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Amplitude Stabilization in Quadrature Oscillator for Low Harmonic Distortion

Andrej Levstek

Faculty of Electrical Engineering, University of Ljubljana, Ljubljana, Slovenia

Abstract: A simple non-linear network for amplitude stabilization in harmonic quadrature oscillator (QO) is analyzed. The oscillation startup conditions for QO topology with two operational amplifiers are used for amplitude adjustment by element values. A closed form expression for the amplitude setting resistance is derived. Total harmonic distortion (THD) is expressed as closed form function of the established amplitude. The derived expressions as well as the oscillator performance are verified by computer simulations and measurements.

Keywords: quadrature oscillator, harmonic distortion, integrator, steady state, sine wave

Stabilizacija amplitude v kvadraturnem oscilatorju za nizko harmonično popačenje

Izvleček: V prispevku je analizirano delovanje preprostega nelinearnega vezja za stabilizacijo amplitude kvadraturnega oscilatorja (QO). Nastavitev amplitude v QO z dvema operacijskima ojačevalnikoma je izvedena z vrednostmi elementov, ki vplivajo na zagonske pogoje. Vrednost upornosti za nastavljanje amplitude je določena z izpeljanim analitičnim izrazom. Harmonično popačenje (THD), ki ga vnaša vezje za amplitudno stabilizacijo, je podano kot funkcija amplitude in vrednosti elementov. Delovanje vezja za stabilizacijo amplitude in ustreznost izpeljanih izrazov je preverjeno z računalniško simulacijo in laboratorijskimi meritvami.

Ključne besede: kvadraturni oscilator, harmonično popačenje, integrator, sinusni signal

** Corresponding Author's e-mail: andrej.levstek@fe.uni-lj.si*

1 Introduction

Oscillators represent an important electronic circuit that is included in almost every electronic device. The majority of them are used for clocking digital circuits that operate in sequential mode. The output signal of such time keeping oscillators is usually a square waveform - digital clock. Stable frequency and steep edges of the rising and falling edge are important properties that have to be assured by adequate design of the circuit. In common, such oscillators are realized as relaxation oscillator [1] (astable multivibrator) utilizing one or two capacitors which in combination with resistors determine the oscillating frequency and its temperature coefficient. For higher accuracy and stability of the frequency crystal and ceramic resonators are used. In the later case the core oscillator actually produces a sinusoidal signal [2] that is then amplified and clipped to become a square wave at the output.

Harmonic oscillators represent an important kind of circuits that operate in a quite different way. Ideally the frequency spectrum of the output signal should consist only of a single spectral line. In reality, there is no such thing as an ideal sine wave generator, but there is a variety of technical solutions approaching to this goal by minimizing certain imperfections. Numerous circuit topologies are used in order to meet the operating frequency band and the tolerated total harmonic distortion (THD).

For the audio frequency band extended up to several 100 kHz RC oscillators with operational amplifiers are an attractive solution. LC oscillators are known to have stable frequency and low distortion, but the physical size of inductances that would be used at the lower part of this frequency band becomes unacceptably big. In this frequency band spectrally pure sinusoidal signal sources with residual distortion less than -66 dB (0.05 %) or -90 dB are required for testing low-distortion devices, such as audio amplifiers or line transformers for digital subscriber lines (xDSL), respectively.

Wien bridge oscillator is a good source of low THD sinusoidal signals if amplitude stabilization is achieved without challenging its low harmonic distortion. This can be achieved with rather complex networks [3]. Amplitude stabilizing solutions that include resistive device with negative temperature coefficient, like NTC or incandescent light bulb, are not suitable for oscillators operating at low voltages. The output signal amplitude is strongly influenced by the temperature of the ambient. In addition, the frequency is not very stable due to the low phase slope $d\varphi/d\omega$ of the feedback network, so relatively big changes of the oscillating frequency are caused by changes of the amplifier's phase response [4].

