About theoretical and practical aspects of current mode RC oscillators design

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Abstract — This paper addresses a group of constructive elements with which, (through adequate combinations) one can generate current mode RC oscillator transfer functions. Obviously this elements set is not unique. From the multitude of possible solutions only the solutions that accommodate the below conditions stand out:

- the active elements can be easily produced in monolithic technology,
- each oscillator must have two resistors or two capacitors connected to the mass.

The latter requirement is very important for the oscillators with variable frequency. It was made RC oscillators and its transfer functions, and it was made a study of errors which affects maintaining gain oscillations and frequency oscillation. The paper point out the experimental results obtain through RC oscillators implementation with PA 630 current conveyors showing that the current mode oscillators appears to be an interesting approach from the perspective of the simplicity/performance compromise.

Keywords— frequency, gain, oscillator.

I. INTRODUCTION

In this paper we present an intuitive method for the creation of a new current mode oscillator structure, based on the understanding of the way the oscillation frequency is being determined and the loop transfer ratio is being adjusted [1]. This method results in a new and elegant structure.

Here are some of the features such a structure is intended to achieve.

- The frequency tuning is to be done through the modification of a single passive component, preferably a resistor. For high frequency usage the use of a varicap diode is accepted. For frequency tuning that is achieved through multiple components, no precise pairing of those is required.
- The loop transfer control is done independently from the oscillation frequency. Although there are multiple structures that comply to this requirement (at least within the boundaries of an initial approximation that ignores the side effects), most of such oscillators have a certain dependency of the loop transfer tuning to the oscillation frequency.
- A wide range of frequency variation must be supported, which usually means more complicated circuitry needs to be put in place.
- It is desirable that all capacitors have one end connected to the ground so that these oscillators can be implemented within an integrated circuit.
- There should also be available a simple means of control for the oscillation amplitude. This is vital for an oscillator because a slight variation of the oscillation amplitude can lead to the oscillation being distorted or even interrupted.
- For RC oscillators it may seem absurd to talk about obtaining a high degree of thermal stability of the oscillation frequency as long as none of them are very stable. We should on the other hand make sure that the oscillation frequency is not dependant on any of the parameters of the active components, since these elements usually have a significant dependency to the environmental temperature.

The block schema of a RC oscillator is presented in figure 1[1], where by A we mean an ideal current oscillator, whose gain is:

$$G = \frac{i_1}{i_i} \tag{1}$$

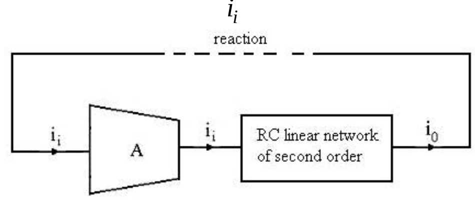

Fig. 1 The ideal current oscillator

For a linear second order RC network, the current transfer function can be described as

$$\beta(s) = \frac{i_0}{i_i} = \frac{a_2 s^2 + a_1 s + a_0}{b_2 s^2 + b_1 s + b_0} \tag{2}$$

Where s is the complex frequency value and a_n(n=0,1,2) are real numbers. In practice only stabile networks are of interest so the b_n (n=0,1,2) coefficients must be positive real numbers. The Barkhausen oscillation initiation condition, applicable for any positive reaction oscillator is leading to the following equation:

$$\frac{a_2 s^2 + a_1 s + a_0}{b_2 s^2 + b_1 s + b_0} A = 1 \tag{3}$$

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$$s^2\left(a_2A-b_2\right)+s\left(a_1A-b_1\right)+a_0A-b_0=0 \qquad (4)$$

For the circuit in figure 1 with the closed reaction loop, the ω_0 angular oscillation frequency and the necessary gain for oscillation maintaining (G_0) have the below expressions

$$\omega_0=\sqrt{\frac{b_0}{b_2}}\sqrt{\frac{1-\frac{a_0b_1}{a_1b_2}}{1-\frac{a_2b_1}{a_1b_2}}} \qquad (5)$$

$$G_0=\frac{b_1}{a_1} \qquad (6)$$

The above scopes will be further called "the oscillation frequency" and "maintaining gain".
Since the gain in 6 must be finite, the a_1 can't be zero while expression (6) allows for a_0 and a_2 to be zero both simultaneously and each at a time.

II.FUNDAMENTAL BLOCKS FOR SECOND ORDER CMO-RC

Different shapes number of acceptable second order transfer functions for harmonic oscillators construction will be given by restrictions towards the equation (2) coefficients in order to equations (5) and (6) make physical sense [14].
Based on the restrictions in the above paragraph, we can identify four types of second order RC networks that can be used for the design of harmonic oscillators and corresponding them, second order RC oscillators. The most general type corresponds to the following parameters: $a_2\neq 0; a_1\neq 0; a_0\neq 0$.
Given this, the oscillation frequency becomes

$$\omega_0=\sqrt{\frac{b_0}{b_2}\frac{1-\frac{a_0b_1}{a_1b_2}}{1-\frac{a_2b_1}{a_1b_2}}}=S_4\sqrt{\frac{b_0}{b_2}} \qquad (7)$$

Where we made the following convention

$$S_4=\sqrt{\frac{1-\frac{a_0b_1}{a_1b_2}}{1-\frac{a_2b_1}{a_1b_2}}} \qquad (8)$$

From equation 7 one notices that, the oscillation frequency is dependent by the S_4 factor, called the scale factor. To investigate the values that S4 factor can equals is a useful way to know the frequency tuning possibilities when ratio b_0/b_2 has a constant value.

For these oscillators the maintaining gain is also given by equation 6.

III. CMO CONSTRUCTIONS

The below paragraph proposes a set of constructive blocks that, by adequate design, will produce the desired transfer ratios. This set is not by far unique, but from the multitude of possible solutions we chose one that complies with the following requirements[20]:

a) the passive components should be designed easily and in a performing manner in monolithic technology.

b) the oscillator should have either two resistors or two capacitors connected to the ground.

The latter requirement has a great practical importance for variable frequency oscillators. For this particular reason a wide known network like the Wien network may not be used for our purpose.

A Active Blocks

The amplifiers presented in figure 2 will be used as active elements[1].

