A Two-Section Phase-Shift Oscillator

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***Abstract*— This paper proposes an alternative phase-shift oscillator that only needs two RC sections. In order to extend the frequency range of operation, a composite amplifier structure is used instead of a single operational amplifier block. This solution has some advantages on the values of the circuit elements and in the frequency range according to the gain-bandwidth of the amplifier. Preliminary experimental results are shown and discussed, which demonstrate the functionalities expected. Ongoing research work is being carried on to improve the oscillator performance.**

***Keywords*—Analog design, Bandwidth extension, Canonical oscillators, Composite amplifiers, Phase-shift oscillators**

I. Introduction

Sinusoidal oscillators are a classical subject and one of the most common applications of operational amplifiers (OA) based active circuits, whose fundamentals are based on feedback amplifier theory and RC frequency-selective network response. The general structure of sinusoidal oscillators is depicted in fig. 1, where $A(s)$ represents the open loop amplifier gain of the amplifier and $\beta(s)$ represents the feedback coefficient imposed by the RC network, which are both frequency dependant. Signal X_s in fig. 1 represents the initial perturbation of the system, e.g. noise when switching on the power supplies. A negative feedback amplifier block is normally used, to take the advantages of negative feedback properties, which most often is based on operational amplifiers [1],[2] .

By definition, the gain of the feedback amplifier is given by

$$A_f(s) \equiv \frac{X_o}{X_s} = \frac{A(s)}{1 - A(s)\cdot\beta(s)} \qquad (1)$$

At a given frequency, ω_0, the loop gain, $L(s) \equiv A(s)\cdot\beta(s)$, equals unity with null phase shift, i.e. $L(j\omega_0) = 1\angle 0^o$ (the Barkausen criterion), the poles are located on the imaginary axis and oscillations start with frequency ω_0.

To verify the oscillation criterion, the RC network has to impose an overall phase shift of 180º, when an inverting amplifier configuration is used, or a 0º overall phase shift when using a non-inverting amplifier.

Phase-shift oscillators are among the classical and most popular RC-oscillator topologies, whose configurations are normally based on inverting amplifiers and a three section RC network. There is a common idea that the minimum number

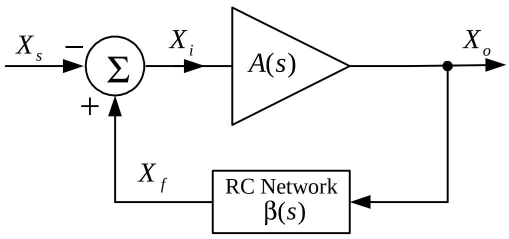

Fig. 1. General structure of a sinusoidal oscillator: a feedback amplifier and a frequency-selective RC network.

of RC sections is three. This is probably due to the most common implementations of the phase-shift oscillator presented in bibliography [1],[2], or the classical topologies [3]-[5] or even due to one of the most authoritative papers on classical theory for this subject [6]. If one uses an inverting amplifier than it requires a minimum of three RC sections. However, if a non-inverting amplifier is used, only two sections are needed in the RC network, given that the overall phase shift imposed is null. This leads to the so called canonical RC oscillators [7], which only require two capacitors and four resistors: two for the RC network and two for the amplifier gain.

On the other hand, for practical reasons, it is often useful to extend the bandwidth in the most common applications of linear active circuits. The frequency-dependant gain behavior of active elements lead some research for improvements on frequency operation in RC type oscillators [8] and several approaches were proposed. The approach to generate composite operational amplifiers proposed by Mikhael and Michael [9] allows for effective improvements on bandwidth in similar applications [10],[11] with interesting results.

In this paper, we discuss and present an alternative phase-shift oscillator that only uses two RC sections and a non-inverting amplifier. In order to extend the frequency range we use a composite amplifier scheme instead of a single amplifier stage. Preliminary results of an ongoing research concerning the design and implementation of a two-section RC network phase-shift oscillator are presented and discussed, thus demonstrating the expected features of the proposed oscillator.

The paper is organized as follows. Section II presents the basics of typical phase-shift oscillators and the fundamentals of the composite amplifier configuration in use. The proposed two-section phase shift oscillator is presented in section III, for a single amplifier implementation, and using composite amplifiers for bandwidth extension. In section IV some evaluation results are presented and discussed, and finally conclusions are made in section V.