In this low to moderate frequency range quadrature oscillator (QO) is a better solution for a low distortion sine wave source then the formerly mentioned Wien bridge oscillator. The name quadrature oscillator is used for autonomous linear circuits that generate a pair of orthogonal or quadrature sinusoidal signals that are also maintaining sustained steady state oscilation. The quadrature signals are of the same amplitude with a phase lag $\pi/2$. If only one output signal is required, e.g., previously mentioned distortion measurements, then the signal with lower distortion can be selected.

QO outperform Wien bridge oscillators in the aspects of frequency stability and simple non inertial amplitude stabilization network. The output signal frequency is determined almost exclusively by time keeping passive elements, namely, capacitors and resistors. The oscillating frequency is virtually insensitive to characteristics of active devices, i.e., operational amplifiers. If temperature coefficients of capacitances and resistances of the elements in the QO are of the same magnitude but of opposite sign, then the frequency temperature coefficient is reduced at least for an order of magnitude.

QO in papers published in scientific and technical journals are mostly treated without considering the necessity of amplitude stabilization [5]. Low distortion of the QO signal can be achieved by a simple non-linear resistive circuitry. In other types of oscillator circuits this is often realized by lowering the gain of the amplifying non-linear device [6].

Analytical relations for the output signal amplitude and THD for the selected QO topology and amplitude stabilizing network are derived in the paper. At first, a detailed analysis of the circuit is performed with the aim to determine the conditions at which self oscillating can evolve. The results are used in the design of the non-linear resistive circuit that is then analyzed for its effects on the output. The obtained results are verified by SPICE simulations and experimental measurements.

2 Operation of quadrature oscillator

2.1 Ideal oscillator

In Fig. 1, the basic circuit concept of QO is shown. The elements in this simplified circuit are considered to be ideal, i.e., the voltage gain of operational amplifiers is infinite and frequency independent.

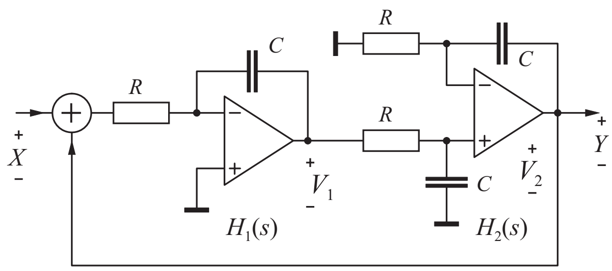

Figure 1: Explanatory scheme of QO based on two operational amplifiers

The circuit contains two integrators and an adding element that is actually just a wire. If the amplifiers are ideal, the system functions are

$$H_1(s) = -\frac{1}{RCs}, \qquad H_2(s) = \frac{1}{RCs} \tag{1}$$

The system function $H(s)$ of the QO is then given as

$$H(s) = \frac{Y(s)}{X(s)} = \frac{T(s)}{1-T(s)} = -\frac{\omega_0^2}{s^2+\omega_0^2} \tag{2}$$

where $T(s) = H_1(s)H_2(s)$ and $\omega_0 = 1/(RC)$. The system function Eq. (2) has a pair of complex poles on the imaginary axis at $\pm j\omega_0$, that means the impulse response of the circuit is an undamped sine wave $V_m \cdot \sin(\omega_0 t)$. The amplitude V_m depends on the input stimulus or on initial conditions, whereas the frequency is set by the time constant RC. In reality it is necessary to obtain oscillating output of defined amplitude and without any starting signal. The system function must include a complex pair of poles lying initially on the right side of the s-plane, i.e., the real part has to be positive in order to obtain an increasing oscillating response. Ubiquitous thermal noise is sufficient to start the oscillatory process [7]. Exponential growth of the amplitude is unlimited as long as the elements remain linear. In order to obtain pure sine wave without any higher harmonic components, the pole pair has to be moved exactly on the imaginary axis when the oscillation reaches the desired amplitude within the linear output range of real amplifiers. This start up process is depicted in Fig.2 by the position of the pair of complex poles.