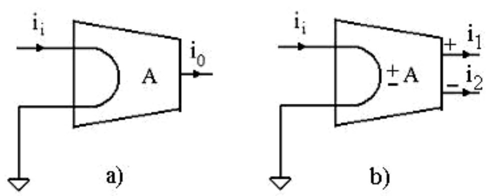

Fig. 2 Active blocks. a) ideal current amplifier; b) ideal current amplifier with a single input and two opposing outputs (SIDO).

The amplifier in figure 2.a is an ideal current amplifier whose gain is

$$\frac{i_0}{i_i}=A \qquad (9)$$

Where A can be both a negative and a positive gain.
The amplifier in figure 2.b is an ideal current amplifier with a single input and two opposing outputs (SIDO). Its behavior can be described by

$$\frac{i_1}{i_i}=A \qquad (10)$$

$$\frac{i_2}{i_i}=-A \qquad (11)$$

where A is positive.

B Passive Blocks

According to the last requirement previously mentioned, the passive blocks taken into consideration will be those from figure 3.
The circuit in figure 3.a is a resistive current divisor with a transfer function described as

$$\frac{i_0}{i_i}=\frac{R_2}{R_1+R_2}=k \qquad (12)$$

The circuit in figure 3.b is a single section CR network whose transfer function is

$$\frac{i_0}{i_i} = \frac{1}{1+sCR} \tag{13}$$

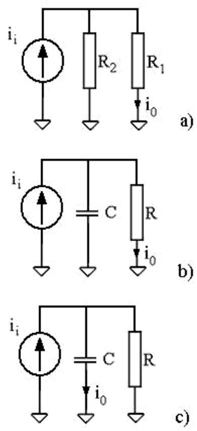

Fig. 3 The passive blocks. a) resistive current divisor; b) single section CR network; c) single section RC network.

And figure 3.c shows a single section RC network with the following transfer function

$$\frac{i_o}{i_i} = \frac{sCR}{1+sCR} \tag{14}$$

C Combined Blocks

This category consists of the simplest combinations of active elements and passive blocks of first order transfer ratios.

The structure that will therefore result and satisfy previous the restrictions be shown in figure 4. The transfer ratio is easily determined by using the (9)-(11) relations in combination with one of the (12)-(14) relations.

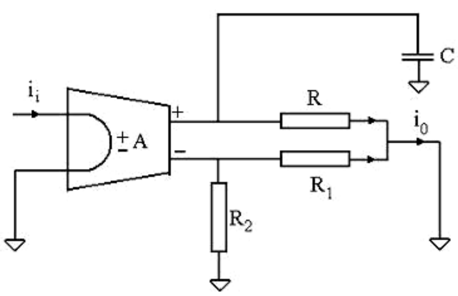

Fig. 4 SIDO amplifier, a current divisor and a single section CR network.

The circuit in figure 4 consists of a SIDO combined with current divisor (figure 3.a) and a single section CR network (figure 3.b) and it will be called *de-phasing network with current divisor and grounded capacitor*. According to relations (11) and (12), the output current of the R_1,R_2 divisor becomes

$$i_D = -kAi_i \tag{15}$$

According to relations (10) and (13), the output current from the RC phase shifting circuit is

$$i_R = \frac{1}{1+sCR} i_i \tag{16}$$

The overall circuit output current is the sum of the above two currents

$$i_0 = i_R + i_D \tag{17}$$

And thus the circuit transfer ratio is

$$\frac{i_0}{i_i} = A\frac{\left(1-k\right)-skRC}{1+sRC} \tag{18}$$

IV. RC OSCILLATOR WITH THE CAPACITOR DE-PHASING NETWORK AND SINGLE SECTION RC NETWORK

With the capacitor de-phasing network(figure 4), a single section RC network(figure 3b) and the ± A and A_1 amplifiers, one obtains the oscillator in figure 5[6].

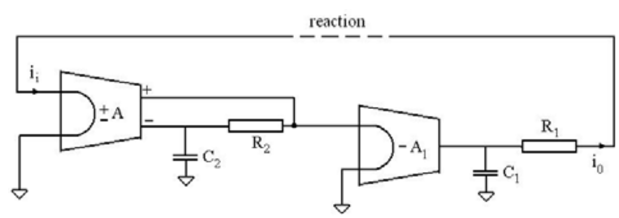

Fig. 5 RC oscillator a single section RC network and capacitor de-phasing network

The loops transfer function is obtained by setting k = 1 in equation 18 and by multiplying the result with the transfer function of the single section RC network which looks like: $\frac{1}{1+sRC}$. Equation 19 is thus obtained

$$\frac{i_0}{i_i} = AA_1\frac{sR_2C_2}{s^2R_1R_2C_1C_2 + s\left(R_1C_1 + R_2C_2\right)+1} \tag{19}$$

The expression of the oscillation sustaining gain is:

$$G_0 = 1+\frac{R_1}{R_2}\frac{C_1}{C_2} \tag{20}$$

And the oscillation frequency is:

$$\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}} \tag{21}$$

The value of the sustaining gain remains constant if the ration of the resistors and capacitors remains constant. For the case when $R_1 = R_2; C_1 = C_2$ (practical use case) one gets G_0=2.

The A_1 amplifier in figure 5 is not replaceable because it has the role of separating the networks. We can observe the oscillator would run similarly should R_1 and C_1 would be interchanged with R_2 and C_2.
Then the transfer ratio becomes:

$$\frac{i_0}{i_1} = AA_1 \frac{sR_1C_1}{s^2R_1R_2C_1C_2 + s\left(R_1C_1 + R_2C_2\right) + 1} \tag{22}$$

The oscillation frequency remains the same and the oscillation sustaining gain is:

$$G_0 = 1 + \frac{R_2}{R_1}\frac{C_2}{C_1} \tag{23}$$

The oscillator behavior is similar to the one of a Wien network oscillator, but has the advantage of having to ground resistors. Actually the above interchange is equivalent to the use of a single section RC network.