II. PHASE-SHIFT OSCILLATOR AND COMPOSITE AMPLIFIER BASICS AND STABILITY

A. Phase-shift Oscillators

The most common configuration of the phase-shift oscillator is presented in fig. 2. It uses an inverting amplifier associated with a three-section RC network. The gain of the inverting amplifier is given by $K = -R_f / R_1$ and each section of the frequency-selective network is equal in terms of configuration and values of the circuit elements. At the RC network in fig. 2 Z_a and Z_b can either be capacitors or resistors. If Z_a are capacitors and Z_b resistors, and assuming independency among sections, the loop gain is given by

$$L(s) = \frac{-\left(R_f / R_1\right) s^3}{s^3 + \frac{6}{RC} s^2 + \frac{5}{(RC)^2} s + \frac{1}{(RC)^3}} \quad (2)$$

Through (2) one obtains $\omega_0 = \frac{1}{\sqrt{6} \cdot RC}$ and $\frac{R_f}{R_1} = 29$.

For its counterpart, i.e. Z_a are resistors and Z_b capacitors, the transfer function is given by

$$L(s) = \frac{-\left(R_f / R_1\right)}{\left(sCR\right)^3 + 5\left(sCR\right)^2 + 6sCR + 1} \quad (3)$$

In this case the frequency is $\omega_0 = \frac{\sqrt{6}}{RC}$ and $\frac{R_f}{R_1} = 29$.

In both cases, the gain condition remains equal and the RC network imposes an overall phase shift of 180º at the natural frequency, i.e., each section imposes a 60º phase shift, either in advance or in delay, depending on the configuration. The finite gain-bandwidth product (GBW) of the OAs, and the higher value of the amplifier gain represent some limitations on the frequency range of the oscillator. Moreover, care should be taken at design and on choosing the values of the circuit elements, namely for R_1, as low values on the amplifier input impedance can significantly deviate the frequency.

B. Composite Amplifiers

Based on the ideal operational amplifier model and a nullor chain transmission matrix model, Mikhael and Michael [9] proposed several structures of composite amplifiers called CNOA's, where N stands for the number of operational amplifiers used in each configuration. The simplest composite amplifier configurations use only two operational amplifiers – C2OA's – and only four meet the required performance criteria, so only these were proposed [9]: C2OA-1, C2OA-2, C2OA-3 and C2OA-4. Within these, the C2OA2 configuration presented in fig. 3 has special interest for application on the two-section phase-shift oscillator due to its simplicity, characteristics and features.

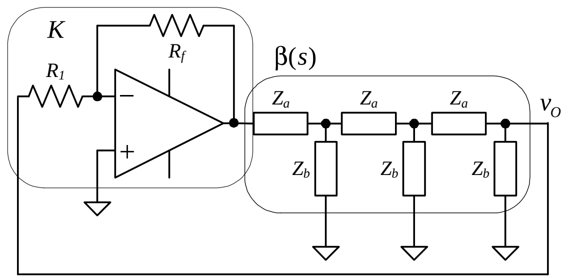

Fig. 2. Typical configuration of a phase-shift oscillator based on an inverting amplifier and a three section RC network.

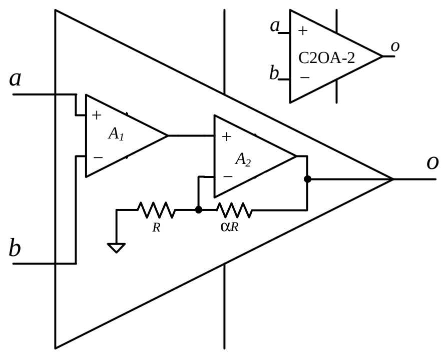

Fig. 3. Structure of the C2OA-2 composite amplifier configuration.