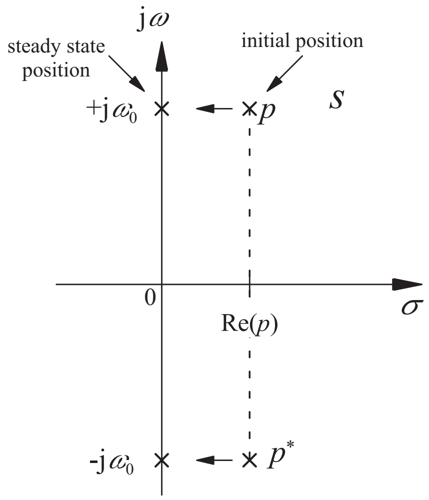

Figure 2: Oscillation determining pole pair during startup

2.2 System function of real QO

In order to find out the necessary starting conditions the most important parameters of real operating amplifiers must be considered. DC parameters like input offset voltage and input bias current have no effect on the operation of QO and are neglected. The most important parameters are open loop voltage gain A and the dominant pole frequency f_{op} that combine in the well known gain bandwidth product $GBW = A \cdot f_{op}$ (unity-gain bandwidth). For the sake of brevity, circular frequency ω_{op} is used in the system function of both operational amplifiers

$$A(s) = \frac{V_{out}}{V_{in}} = -\frac{A}{1+\dfrac{s}{\omega_{op}}} = -\frac{A\omega_{op}}{s+\omega_{op}} \tag{3}$$

The inverting integrator shown in Fig. 3 is slightly changed by the added resistance R_i to provide a means for altering the system function.

Using Eq. (3) the system function of the modified integrator is obtained

$$H_1(s) = \frac{V_1}{V_x} = \\ = \frac{-A\omega_{op}\omega_0}{s^2 + \left[\omega_0(1+\dfrac{R}{R_i}) + (A+1)\omega_{op}\right]s + \left[1+\dfrac{R}{R_i}(A+1)\right]\omega_{op}\omega_0} \tag{4}$$

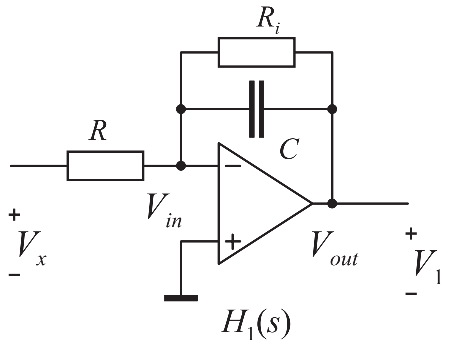

Figure 3: Inverting integrator with resistance R_i. Voltages V_{in} and V_{out} are indicated to clear up the designations used in Eq.

where $\omega_0 = 1/(RC)$. Eq. (4) is simplified by noting $A >> 1$ and $R/R_i << 1$ to obtain

$$H_1(s) = -\frac{A\omega_{op}\omega_0}{s^2 + (A\omega_{op} + \omega_0)s + A\omega_{op}\omega_0 \dfrac{R}{R_i}} \tag{5}$$

without any loss of accuracy, because the exact value of DC open loop gain is not known. The poles of $H_1(s)$ can be determined if the values of gain bandwidth product $A \cdot f_{op}$ of usual amplifiers and the desired frequency f_0 are considered. Values of $A \cdot f_{op}$ are in the range 2 MHz to 10 MHz and thus bigger then the desired frequency f_0 <1 MHz. The denominator of $H_1(s)$ Eq. (5) can be factorized

$$H_1(s) = -\frac{\omega_0\omega_{op}A}{(s + A\omega_{op})(s + \dfrac{R}{R_i}\omega_0)} \tag{6}$$

yielding two real poles

$$p_1 = -\frac{R}{R_i}\omega_0 \ , \ p_2 = -A\omega_{op} \tag{7}$$

The lower pole p_1 depends on the oscillating frequency, whereas p_2 lies at the unity gain bandwidth of the amplifier. The same procedure is carried out for the system function $H_2(s)$ of the noninverting integrator

$$H_2(s) = \frac{A\omega_{op}\omega_0}{s^2 + (A\omega_{op} + \omega_0)s + \omega_{op}\omega_0} \approx \\ \approx \frac{A\omega_{op}\omega_0}{(s + A\omega_{op})(s + \dfrac{\omega_0}{A})} \tag{8}$$