A The study of errors

1. Errors caused by the actual values of the conveyors impedances.
We will continue by computing the errors produced by the parameters which are involved in the expressions of the oscillation frequency sustaining gain. The estimation of these errors was done for the oscillator in figure 5.
There were also taken into account the resistances and capacitors for the inputs/outputs of the integrated PA630 conveyors, opposite to the ideal conditions (R_X=0; R_Z=∞; C_Z=0). According to the catalogue information, the input resistance at input X of the inverted current repeater has the typical value R_X=2 Ω ,the capacitor C_Z=11pF, and the output resistance R_Z=3MΩ. For the schematics in figure 5, the output capacitors of the (C_X) conveyors come in parallel with C_1 and C_2, while the Rx resistance of SIDO comes in series with R_1. If one neglects the input resistance of A_1, one gets the following loop transfer ratio

$$\begin{aligned}\left(\frac{i_0}{i_i}\right)^* &= [AA_1sR_2\left(C_2 + C_Z\right)]/ \\ &/\{s^2\left(R_1 + R_X\right)R_2\left(C_1 + C_Z\right)\left(C_2 + C_Z\right) + \\ &+ s\left[\left(R_1 + R_X\right)\left(C_1 + C_Z\right) + \left(C_2 + C_Z\right)R_2\right] + 1\}\end{aligned} \tag{24}$$

The expression of the loop gain becomes

$$G_0^* = 1 + \frac{\left(R_1 + R_X\right)}{R_2}\frac{C_1 + C_Z}{C_2 + C_Z} \tag{25}$$

While the expression of the oscillation frequency is:

$$\omega_0^* = \frac{1}{\sqrt{\left(R_1 + R_X\right)R_2\left(C_1 + C_Z\right)\left(C_2 + C_Z\right)}} \tag{26}$$

By (*) we mark here the parameters that are prone to errors.

$$\varepsilon_A^M = \left|\frac{M_A - M_0}{M_0}\right| 100\left[\%\right] \tag{27}$$

Where by ε_A^M we mean the relative error in percentages, M_0 stands for the errorless value of M and M_A stands for the value of M affected by error A.

For the estimation of the inserted errors we will take into consideration the worst case scenario, meaning R_1=R_2=10 Ω, respectively C_1=C_2=68pF (corresponding to the minimal values of R and C used in the practical implementation).

We therefore obtain the numerical values of: $\varepsilon_{imp}^{G_0} = 10\%$ and $\varepsilon_{imp}^{\omega_0} = 27,6\%$.
For the usual values R_1=R_2=100 Ω respectively C_1=C_2=1nF one obtains $\varepsilon_{imp}^{G_0} = 1\%$ and $\varepsilon_{imp}^{\omega_0} = 1,19\%$.

2. Errors due to the inaccuracy of the conveyor current transfer

To estimate these errors we will consider the below expression of the conveyor current transfer

$$\frac{i_0}{i_i} = 1 \pm \delta \tag{28}$$

Where $\delta = 0,5\%$ is according to the catalogue data.
We must take into account the fact that a differential stage is made of two current conveyors. Considering $i_+ = \left(1 + \delta_2\right)i_i$ respectively $i_- = \left(1 + \delta_1\right)i_i$, the expression of the transfer ratio becomes:

$$\left(\frac{i_0}{i_i}\right)^* = AA_1 \frac{sC_2R_2\left(1 + \delta_2\right) + \delta_2 - \delta_1}{s^2C_1C_2R_1R_2 + s\left(R_1C_1 + R_2C_2\right) + 1} \tag{29}$$

The oscillation frequency:

$$\omega_0^* = \sqrt{\frac{1}{C_1C_2R_1R_2}}\sqrt{1 - \frac{\left(\delta_2 - \delta_1\right)}{1 + \delta_2}\left(C_1R_1/C_2R_2 + 1\right)} \tag{30}$$

The sustaining gain is:

$$G_0^* = \frac{1}{1 + \delta_2}\left(1 + C_1R_1/C_2R_2\right) \tag{31}$$

With expressions 30 and 31 we can now calculate the relative errors of the oscillation frequency and the sustaining gain:

$$\varepsilon_\delta^{\omega_0} = \left|1 - \sqrt{\frac{1 - \delta_2 + 2\delta_1}{1 + \delta_2}}\right| \tag{32}$$

$$\varepsilon_{\delta}^{G_0} = \left| 1 - \frac{1}{1+\delta_2} \right| \qquad (33)$$

In the worst case scenario ($\delta_1 = 0,5\%$ respectively $\delta_2 = -0,5\%$) one gets $\varepsilon_{\delta}^{G_0} = 0,5\%$ respectively $\varepsilon_{\delta}^{\omega_0} = 1\%$.

3. Errors due to the miss-pairing of the passive components

We only consider the tolerance of the manufacturing process of the components and neglect the temperature dependency of the components values.
Thus

$$R_1 = R_{1n}\left(1 \pm \lambda_1\right); R_2 = R_{2n}\left(1 \pm \lambda_2\right);$$
$$C_1 = C_{1n}\left(1 \pm \lambda_3\right); C_2 = C_{2n}\left(1 \pm \lambda_4\right),$$

where λ_{1-4} is the components tolerance (in percentages).

We can say that the components tolerance has a small impact on the sustaining gain expression, because they only come in the shape of ratios R_1 / R_2, while these ratios have small values (2%) for integrated implementations.
The oscillation frequency is

$$\omega_0^* = \frac{1}{\sqrt{R_{1n}R_{2n}C_{1n}C_{2n}\left(1\pm\lambda_1\right)\left(1\pm\lambda_2\right)\left(1\pm\lambda_3\right)\left(1\pm\lambda_4\right)}} \qquad (34)$$

And the relative error of the oscillation frequency is:

$$\varepsilon_{\lambda}^{\omega_0} = \left| 1 - \frac{1}{\sqrt{\left(1\pm\lambda_1\right)\left(1\pm\lambda_2\right)\left(1\pm\lambda_3\right)\left(1\pm\lambda_4\right)}} \right| \qquad (35)$$

Should we consider equal tolerances for all four components, the worst case scenario is when they all are positive:
For $\lambda_1 = \lambda_2 = \lambda_3 = \lambda_4$ =1% we get $\varepsilon_{\lambda}^{\omega_0}$ =1,9%, for $\lambda_1 = \lambda_2 = \lambda_3 = \lambda_4$ =5%, $\varepsilon_{\lambda}^{\omega_0}$ =9,2% and for $\lambda_1 = \lambda_2 = \lambda_3 = \lambda_4$ =10% we get $\varepsilon_{\lambda}^{\omega_0}$ =17,35%.