The open-loop gain of the single OA used in modeling the C2OA´s configuration (assuming a single-pole model) is

$$A_i = \frac{A_{oi}\omega_{Li}}{\omega_{Li} + s} = \frac{\omega_i}{\omega_{Li} + s} \qquad i = 1, 2 \quad (4)$$

where A_{oi}, ω_{Li} and ω_i are the dc open-loop gain, the 3-dB bandwidth and the GBW of the *i*th single OA, respectively. For C2OA-2 the open loop input-output relationship is given by

$$V_o = V_a \frac{A_1 A_2 (1+\alpha)}{A_2 + (1+\alpha)} - V_b \frac{A_1 A_2 (1+\alpha)}{A_2 + (1+\alpha)} \Leftrightarrow$$
$$V_o = (V_a - V_b) \frac{A_1 A_2 (1+\alpha)}{A_2 + (1+\alpha)} \quad (5)$$

where α is a resistor ratio.

This composite amplifier has a single-pole roll off from ω_i / A_o to $\omega_i / (1+\alpha)$ where the second pole occurs. Moreover, it is interesting to note that, among the four above mentioned configurations, only C2OA-2 has identical expressions for the positive and negative open-loop gain as can be observed in (5). Thus, common-mode rejection ratio (CMRR) problems should not occur, even for relatively large common-mode signals. The C2OA-2 is then suited for application on the proposed phase-shift oscillator as it fits the non inverting characteristics.

C. Stability of the C2OA Composite Amplifiers

The stability conditions of the C2OA composite amplifiers are studied in [9] and [12], using the two-pole model for each amplifier and the Routh-Hurwitz stability criterion. The later considers also the input impedance of the non-inverting amplifiers. Noting that the input impedance, Z_{in}, is only due to the finite GBW of the amplifier used, one assumes that the open-loop gain of a single OA is given by

$$A(s) = \mathrm{GBW}/s \qquad (6)$$

For practical reasons, one also assumes that the two OAs are identical. Hence, neglecting the second-order effects, it can be shown that for the C2OA-1 and C2OA-2 cases:

$$Z_{in} = R + \frac{1}{j\omega C_{eq}} \qquad \text{where} \quad C_{eq} = \frac{K}{\mathrm{GBW}(1+\alpha)} \qquad (7)$$

K is the overall gain of the amplifier. Thus, using the two-pole model, the open loop gain of a single OA is give by

$$A(s) = \frac{A_0}{(1+s/\omega_l)(1+s/\omega_h)} \qquad (8)$$

Assuming that ω_l is much smaller than any frequency of interest and applying the Routh-Hurwitz criterion, one obtains the stability condition for the C2OA-1 and C2OA-2 given by

$$K > \frac{2(1+\alpha)^2}{\left[0.5+(1+\alpha)E\right]} \qquad (9)$$

where α is the resistor ratio and $E = \omega_h/\mathrm{GBW}$.

III. The Two-Section Phase-Shift Oscillator

The proposed phase-shift oscillator is based on a non-inverting amplifier associated with a two-section RC network. Since the non-inverting amplifier does not affect the phase within the limitations of the GBW product, the same is required from the RC network, in order to verify the Barkausen criterion. Therefore, in terms of phase, each section has to behave in a complementary way, in order to have a null overall phase-shift.

The practical single amplifier configuration of this oscillator is presented in fig. 4, considering one of the two possible topologies for the RC network; the other one is its counterpart. The gain of the amplifier is defined by $K = 1 + R_f/R_1$ and is adjustable for practical reasons. For either configuration of the RC network, assuming equal values for the resistors and capacitors on each section, the transfer function of the frequency-selective network is given by

$$\beta(s) = \frac{sCR}{(sCR)^2 + 3RCs + 1} \qquad (10)$$

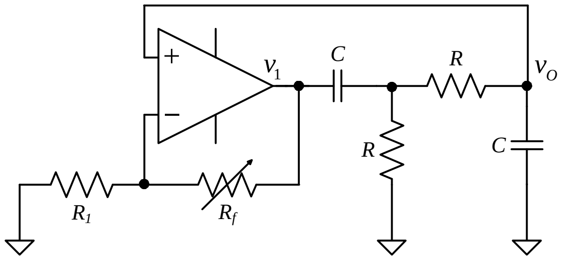

Fig. 4. Electrical scheme of the proposed two-section phase-shift oscillator using a single amplifier stage.