The system function $H(s)$ of the oscillator is obtained from Eq. (2) by applying the derived results in Eqs. (6) and (8).

$$H(s) = -\frac{(A\omega_{op}\omega_0)^2}{\left(s^2 + (A\omega_{op} + \omega_0)s + \omega_0\omega_{op}\right)\left(s^2 + (A\omega_{op} + \omega_0)s + A\omega_{op}\omega_0 \frac{R}{R_i}\right) + (A\omega_{op}\omega_0)^2} \tag{9}$$

Poles of $H(s)$ are can be estimated from the denominator $D(s)$ in and taking into account the approximate values of polynomial coefficients by omitting negligible terms.

$$D(s) = s^4 + 2A\omega_{op}s^3 + (A\omega_{op})^2 s^2 + (A\omega_{op})^2\omega_0 \frac{R}{R_i} s + (A\omega_{op}\omega_0)^2 = \\ = \left(s^2 + 2A\omega_{op}s + (A\omega_{op})^2\right)\left(s^2 + bs + \omega_0^2\right) \tag{10}$$

The oscillator's system function has two real poles at $A\omega_{op}$ and a pair of complex poles whose real part depends on the coefficient b. By matching both sides of Eq. (10) the estimate for b is obtained as well as the position of the complex pair

$$b = \left(\frac{R}{R_i} - 2\frac{\omega_0}{A\omega_{op}}\right)\omega_0 \tag{11}$$

$$p_{1,2} = -\frac{b}{2} \pm j\omega_0 \tag{12}$$

The real part of the pole pair must be positive during start up and zero at steady state. The critical value of R_i is obtained from Eq. (11) respecting $b \leq 0$

$$R_i \geq \frac{R}{2}\frac{A\omega_{op}}{\omega_0} = \frac{R}{2}\frac{Af_{op}}{f_0} = R_0 \tag{13}$$

The amplitude of oscillation increases as long as $R_i > R_0$. The critical value depends on f_0 and the unity gain frequency of the used operational amplifier. The result obtained for the critical resistance in Eq. (13) is not always accurate because several approximations are involved in its derivation. The extent of inaccuracy of R_0 depends on the impact of the involved approximations. If a neglected term is two or more orders of magnitude smaller than the sum then such approximation has virtually no impact.

3 Amplitude stabilization

3.1 Non-linear circuit for amplitude stabilization

The circuit diagram of the oscillator shown in Fig. 4. includes the amplitude stabilizing network, which is made of two *pn* diodes and three resistors, is intended for amplitudes in the range 0.7 V to 2.5 V. Diodes D_1 and D_2 provide the required voltage threshold above which the stabilization process starts. Resistor R_3 improves the sharpness of the voltage threshold V_F by loading the diodes with a current that is several orders of magnitude greater than the current through R_2. Furthermore, resistor R_3 provides a path for the capacitive current of the diodes, so this current is shunted to the ground, thus the phase difference between the output voltage and the current of R_2 and is minimized.

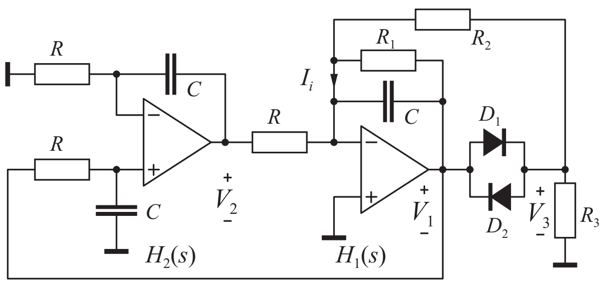

Figure 4: Two amplifier QO with amplitude stabilizing network

The amplitude V_m that is established at steady state can be determined from the plot shown in Fig. 4, that shows the relation between instantaneous values of voltage V_1 and current I_i as denoted in Fig. 3. For voltages $|v_1(t)| < V_F$ the current $i_i(t)$ is proportional to the voltage $i_i(t) = v_1(t) / R_1$. When $|v_1(t)|$ exceeds V_F the current $i_i(t)$ is increased by the contribution that flows through resistor R_2 causing that the apparent resistance felt by the integrator is decreased. Steady state amplitude V_m is obtained at the intersection point where currents are equal

$$I_i(V_m) = \frac{V_m}{R_0} \tag{14}$$

This description is not quite exact, because the wave shape of $i_i(t)$ is not sinusoidal. Only the fundamental harmonic component of current $i_i(t)$ must be respected. The output signal becomes distorted by higher harmonic components that are generated by this nonlinear network. These components should be kept as low as possible in order to minimize *THD* of the generated signal.

