for output voltage saturation is given by:

$$N(a) = 1 \quad a \le \alpha, \qquad N(a) = \frac{2}{\pi}\sqrt{\sin^{-1}\left(\frac{\alpha}{a}\right) + \frac{\alpha}{a}\left[1 - \left(\frac{\alpha}{a}\right)^2\right]} \quad a > \alpha \tag{11}$$

where a is the amplitude and α is the saturation amplitude.

A general second order characteristic equation can then be written in the form:

$$(b_2 + a_2 N)s^2 + (b_1 + a_1 N)s + (b_0 + a_0 N) = 0 \tag{12}$$

where N is as given by Eq. (11).

In order to investigate the stability of the limit cycle, the variable s and its real and imaginary components are considered to be functions of the amplitude of oscillation [15]. Hence,

$$s(a) = \sigma(a) + j\omega(a) \tag{13}$$

Substituting in Eq. (12) results in:

$$B_1 + C_1 N = 0 \quad \text{and} \quad B_2 + C_2 N = 0 \tag{14}$$

where

$$\begin{aligned} &B_1 = b_2(\sigma^2 - \omega^2) + b_1\sigma + b_0 \qquad B_2 = 2b_2\sigma\omega + b_1\omega \qquad C_1 = a_2(\sigma^2 - \omega^2) + a_1\sigma + a_0 \\ &C_2 = 2a_2\sigma\omega + a_1\omega \end{aligned} \tag{15}$$

The partial derivatives of ω and σ with respect to the amplitude can be evaluated and simplified using the Cauchy–Riemann conditions. The resulting expression for the derivative

$\partial\sigma/\partial a$ for small perturbations of the roots away from the imaginary axis ($\sigma = 0$) is found to be

$$\frac{\partial\sigma}{\partial a} = \frac{-\omega^2(b_2 + a_2N)a_1 + (b_1 + a_1N)(-a_2\omega^2 + a_0)}{2(b_1 + a_1N)^2 + 2\omega^2(b_2 + a_2N)^2}\frac{\mathrm{d}N}{\mathrm{d}a} \tag{16}$$

which is a special case of equation (8) in Ref. [15].

The derivative $\partial\sigma/\partial a$ contains significant information about the behavior of the oscillator when perturbed a small distance away from its limit cycle. The stability of a limit cycle can be determined from the sign of $\partial\sigma/\partial a$ which is negative for a stable limit cycle.

For a saturation type nonlinearity $\mathrm{d}N/\mathrm{d}a$ is negative, hence, stability requires that

$$-\omega^2(b_2 + a_2N)a_1 + (b_1 + a_1N)(-a_2\omega^2 + a_0) > 0 \tag{17}$$

A choice of a_2, a_1 and a_0 that well fits most practical oscillators is to take a_2 positive, a_1 negative and a_0 equal to zero.

In this case, and with b_2 positive, it can be shown that the limit cycle is stable if b_1 is negative.

For the oscillator circuits of Fig. 1 and Fig. 3(b) it is necessary to move the imaginary poles slightly into the right half plane in order to start oscillations. This is achieved by increasing the value of one resistor from the design value required to set $b_1 = 0$. With the increased resistor value, the sign of b_1 is found to be negative for both oscillators, hence the limit cycles are stable.

Table 3
The experimental results for the two new circuits of Fig. 1 and Fig. 3(b)

Oscillator circuit	Circuit parameters				Oscillation frequency		
	C	R (kΩ)	R_4 (start) exp. (kΩ)	R_4 (start) simul. (kΩ)	theor. (kHz)	exp. (kHz)	simul. (kHz)
Fig. 1	0.1 μF	10	51.5	51	0.1949	0.196	0.1959
	0.1 μF	1	6	5.7	1.949	1.667	1.866
	1 nF	10	57	51	19.492	19.231	18.992
	4.7 nF	1	6	5.7	41.473	40.883	40.409
	1 nF	1	6.2	5.7	194.924	156.25	184.159
Fig. 3(b)	0.1 μF	10	23.3	21	0.0398	0.040	0.0399
	4.7 nF	10	27.5	21	0.8466	0.900	0.855
	4.7 nF	1	2.6	2.4	8.466	9.091	8.532
	1 nF	1	2.6	2.4	39.789	41.667	39.998
	130 pF	1	4.3	2.4	306.067	243.902	301.196

4. Experimental results

The circuits of Fig. 1 and Fig. 3(b) have been tested experimentally using the Analog Devices AD844 biased with ± 9 V supplies.

The design equations for the circuit of Fig. 1 were taken as given by Eq. (4). The circuit of Fig. 3(b) was designed with $C_2 = C$, $C_1 = 4C$, $R_1 = R_3 = R$ and $R_2 = 4R$. In both cases, oscillations are started by tuning the resistor R_4.

Table 3 includes the experimental data obtained for both circuits together with the theoretical and the PSpice simulated results. It is seen that the deviation in the frequency of oscillation increases at high frequencies, this is due to the parasitic impedances of the AD844.

5. Conclusions

An alternative approach different from those suggested in Refs. [11, 13] for realizing Wien type oscillators using the CCII is given. The proposed oscillators have the advantage that the condition of oscillation can be adjusted by varying a single resistor without affecting the frequency of oscillation. The first proposed oscillator is based on the bandpass circuit given in Ref. [12] and it includes the oscillator VII of Ref. [13] as a special case. The grounded capacitor oscillator includes the oscillator III of Ref. [13] which was also given in Ref. [10] as a special case, with the advantage of having independent control on the condition of oscillation. This is an important advantage since to start oscillations in the circuit of Fig. 3(b) it is always necessary to increase R_4 (or decrease R_3), this will not affect the frequency of oscillation. This property cannot be achieved in the NIC oscillator given in Refs. [10, 13]. PSpice simulations and experimental results indicating the oscillators performance are included.

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