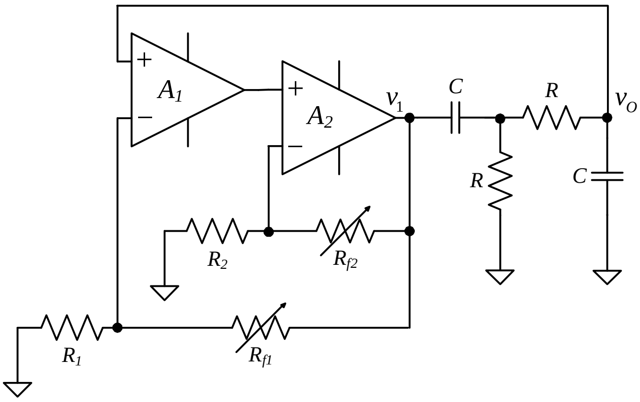

Fig. 5. The two-section phase-shift oscillator using a C2OA-2 non-inverting amplifier stage.

Therefore, the closed loop gain is equal for both topologies

$$L(j\omega) = K\cdot\beta(j\omega) = \frac{jK\omega CR}{\left[1-(\omega RC)^2\right] + j3\omega RC} \qquad (11)$$

The oscillation frequency is $\omega_0 = 1/RC$ and $K = 3$. The amplifier gain must be $K \approx 3$ to have sustainable oscillations and slightly greater of that value safely to start the oscillations, i.e., $R_f = 2R_1 + \delta$. Without any automatic gain control (AGC) scheme, the stability is controlled by the non-linearities of the circuit, similarly to other sinusoidal RC oscillator configurations.

One sees that the two-section phase-shift oscillator is simpler than other phase-shift oscillator configurations. On the other hand, when compared to the typical phase-shift configurations, one verifies that the two-section oscillator requires an amplifier gain that is about ten times lower. This feature has advantages on the frequency range of operation, at the noise sensitivity and at the stability, especially at higher frequencies. Moreover, the use of a non-inverting amplifier has advantages on the input impedance of the amplifier, thus overcoming the drawbacks of the inverting amplifier configuration (e.g. the input impedance) while maintaining all the benefits of negative feedback structure, namely gain sensitivity and gain-bandwidth.

The frequency range can also be improved by replacing the single amplifier with a composite amplifier stage. The electrical scheme for the C2OA-2 composite amplifier is presented in figure 5 and includes variable resistors to adjust the gain in each amplifier stage, for practical reasons. To have sustainable oscillations the gain in each stage has to be set to $\sim\sqrt{3}$ thus, an average gain of ~3 on the frequency range is expected when using composite amplifiers stage.

IV. Evaluation and Experimental Results

To evaluate the proposed phase-shift oscillator, we implemented the configurations depicted on fig. 4 and 5 using the TL084 OA (slew rate of 16 V/µs and GBW ~4 MHz) for the two possible RC networks and for different values of resistors and capacitors. Regarding the GBW product and the frequency dependant parameters of the OA, a non-inverting amplifier with a gain of 3 (K=3) using the TL084 allows a frequency range up to ~1 MHz without distortion, depending on the amplitude of the generated sine wave.

The evaluation of the oscillator considers mainly the frequency of oscillation, the relative frequency deviation, signal geometry (duty-cycle) and the total harmonic distortion computed up to the seventh harmonic. One also evaluates the gain on frequency range achieved with the composite amplifier implementation. The measurement equipment list includes an Agilent 54615B digital oscilloscope with an Agilent 54659A data acquisition and PC link module, and an Advantest R3131A Spectrum Analyzer.

TABLE I
Results for the Two-RC sections phase-shift oscillators: the single amplifier and the C2OA-2 composite amplifier configurations

		R=150 Ω	**R=220 Ω**	**R=390 Ω**	**R=560 Ω**
Single Amp.	f_o^* [Hz]	$1.06\text{x}10^6$	$723\text{x}10^3$	$408\text{x}10^3$	$284\text{x}10^3$
	f_{o1} [Hz]	$420\text{x}10^3$	$342\text{x}10^3$	$240.4\text{x}10^3$	$198.4\text{x}10^3$
	$\Delta f / f_o$ [%]	60.3	52.7	41	30.1
	Sym [%]	51	50	49.5	50
	THD [%]	0.12	0.28	0.39	0.6
	K	3.5	3.4	3.2	3.1
Composite Amp.	f_{o2} [Hz]	N/A	N/A	$581\text{x}10^3$	$813\text{x}10^3$
	G [f_{o2}/f_{o1}]			2.4	4.09
	Sym [%]			49	49.2
	THD [%]			0.17	0.45
	K (apx)			3.27	4

* Estimated frequency of oscillation according to (11)

Table I shows some evaluation results for the two-section phase-shift oscillator, for the single amplifier configuration and the C2OA-2 composite amplifier configuration, both without AGC, with $C = 1\,\text{nF}$ and considering various resistor values. All the values presented refer to v_O (see fig. 4 and 5), measured after a buffer amplifier interface according to the input impedance requirements of the spectrum analyzer.