In spite of the fact that the piecewise-linear function shown in Fig. 5 is not suitable for accurate determination of the amplitude V_m, this function is useful to understand the influence of circuit parameters on amplitude stability and distortion. The impact of critical resistance's uncertainty ΔR_0 on the amplitude uncertainty ΔV_m increases as the angle of intersection between transfer function $I_i(V_1)$ and straight line V_i/R_0 is being decreased. Stable and well defined amplitude V_m is obtained if the value R_1 is high and R_2 is small, since this

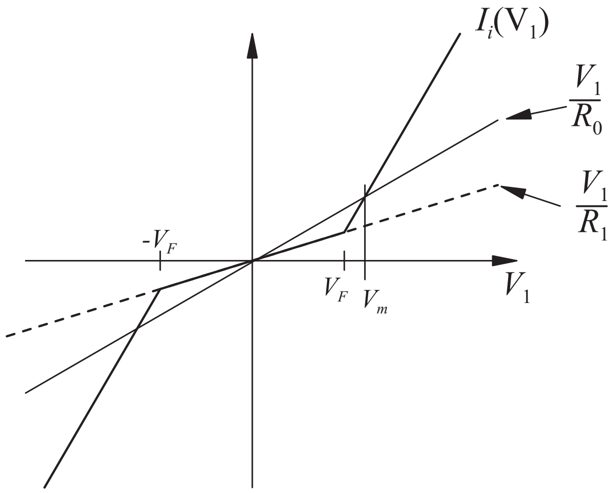

Figure 5: Plot of piecewise-linear transfer function $I_i(V_1)$

gives the largest angle of intersection, and as a consequence lowest sensitivity $\Delta V_m / \Delta R_0$. On the other hand, lower distortion can be obtained if larger tolerance of the amplitude is acceptable. In this case the elements are selected so that the lines intersect at smaller angle. In the next section this trade-off between stability and distortion is clarified by derived analytical expressions for harmonic distortion of the stabilized signal.

3.2 Harmonic analysis

As it is mentioned in the previous section, the plot in Fig. 5. is not adequate for the determination of sustainable amplitude. The effect of amplitude dependant resistance is achieved by the fundamental harmonic component of the current through R_2 that increases for voltages that are above or below V_F or $-V_F$, respectively.

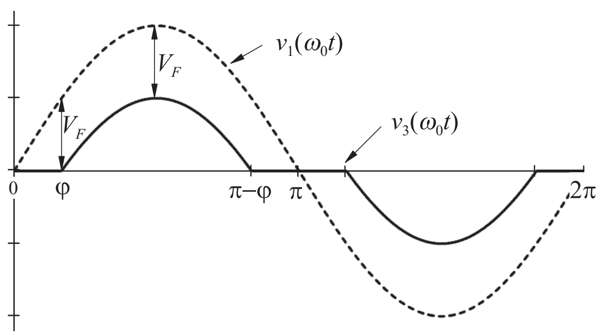

Figure 6: Idealized waveforms of the stabilizing network

For the purpose of harmonic analysis the diodes in the stabilizing network are replaced by piecewise linear relation that can be modeled as serial connection of ideal diode and a voltage source V_F. The obtained idealized waveform is shown in Fig. 6. For brevity, the harmonic components of the periodic voltage V_3 are denoted with C_k.

$$C_k = \frac{2}{T}\int_0^T v_3(t)\sin(k\omega_0 t)dt = \frac{2}{\pi}\int_{\varphi}^{\pi-\varphi}(V_m \sin(\alpha) - V_F)\sin(k\alpha)d\alpha \tag{15}$$