4. Errors due to the oscillation amplitude limitation circuit

For a minimal implementation the oscillation amplitude limitation circuit is made of a resistance divisor bridged by two anti-parallel coupled diodes (figure 6a). Since the limitation is done by a non-linear circuit, important errors may appear both related to variations from the nominal oscillation frequency, and the frequency specter of the resulting signal (we may get typical harmonica distortions of 3-5 %).

The evaluation of such errors is very complex and can only be approached by the use of a computer. One should also take into account the fact that these latter errors can be used to compensate for the previous ones.

B Practical Implementations

An oscillator as shown in figure 5 has been implemented in practice. On the printed circuit we created, there was a ground plane created on the components side, with the purpose of protecting the assembly from external interference. With the same purpose the entire circuit has also been inserted in a metal case. The circuit power source is a $V_{CC}=V_{SS}=\pm 10V$ (E4109), and the oscillations visualization was done by an E0103-A oscilloscope, while the oscillation frequency was done with a BM526 (TESLA) frequency meter. For the created oscillator, the required gain for the sustaining of a unitary transfer ration in the loop was ensured by the first differential stage, all the other active stages having a unitary gain. One can't discard the use of the current repeating stage because we require the RC de-phasing networks to be evaluated against their current behavior.

For the implementation of a SIDO stage we require two PA630 current conveyors, while for the implementation of a current repeater there was only one PA630 current conveyor required[8].

Figure 6a schematically displays the configuration of a SIDO, while figure 6b is showing the configuration of a simple current repeater.

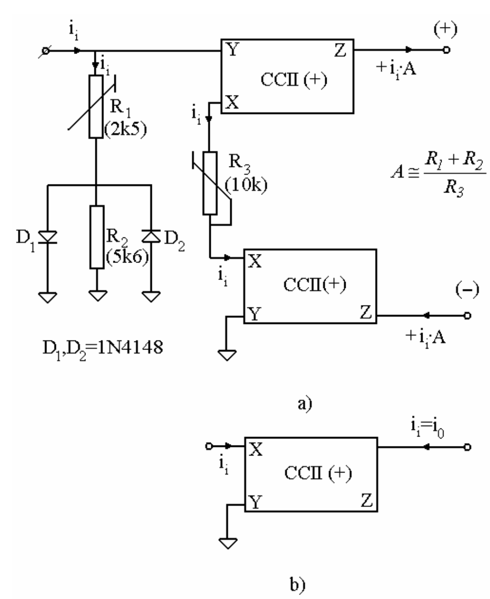

Fig. 6a) SIDO amplifier, b) The configuration of a simple current repeater.

Note: The collectors of the Y input transistors we're connected to –Vss for all CCII.
From figure 6a stands out the oscillation limitation circuit, formed of the D_1 and D2 diodes, which are bridging the R_1, R_2 input resistive divisor. With an initial approximation the SIDO gain is given by $\left(R_1+R_2\right)/R_3$. We should take notice that if for any given reason the oscillation amplitude may tend to grow, the D_1 and D_2 diodes will tend to further open, thus decreasing the initial value of $R_1 + R_2$ and so the entire transfer ratio of the loop will decrease.

To avoid the latch-up phenomenon specific to the PA630 current conveyor, we will use (if required) on the Z outputs of the conveyor, some 3.6V Zenner diodes (DZ3V6), in series with 1N4148 fast diodes, as shown in the PA630 catalogue. Obviously the use of such diodes will lead to a change in the loop transfer ratio and also of the oscillation amplitude.

The conveyors polarization was done according to the catalogue schemas, by ensuring a polarization current of 1.26mA[8]. In order to implement the RC de-phasing circuits, we used 1% tolerance metal foil resistors, high precision capacitors (1-2% precision) and high tolerance electrolytic capacitors (20-50% tolerance).

The following measurements have been performed:

a) The oscillation amplitude has been fixed to 280mV, also $C_1=C_2=47nF$ and also $R_1=R_2$ we're varied from 10 to 196 KΩ. For resistor values higher then 100 KΩ, there was observed a slight drop in the oscillation frequency. For values of R1=R2 ranging from 215Ω to 196 KΩ(911:1) we got a frequency variation ranging from 17Hz to 15.7kHz (923:1), with a 5% precision.

For values of $R_1=R_2\leq 200\Omega$, we noticed an increase of errors up to about 55%. The experimental results are displayed in figure 7.

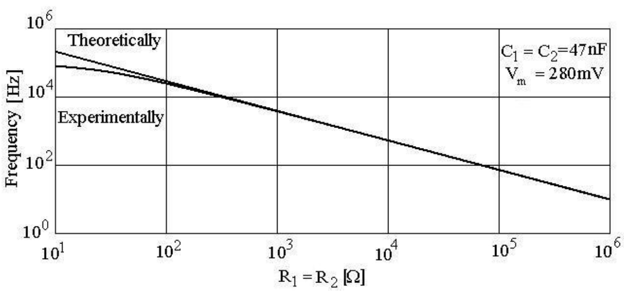

Fig. 7

b) We continued by setting $R_1=R_2=1K\Omega$ and modifying $C_1=C_2$ ranging from 22,6pF to 100μF. For values of $C_1=C_2$ ranging from 10nF to 10μF (1000:1) we obtained a frequency variation from 16Hz to 16KHz (1000:1) with a 5% precision. For values of $C_1=C_2=1nF$ errors may grow up to 85%. Also for values of $C_1=C_2>10\mu F$ errors may grow up to 30% and are caused by the high tolerances of the electrolytic capacitors. The experimental results are graphically presented in figure 8.