For 1nF ceramic capacitors and E12 series resistors ranging from 150 to 560 Ω, the expected frequencies ranges from 1.06 MHz down to 284 kHz. In real circuit implementations with the TL084 and for the single amplifier configuration, the frequency values range from ~420 kHz down to ~199 kHz with signal amplitudes from ~700 mV to ~3 V. The real frequency values shown in table I consider the start of the oscillations condition. These values differ more significantly from the expected frequencies as the resistor values decreases, i.e., as the value of the theoretical frequency increases, which is also noticeable with the frequency deviation relative to the expected value, and for the symmetry of the waveform at the output signal, measured through the duty-cycle. The relative deviation on symmetry is lower than 2%. The harmonic distortion, in general, is better than 0.6% and tends to decrease as the frequency increases. The value of the amplifier gain, K, also taken for the start of oscillations condition, is almost equal to 3 but increases slightly as the frequency of operation increases. This is related to the compensation of the gain imposed by the RC network, and also due to the real influence of the RC sections.

The evaluation of the C2OA-2 composite amplifier with respect to the single amplifier considers only two conditions that allow an effective extension on bandwidth, both with C=1nF: R=390 and R=560 Ω. The other two conditions suffer from gain-bandwidth limitations of the amplifiers and the real frequency is nearly the same as for the single amplifier configuration thus, no significant improvements are obtained. Moreover, from (9) one verifies that for a given resistor ratio stability problems may occur. For the sake of stability precision resistor are preferred to define the resistor ratio and the gain. One notices that the harmonic distortion for the composite amplifier realization has the same behavior as for the single amplifier implementation and although not so significant, it shows better results for the two conditions evaluated.

To evaluate the possible gain on bandwidth, symmetry and harmonic distortion we compare the both implementations without automatic gain control for the lower frequency condition ($C = 1\,\text{nF}$ and $R = 560\,\Omega$), whose results are shown in fig. 6 to 9. There is a gain of 3.2 to 4 on frequency range, thus confirming the expected expansion on bandwidth. At the spectrum analyzer figures, one can see that the difference on level between the fundamental frequency and the second harmonic in both cases. For the single amplifier the difference is ~28 dB and is ~23 dB for the composite amplifier. As the frequency of oscillation varies from ~198 kHz to ~813 kHz, from the single to composite amplifier, respectively (C=1nF and R=560Ω), at the same conditions the amplitude of the output signals varies from ~2.8 V down to ~350 mV, which means an attenuation of ~18 dB. Hence, signal distortion is more evident in the later case. The stability problem is also more noticeable for the composite amplifier. Even though, as the symmetry of the signal is not seriously affected, it is possible to meet a better quality by using an adequate low pass filtering process.

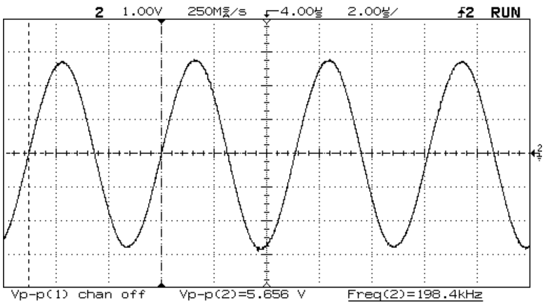

Fig. 6. Waveform of the output signal for the single amplifier configuration with C=1nF and R=560 Ω.