The calculation is simplified since the analyzed signal is odd and so all cosine terms are zero. The integration is preformed by substituting time with phase angle $\alpha = \omega_0 t$. The fundamental component is given by

$$C_1 = V_m\left[\left(1 - \frac{2\varphi}{\pi}\right) + \frac{\sin(2\varphi)}{\pi}\right] - V_F\frac{4}{\pi}\cos\varphi \tag{16}$$

$$\varphi = \arcsin\frac{V_F}{V_m} \tag{17}$$

Higher odd components are given by

$$C_k = V_m\frac{2}{\pi}\left[-\frac{\sin((k-1)\varphi)}{k-1} + \frac{\sin((k+1)\varphi)}{k+1}\right] - V_F\frac{4}{\pi}\frac{\cos(k\varphi)}{k} \tag{18}$$

all even components are zero.

The expression for C_1 in Eq. (16) can be simplified by replacing Eq. (17) with equality.

$$C_{1ap} = V_m\left[\left(1 - \frac{2}{\pi}\frac{V_F}{V_m}\right) + \frac{\sin(2\frac{V_F}{V_m})}{\pi}\right] - V_F\frac{4}{\pi}\cos\left(\frac{V_F}{V_m}\right) \tag{19}$$

First harmonic component of the distorted voltage V_3 is calculated using the both expressions. The results are plotted in Fig. 7 for values V_m above the threshold.

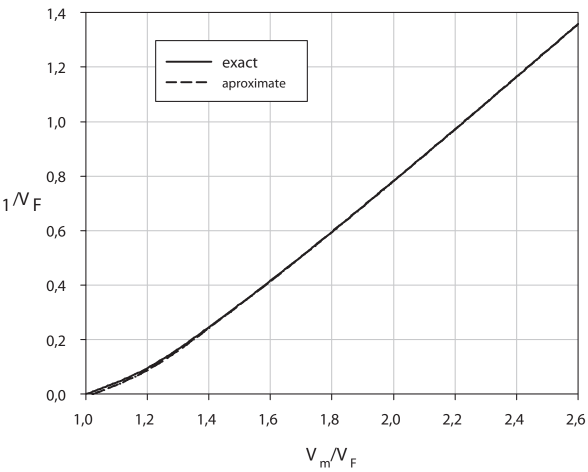

Figure 7: First harmonic component of the voltage V_3 vs. normalized amplitude V_m/V_F

The diagram in Fig. 7 shows that the difference between plots is negligible, so the simpler expression C_{1ap} is used for calculation of the resistances.

At first, R_1 is selected from the range $R_1 > R$, though it is advisable to choose at least $1.5 \times R_0$ to ensure a reliable startup. The selection of R_1 offers the possibility to control the distortion to certain extent. The value of R_2 is obtained as the solution of

$$\frac{V_m}{R_0} = \frac{V_m}{R_1} + \frac{C_{1ap}(V_m, V_F)}{R_2} \tag{20}$$

For the desired amplitude V_m, the value R_2 is then given by

$$R_2 = \left[1 - \frac{1}{\pi}\left(\frac{2}{a} - \sin\left(\frac{2}{a}\right) + \frac{4}{a}\cos\left(\frac{1}{a}\right)\right)\right]\left(\frac{1}{R_0} - \frac{1}{R_1}\right)^{-1} \tag{21}$$

whera a is the amplitude V_m normalized by V_F

$$a = \frac{V_m}{V_F} \tag{22}$$

3.3 Harmonic distortion

Harmonic distortion is caused by the higher harmonic components of the current flowing through R_2. This current flows through the integrating capacitor. The amplitude of harmonic components at the output is given by

$$V_k = \frac{|C_k|}{R_2}|Z_C(k\omega_0)| = \frac{|C_k|}{R_2}\frac{1}{k\omega_0 C} = \frac{|C_k|}{k}\frac{R}{R_2} \tag{23}$$

$$\frac{V_k}{V_m} = \frac{R}{k\,R_2}\frac{2}{\pi} \cdot \left| -\frac{\sin((k-1)a^{-1})}{k-1} + \frac{\sin((k+1)a^{-1})}{k+1} - \frac{2}{a}\frac{\cos(ka^{-1})}{k} \right| \tag{24}$$