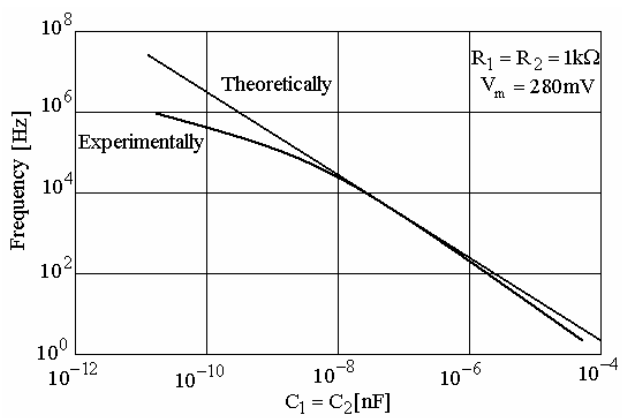

Fig. 8

V. RC OSCILLATOR USING A CR NETWORK WITH TWO SECTIONS

The oscillator in figure 9 is built using a CR net with two sections that are differentially coupled to the SIDO amplifier (± A) and to the simple amplifier $+A_1$[7]. The transfer ratio for the loop is obtained by multiplying the transfer loop of the double section CR net with the gains of the two current amplifiers (A and A_1). So, the expression is:

$$\frac{i_0}{i_i} = AA_1 \frac{sR_2C_2}{s^2R_1R_2C_1C_2 + s\left(R_1C_1 + R_2C_2 + R_1C_2\right)+1} \qquad (36)$$

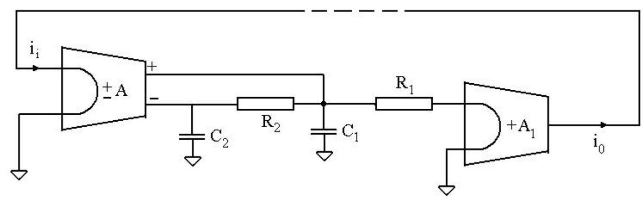

Fig. 9 RC Oscillator using a CR network with two sections

The sustaining gain expression is (37) while the oscillation frequency is (38)

$$G_0 = 1 + \frac{R_1}{R_2} + \frac{R_1}{R_2} \cdot \frac{C_1}{C_2} \qquad (37)$$

$$\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}} \qquad (38)$$

Expression (37) shows that the value of the sustaining gain is constant if the ratios between the resistors and capacitors remain constant, even if their particular values may be changed.

The schematic in figure 9 can be utilized to build variable frequency oscillators that can be brutally tuned by switching two equal resistors of fine tuned by the use of a variable capacitor with two identical sections.

In this case, to maintain the oscillations we need the following condition to be true:

$$A \cdot A_1 = 3 \qquad (39)$$

It is noticed that transfer function is identically with a Wien net one which is working in current mode, but in comparison with an oscillator with Wien nets, the oscillator from figure 9 has the advantage that both capacitors are connected to the ground.

Theoretically speaking the A_1 amplifier in figure 9 could be eliminated if the SIDO amplifier offers sufficient gain, so that the required sustaining gain to be obtained.

One can also note that theoretically speaking the oscillator in figure 9 works in a similar way if the resistors and capacitors can be interchanged. Thus if R_1 and C_1 are interchanged, as well as R_2 and C_2, the transfer ratio becomes:

$$\frac{i_0}{i_1} = \frac{AA_1 sR_2C_2}{s^2R_1R_2C_1C_2 + s\left(R_1C_1 + R_2C_2 + R_2C_1\right) + 1} \quad (40)$$

The oscillation frequency is then given by (38), and the sustaining gain is:

$$G_0 = 1 + \frac{C_1}{C_2} + \frac{C_1}{C_2} \cdot \frac{R_1}{R_2} \quad (41)$$

The oscillator's behavior is again equivalent to that of a Wien net oscillator, but in this second case it has the advantage of having two resistor connected to the ground.

In fact the above change is equivalent to the use of a double section RC net.

A The study of errors

1 Errors due to the actual conveyor impedances

For the circuit in figure 9, the output capacitors (C_Z) of the SIDO(made out of two CCII) are put in parallel to the C_1 and C_2 capacitors, the (R_Z) output resistor of the non-inverting terminal of the SIDO is in parallel with C_1 , R_1 and the output (R_Z) output resistor of the inverting terminal of the SIDO is in parallel with C_2 . If R_1, R_2<< R_Z their effect is neglected.

The input resistance of the A_1 amplifier (R_X) is in series with R_1. In this case the loop transfer ratio becomes:

$$\left(\frac{i_0}{i_i}\right)^* = AA_1 sR_2\left(C_2+C_Z\right) / \left\{s^2\left(R_1+R_X\right)R_2\left(C_1+C_Z\right)\left(C_2+C_Z\right)+ \right. \\ \left. +s\left[\left(R_1+R_X\right)\left(C_1+C_Z\right)+R_2\left(C_2+C_Z\right)+\left(R_1+R_X\right)\left(C_2+C_Z\right)\right]+1\right\} \quad (42)$$

The sustaining gain is:

$$G_0^* = 1 + \frac{R_1+R_X}{R_2} + \frac{R_1+R_X}{R_2} \cdot \frac{C_1+C_Z}{C_2+C_Z} \quad (43)$$

While the oscillation frequency is:

$$\omega_0^* = \frac{1}{\sqrt{\left(R_1+R_X\right)R_2\left(C_1+C_Z\right)\left(C_2+C_Z\right)}} \quad (44)$$

For the estimation of the inserted errors we will take into consideration the worst case scenario, meaning $R_1=R_2=10\,\Omega$, respectively $C_1=C_2=68$pF (corresponding to the minimal values of R and C used in the practical implementation). We therefore obtain the numerical values of: $\varepsilon_{imp}^{G_0} = 13,3\%$ and $\varepsilon_{imp}^{\omega_0} = 27,6\%$. For the usual values $R_1=R_2=100\,\Omega$ respectively $C_1=C_2=1$nF one obtains $\varepsilon_{imp}^{G_0} = 1,3\%$ and $\varepsilon_{imp}^{\omega_0} = 1,19\%$.