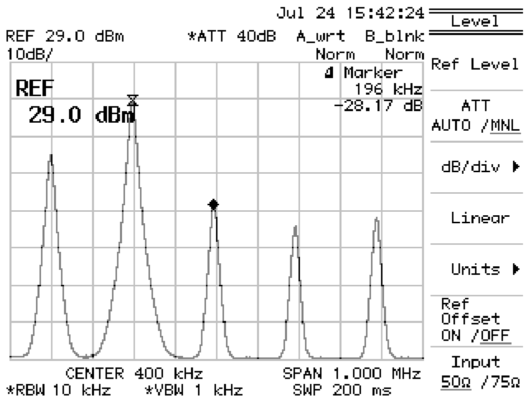

Fig. 7. Frequency spectrum for the single amplifier realization with C=1nF and R=560 Ω.

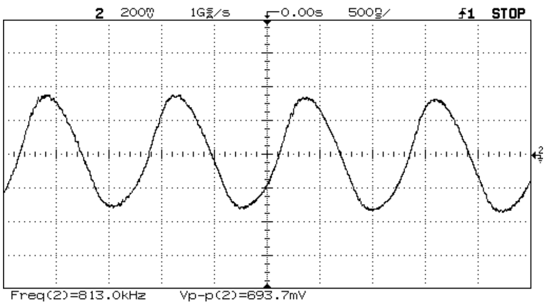

Fig. 8. Output signal for the C2OA-2 composite amplifier with C=1nF and R=560 Ω.

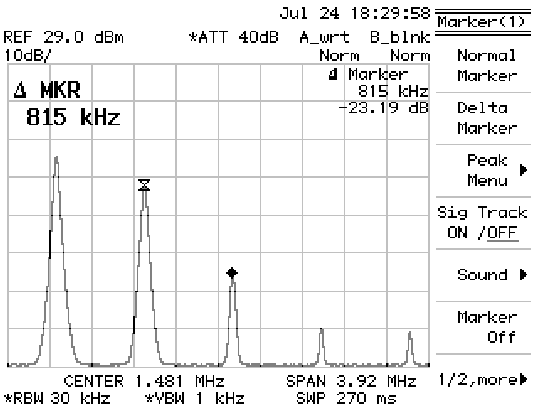

Fig. 9. Frequency spectrum for the C20A-2 composite amplifier with C=1nF and R=560 Ω.

To have a better idea on the two-section oscillator with respect to the typical implementations, we evaluated a three-section phase-shift oscillator for similar conditions, and compared it with the two-section oscillator. Although the values of the circuit elements are different from those of the two-section oscillator the frequencies of oscillation are similar. The resistors in this implantation are also from E12 series.

Table II shows the evaluation results for a three-section phase-shift oscillator with low-pass type RC network.

TABLE II
RESULTS FOR THE TYPICAL PHASE-SHIFT OSCILLATOR WITH A THREE-RC SECTION FREQUENCY SELECTIVE NETWORK (LOW-PASS CONFIGURATION)

	R=390 Ω	**R=560 Ω**	**R=1 kΩ**	**R=1.5 kΩ**
f_o^* [Hz]	$1.0x10^6$	$696.2x10^3$	$389.8x10^3$	$259.9x10^3$
f_o [Hz]	$242.7x10^3$	$250x10^3$	$195.3x10^3$	$149.7x10^3$
$\Delta f / f_o$ [%]	75.7	64	49.9	42.4
Sym [%]	50.1	50	50	50
THD [%]	0.07	0.01	0.005	0.001

* Estimated frequency of oscillation according to (3)

Comparing the results from table II with respect to table I one notices that for the three-section oscillator: the real frequency of oscillation obtained is lower; the relative deviation in frequency is higher; the symmetry is in general better and that the harmonic distortion is significantly better.

V. CONCLUSIONS

A two-section phase shift oscillator has been proposed and discussed in this paper. The oscillator is canonical with respect to the number of passive components. The two-section RC network is only possible due to the use of non-inverting amplifier stage, either using a single amplifier or a composite amplifier arrangement for bandwidth extension.

Preliminary results of ongoing research are presented and demonstrate the features of the proposed oscillator for simple implementations without AGC. The frequency range has been extended with the use of composite amplifier stage. The gain achieved on frequency bandwidth demonstrates the expected theoretical values. Harmonic distortion and stability are some drawbacks that deserve further attention, as well as other performance improvements. Work is in progress compare the performance of the proposed oscillator with other configurations, and also to evaluate the advantages of other possible composite amplifier implementations. The proposed oscillator is suited for analog IC design implementations, due to its simplicity and type of circuit elements (passive and active).

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