The relative distortion in Eq. (24) is valid for normalized amplitudes $a > 1.2$. If the amplitude is closer to the threshold voltage, i.e., $a < 1.2$, then Eqs. (18) and (17) should be inserted in Eq. (23). The calculation of total harmonic distortion is simplified by considering only the third harmonic, because the contribution of higher frequency components to the RMS total distortion voltage is negligible. The distortion expressed in dB is obtained from Eq. (24)

$$THD[\mathrm{dB}] = 10\log\frac{V_{D\,RMS}^2}{V_{1\,RMS}^2} \cong 20\log\frac{V_3}{V_m} \tag{25}$$

The curves of THD shown in Fig.7 can be used for the assessment of achievable performance of generated signal. Distortion of the output voltage depends on the critical value R_0 (Eq. (13)) and the resistance R. Plots in Fig. 8 are obtained for R_0 = 150 kΩ and R = 3.9 kΩ.

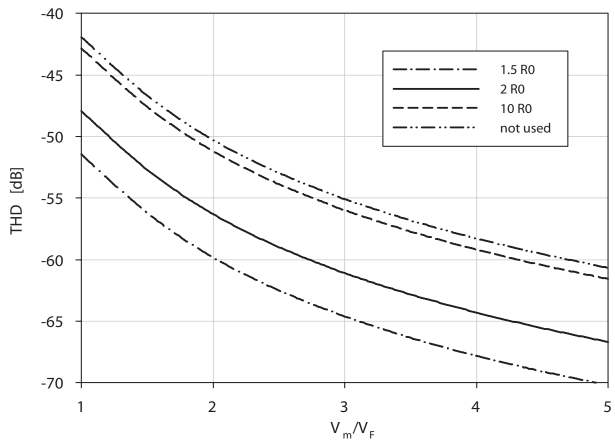

Figure 8: THD of QO with the proposed stabilizing network for typical values of R_1 against normalized amplitude

QO with larger normalized amplitude are featured with lower THD but the actual amplitude can substantially differ from the expected value. If R_1 is selected very close to the critical value R_0 then the maximal value from the worst case analysis must be used.

4 Results

4.1 SPICE simulations

The derived expressions for the amplitude V_m and THD were verified by simulations with SPICE and measurements of the realized circuit. Operational amplifiers TL071 with unity gain bandwidth 3 MHz have been used in the QO for 40 kHz. Ideal and real pn diodes have been used in the stabilizing network. The $I(V)$ characteristic of real diodes has actually no pronounced voltage threshold therefore is V_F selected rather arbitrary. The threshold voltage of the diodes has become more pronounced by the load R_3 = 1 kΩ, which is low when compared to R_1 and R_2. The element values have been calculated with the assumption V_F = 0.65V. The ideal diode has been modeled by b-voltage source connected in parallel with R_3 and controlled by the output voltage V_2.The implemented threshold voltage has been the same as the assumed voltage for real diodes, i.e., 0.65V.

For the desired amplitude V_m, resistance R_2 has been obtained using Eq. (21). Steady state waveform for the defined circuit has been generated by transient analysis. Harmonic distortion has been determined from the spectrum. Flat top window has been applied on the generated signal. This window is suitable for determining discrete harmonic components of periodic signals. The broad and flat frequency response of this window has low spectral resolution, but has the important advantage that discrete spectral lines of periodic signals can-not miss the flat part, so the spectral peak exactly corresponds to the amplitude of the harmonic component.

Table 1: Simulation results for $R_1 = 2 \cdot R_0 = 300$ kΩ

Input			Ideal diode		1N4148	
V_m/ V_F	V_m [V]	R_2 [kΩ]	V_m [V]	THD [dB]	V_m [V]	THD [dB]
2	1.3	98	1.11	-53.0	0.9	-60.0
3	1.95	175	1.76	-59.5	1.7	-70.2
5	3.25	225	2.7	-63.0	3.4	-76.0

Table 2: Simulation results for circuit without R_1

Input			Ideal diode		1N4148	
V_m/ V_F	V_m [V]	R_2 [kΩ]	V_m [V]	THD [dB]	V_m [V]	THD [dB]
2	1.3	58	1.25	-49.0	1.04	-53.3
3	1.95	87	1.8	-53.1	1.68	-59.1
5	3.25	112	2.8	-58.3	3.13	-66.1

Simulated THD for real pn diode in the limiting network is roughly 6 dB lower than for ideal diodes, which is a logical consequence of a smoother waveform $v_3(t)$. But these values are also too low when compared with theory, especially in the case when R_1 is missing.