2 Errors due to the inaccuracy of the conveyor current transfer

To estimate these errors we will consider the below expression of the conveyor current transfer

$$\frac{i_0}{i_i} = 1 \pm \delta \quad (45)$$

Where $\delta = 0,5\%$ is according to the catalogue data. We must take into account the fact that a differential stage is made of two current conveyors. Considering $i_+ = \left(1+\delta_2\right)i_i$ respectively $i_- = \left(1+\delta_1\right)i_i$, the expression of the transfer ratio becomes:

$$\left(\frac{i_0}{i_i}\right)^* = \frac{AA_1\left[sC_2R_2\left(1+\delta_2\right)+\delta_2-\delta_1\right]}{s^2C_1C_2R_1R_2 + s\left(R_1C_1+R_2C_2+C_2R_1\right)+1} \quad (46)$$

The oscillation frequency:

$$\omega_0^* = \sqrt{\frac{1}{C_1C_2R_1R_2}}\sqrt{1-\frac{\left(\delta_2-\delta_1\right)}{1+\delta_2}\left(1+\frac{R_1}{R_2}+\frac{C_1R_1}{C_2R_2}\right)} \quad (47)$$

The sustaining gain is

$$G_0^* = \frac{1}{1+\delta_2}\left(1+\frac{R_1}{R_2}+\frac{C_1R_1}{C_2R_2}\right) \quad (48)$$

With expressions 47 and 48 we can now calculate the relative errors of the oscillation frequency and the sustaining gain:

$$\varepsilon_\delta^{\omega_0} = \left|1-\sqrt{\frac{1-2\delta_2+3\delta_1}{1+\delta_2}}\right| \quad (49)$$

$$\varepsilon_\delta^{G_0} = \left|1-\frac{1}{1+\delta_2}\right| \quad (50)$$

In the worst case scenario ($\delta_1 = +0,5\%$ respectively $\delta_2 = -0,5\%$) one gets $\varepsilon_\delta^{G_0} = 0,5\%$ respectively $\varepsilon_\delta^{\omega_0} = 1,49\%$.

3. Errors due to the miss-pairing of the passive components. Identically as in figure 5.

4. Errors due to the oscillation amplitude limitation circuit. Identically as in figure 5.

VI. PRACTICAL IMPLEMENTATIONS

An oscillator as in figure 9 has been implemented in practice.

The following measurements have been performed:

The oscillation amplitude has been fixed to 280mV.

a) We fixated $R_1=R_2=1$KΩ and tuned $C_1=C_2$ between 136pF and 100µF. For capacitor values under 1nF the loop gain needed to be diminished.

For values of $C_1=C_2$ between 5,651nF and 10µF (1785:1) we went through a frequency domain between 16Hz and 28,164 KHz (1760:1) with a precision of approximately 5%. We can demonstrate that for capacitors values below 1nF, the errors rise up to about 67%. Also for large values of the capacitors, the errors also reach to 50%, but in this case these errors can be blamed on the high tolerance of the electrolytic capacitors being used.

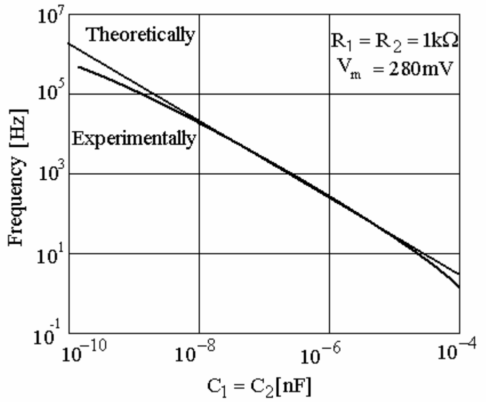

Fig. 10

b) We fixated C_1=C_2=47nF and R_1=R_2 have been modified between 10Ω and 61,9KΩ. The oscillation amplitude has been fixed at 280mV. For values of R_1=R_2>50KΩ the loop gain also needed to be changed and same for R_1=R_2<100Ω. For values of R_1=R_2 ranging from 215Ω to 61,9K Ω(283:1) the obtained frequencies ranged from 55Hz to 15KHz (272: 1) with a 5% precision. When R_1=$R_2 \leq$200Ω the errors grow up to 100%.

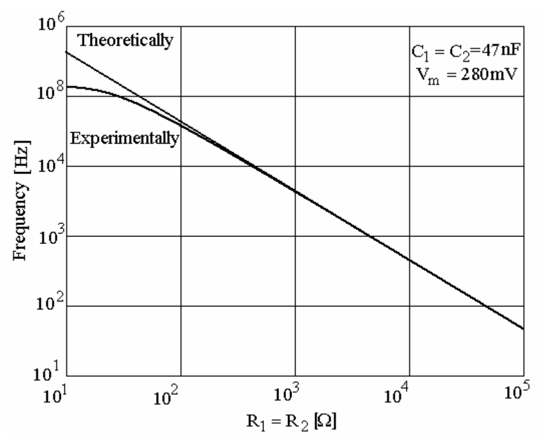

Fig. 11

VII. THE CIRCUIT OF THE SECOND ORDER CURRENT MODE RC OSCILLATOR

A wide variety of second order oscillators can be synthesized just by using the above functional blocks. We will now present an oscillator whose reaction loop is characterized by a complete second order transfer ratio. The circuit in figure 10 is a cascade connection of two de-phasing network with current divisor and grounded capacitor, the likes of the one in figure 4[12]. The transfer ratio of the loop is obtained from expression (18) by multiplication with A_3's gain

$$\frac{i_0}{i_i} = A_1 A_2 A_3 \frac{s^2 k_1 k_2 R_1 R_2 C_1 C_2 + (k_1 - 1)(k_2 - 1)}{s^2 R_1 C_1 R_2 C_2 + s(R_1 C_1 + R_2 C_2) + 1} + \\ + \frac{s\left[k_1 (k_2 - 1) R_1 C_1 + k_2 (k_1 - 1) R_2 C_2\right] +}{s^2 R_1 C_1 R_2 C_2 + s(R_1 C_1 + R_2 C_2) + 1} \tag{51}$$

The expression of the scale factor is

$$S_4 = \sqrt{\frac{(1-k) + \frac{R_1}{R_2}\frac{C_1}{C_2}(1-k_2)}{k_2 + \frac{R_1}{R_2}\frac{C_1}{C_2}k_1}} \tag{52}$$

The maintaining gain is

$$G_0 = \frac{1 + \frac{R_1}{R_2}\frac{C_1}{C_2}}{k_2(k_1 - 1) + \frac{R_1}{R_2}\frac{C_1}{C_2}k_1(k_2 - 1)} \tag{53}$$

The general expressions (51) to (53) are too complex to emphasize possible remarkable properties of those oscillators.

From a practice perspective only the situation where the two current divisors from the de-phasing circuits have identical transfer ratios is of interest, therefore $k_1 = k_2 = k$.