4.2 Experimental results

The oscillator shown in Fig.3 has been used for experimental measurements to verify the possibility to build simple sinusoidal signal source for very low harmonic distortion. The same operational amplifier has been used as in simulations, namely TL071. It has turned out that the performance of the actual integrated circuit performed better than the spice subcircuit model provided by the manufacturer. The unity-gain bandwidth has proved to be higher, hence higher critical resistance has been used for the calculation of circuit elements in the amplitude stabilizing network. Harmonic components of the generated signal have been measured with spectrum analyzer HP 3589A.

The results are summarized in Table 3.

Table 3: Measured results

Expected value		$R_1 = 2 \times R_0 = 440$ kΩ			R_1 is not used		
V_m/ V_F	V_m [V]	R_2 [kΩ]	V_m [V]	THD [dB]	R_2 [kΩ]	V_m [V]	THD [dB]
2	1.3	160	0.78	-61.1	86	0.87	-56.6
3	1.95	270	1.15	-60.0	130	1.22	-58.4
5	3.25	330	1.6	-61.7	180	1.72	-60.8

There is quite a substantial difference between expected amplitude and measured values for the circuit that includes resistance R_1. The voltage threshold of real diodes is not well pronounced but a better estimate can be acquired from the results for the circuit without resistor R_1, specifically $V_F = 0.4$ V seems to be a better choice for this purpose. The derived theoretical results concerning the relationship between R_2 and V_m are in sufficiently good agreement in simulations in which ideal diodes have been used. This is not a surprise as the relations are derived by the same approximation.

Measured THD for QO without resistor R_1 is in Fig.8 compared with analytical results for the usual range of normalized amplitudes. This range is featured with good amplitude stability and fair THD. A good match between the THD in Eqs. (25) and (24) and measured distortion can be noted.

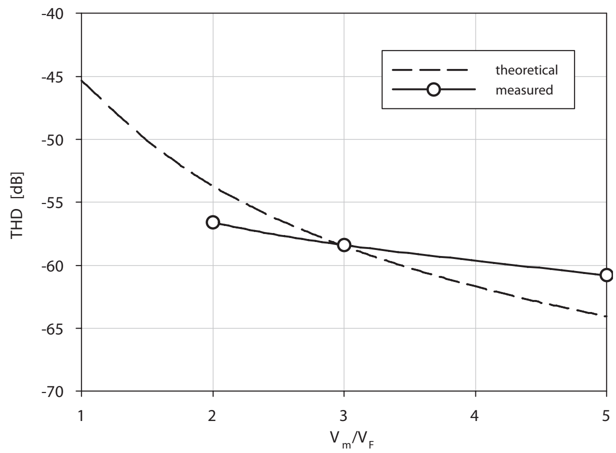

Figure 9: THD vs. normalized amplitude for QO without R_1

5 Conclusion

In this paper the performance of amplitude stabilizing network for QO has been analyzed. The derived relations offer a possibility to improve the performance of the circuit. The investigated oscillator is intended for low voltage operation, therefore only diodes are used for the non-linear network. Supply voltages and resistive dividers are usually used if stable larger amplitude is desired. In this case good stability of supply voltages is required. QO is a good and cheap source for sinusoidal voltages with moderate to low distortion for stand alone applications where digital solutions are not desired. The attained THD -61 dB has been measured at the output of the inverting integrator. THD of the quadrature signal V_2 is roughly 10 dB lower, because the third and higher harmonic components that are contained in V_1 are additionally attenuated. Low cost simple QO with THD = -70 dB or 0.03 % can be achieved with proper circuit design based on the presented work.

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