Given this, the scale factor becomes

$$S_4 = \sqrt{\frac{1-k}{k}} \tag{54}$$

And the maintaining gain is

$$G_0 = \frac{1}{k(1-k)} \tag{55}$$

Expression (55) shows that in this case the maintaining gain is independent of the R_1,R_2,C_1 and C_2 variations. This means that the oscillator can be tuned by varying one ore several components at a time or simultaneously without having to vary the loop gain.

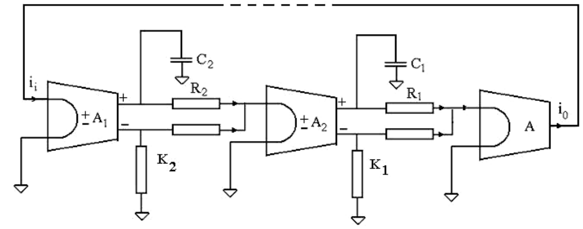

Fig.12 The second order current mode RC oscillator

A Practical Implementations

We used OrCad PSpice 9.0 to verify the design. For this we created a PA 630 Phototronics current conveyor model, by using the NS3905 and NS3904 transistors as active elements (see figure 13). The sinusoidal oscillation (see figure 14) can be visualized at the measuring point situated between the collectors of the Q22 and Q26 transistors. One must note the stability of the amplitude and oscillation frequency.

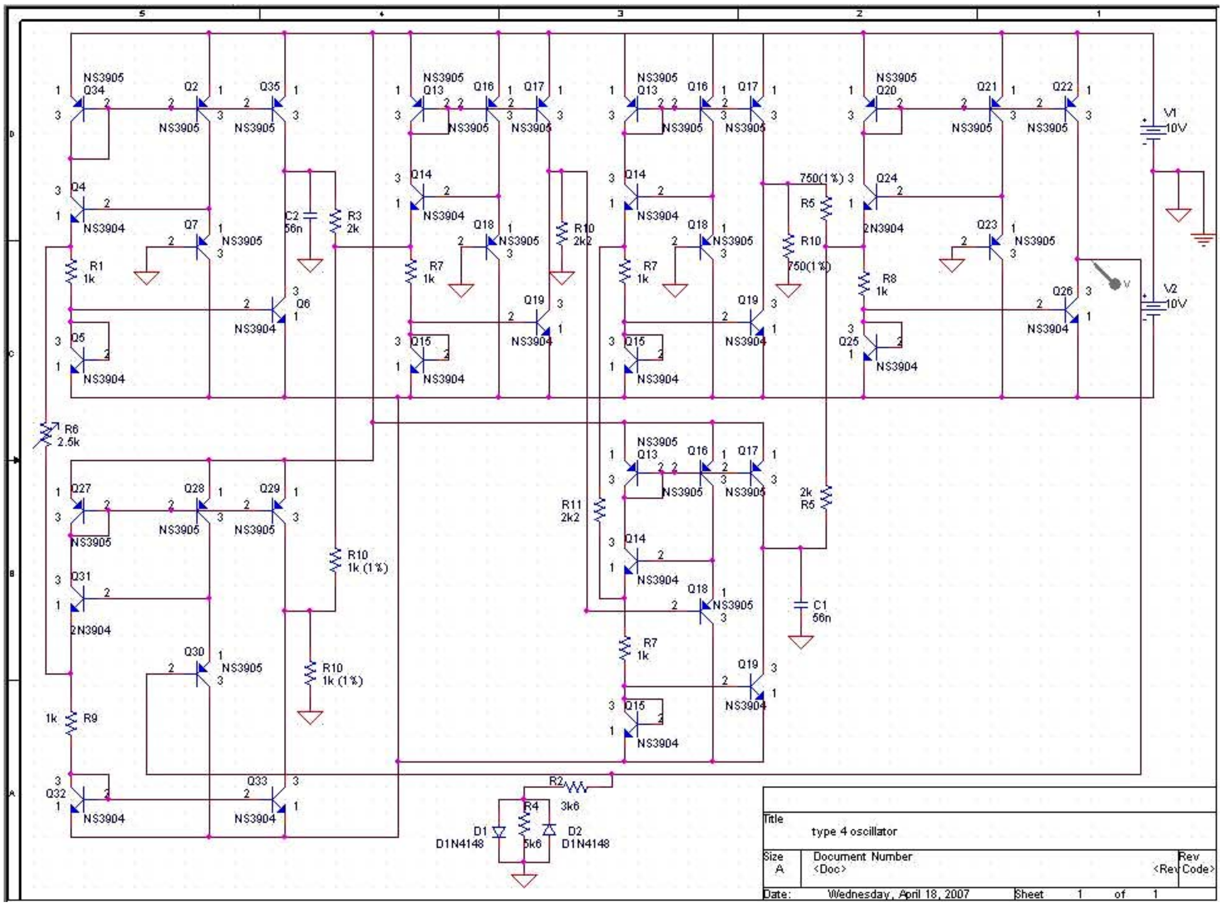

Fig.13 RC oscillator with PA630

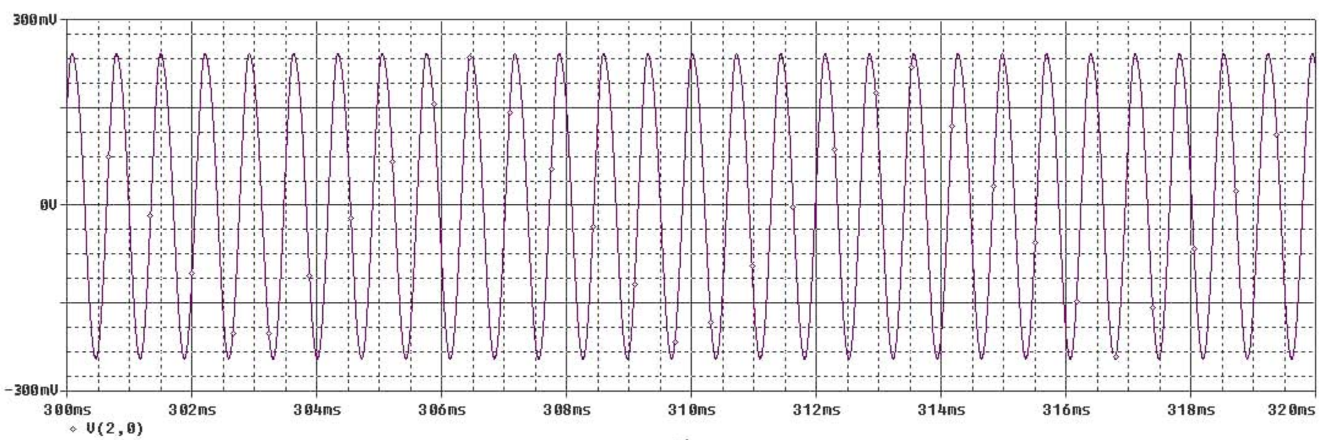

Fig. 14 The results of the numerical simulation with PSpice

For the circuit in figure 12 the following measurements have been performed:

a) The following components with fixed values have been used: R_1=R_2=1KΩ, C_1=56nF respectively k_1=k_2=1/2, while C_2 has been varied within 68pF-1000µF (electrolytic). The oscillation amplitude has been tuned to approximately 280mV and remained almost the same for the entire C_2 value variation. <u>Also the loop gain needed no adjustments.</u> For a variation of C_2 from 10,94nF to 70µF (about 6500:1) the frequency variation ranged from 76Hz to 6127Hz (80:1) with a precision under 5%. For a variation of C_2 from 2nF to 1000µF(500000:1), the frequency varied from 18Hz to 12723Hz (706:1), with a precision below 15%. We should note that for values of $C_2 \leq$ 1nF errors may grow up to 70% even if the oscillation frequency is in the order of tens of kHz.

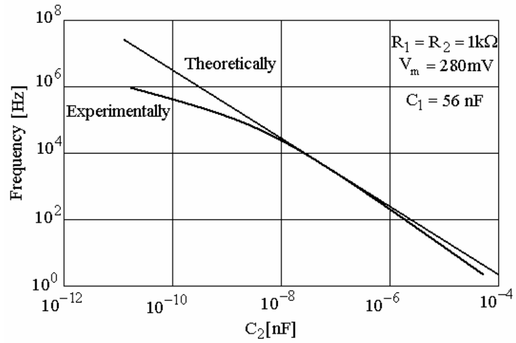

Fig. 15 The experimental results for R_1=R_2=1KΩ, V_m=280mV

b) The following components with fixed values have been used: C_1=C_2= 47nF, R_1=1KΩ and respectively k_1=k_2=1/2 , while R_2 will vary from 10Ω to 1196KΩ. The oscillation frequency was initially adjusted to 280mV and later dropped for R_2 $\leq$100Ω. Also for R_2 $\leq$100Ω the loop gain needed adjustments (gain increase). For a variation of R_2 ranging from 215Ω to 1196KΩ (5500:1) the frequency range varied was 95Hz to 6972Hz(73:1) with a precision higher than 5%. For values of R_2 $\leq$100Ω, errors raise up to 50% (R_2=10Ω) and are caused by the increase in the input impedance module value on the X conveyor connection, as the frequency increases.

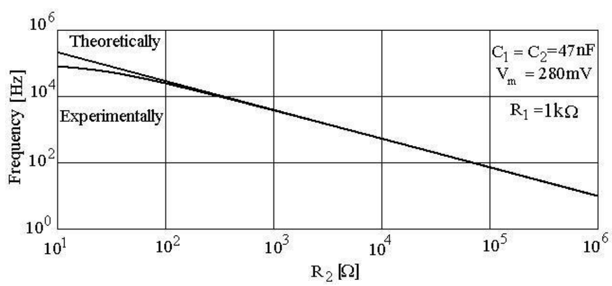

Fig. 16 The experimental results for C_1=C_2=47nF, V_m=280mV

c) The following components with fixed values have been used: R_1=R_2=2KΩ, C_1=C_2=56 nF respectively k_2=1/2 and k_1 was varied. The oscillation amplitude was maintained at 280mV. One must mention that the loop gain needed adjustment for each change in the value of k_1 . Since for each two consecutive values (they change by about 2% each time) it was required for the gain to be readjusted, it can be supposed that a 2% variation can lead to the oscillations fading out. The errors that occur (around 8%) could be due to $k_2 \neq 1/2$ which impacts the scale factor formula.

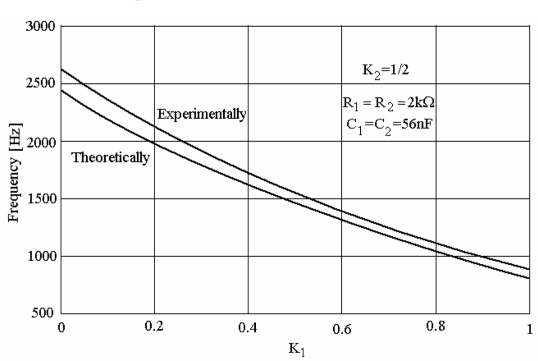

Fig. 17

VIII. Conclusions

From the experimental measurements we can conclude the following related to the design of current mode oscillators by current conveyors:

1. First of all one must notice that the circuit presented here can work in a wide range of frequency just by modifying a passive component and without the use of a complex AAC (automatic amplitude control) circuit. That is why this oscillator is useful for the design of vobulated oscillators.
2. Circuits may work in a wide range of frequencies, should the components pairing be better than 2-5 %, in order to have a reasonable precision.
3. We also experience a higher precision of the oscillation frequency with the change of the resistors when the capacitors are constant than with the change of the capacitors with constant resistances. This is a consequence of the low tolerance of the resistors used (1%) compared with the one of the capacitors (2-10%).
4. For low values of the capacitors (<1nF) even at tens of KHz frequency values errors can be quite large and this can be explained by the inductive behavior of the conveyors input impedances.
5. Also for small values of resistances (smaller than several hundreds ohms) errors can be quite significant and can be explained by the growth with the frequency of the conveyor input impedance value.
6. From a practical perspective the solution has the disadvantage of not ensuring the polarization stability through negative continuous current reaction on the loop.

As a final conclusion, the conveyor based design of the current mode oscillators appears to be an interesting approach from the perspective of the simplicity/performance compromise.

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