Tunable Oscillator Derived from Colpitts Structure with Simply Controllable Condition of Oscillation and Synthetic Inductor Based on Current Amplifier and Voltage Differencing Transconductance Amplifier

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Abstract

Our contribution discusses modification of well-known Colpitts LC oscillator to fully electronically controllable type utilizing adjustable current amplifier and inductance simulator based on voltage differencing transconductance amplifier (VDTA) instead of common bipolar transistor and passive metal coil. Implementation of current amplifier allows independent and direct control of oscillation condition, simplifies design (capacitors can have equal values because condition is not dependent on them) and used inductance equivalent based on VDTA allows linear control of frequency of oscillation and quadrature outputs with unchangeable amplitudes during the tuning process. Theory was verified by PSpice simulations and experimental measurement.

1. Introduction

LC sinusoidal oscillators are very important parts of applications in analog and mixed-signal electronics. However, presence of metal coil(s) and complicated tuning (only change of external passive elements L and C is possible) are main drawbacks of such solutions. Electronic control is not possible easily in many cases. Fortunately, there are many various active elements that allow direct electronic control of its parameter(s) and also control of features of particular application, [1] for instance. The most well-known and frequently used methods are: current gain control [2-4], control of intrinsic resistance of current input [5-7] and transconductance control [8].

In recent years, many advanced methods of control in frame of one active device were developed. Active elements with interesting and useful features were introduced in works [7, 9-12]. Presented approaches combine more than one type of electronic control of parameter in frame of one active device. For example, Minaei et al. [9] utilized electronic control of current gain and intrinsic resistance of current input terminal in frame of current conveyor of second generation [13-14]. Similar approaches of control were used in construction of special type of current feedback amplifier in [10-11]. Similarly de Marcellis et al. [12] developed current conveyor that offers possibility of current and voltage gain control.

Our effort in this contribution is focused on enlargement of controllable (electronically) possibilities of the known Colpitts LC circuit/oscillator [15] by using some attractive and popular methods of control (transconductance and current gain control) in circuit synthesis.

We intend to replace metal coil by synthetic equivalent. Operational transconductance amplifiers (OTAs) seem to be the best and the simplest choice for these purposes. Method of impedance conversion based on OTAs was discussed in tutorial [8] in detail. Active element including two OTAs is very useful and its definition was introduced in [1]. Voltage differencing transconductance amplifier (VDTA) [1] is quite new active element and it is very useful for our intentions in synthetic inductance equivalent. This element received attention in some applications that were discussed in recent works (active filters [16-18], oscillators [19-20]). We can also use knowledge of application of VDTA as inductance simulator presented in [21].

2. Discussion of Intended Modification

Small-signal linear model of Colpitts circuit [15] is shown in Fig. 1. Current transfer of bipolar transistor between emitter and collector is given by its current gain (h_{FE}, h_{21}, β) in such configuration (common base [22]) and is equal to 1 for very high β [22] ($\beta \cong i_{collector}/i_{base}$). Transistor in such configuration behaves as current follower (emitter is low-impedance input and collector serves as output with high-impedance character). Therefore, we can replace bipolar transistor in Fig. 1 by simple current follower and obtain following characteristic equation:

$$s^3 + \frac{C_1 + C_2}{LC_1C_2}s + \frac{1}{R_i LC_1C_2} = 0, \quad (1)$$

where R_i is an input resistance of the current follower. Equation (1) is not characteristic equation of an oscillator. However, careful study of the full symbolical representation (general current gain instead of unity gain) reveals that adjustable current amplifier instead of the current follower can bring this circuit to boundary of stability (oscillation). We suppose replacement of the current follower by the adjustable current amplifier in further explanation (R_i is now input resistance of the current amplifier). Modified circuit has following characteristic equation:

$$s^3 + \left(\frac{1-B}{R_i C_1}\right)s^2 + \frac{C_1 + C_2}{LC_1C_2}s + \frac{1}{R_i LC_1C_2} = 0, \tag{2}$$

where B is controllable current gain of the current amplifier which is now included in the circuit structure. Obtained characteristic equation (2) represents oscillator with favorable condition of oscillation (CO) in form:

$$B = 1 - \frac{C_1}{C_1 + C_2}. \tag{3}$$

In case of equal capacitors ($C_1 = C_2 = C$), we can drive CO easily by continuous change of B. Theoretical requirement for gain sufficient for fulfillment of CO is given by $B = 0.5$ (in case of $C_1 = C_2 = C$). The circuit obtained by discussed modification is shown in Fig. 2.

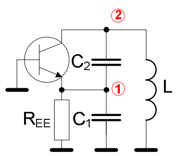

Fig. 1. Small-signal linear model of Colpitts circuit (adapted from circuit presented in [15])

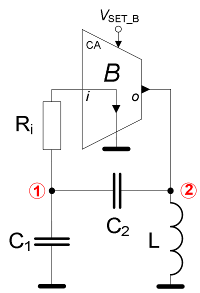

Fig. 2. Modification of the oscillator employing controlled current amplifier

Frequency of oscillation (FO) in above discussed case has typical form (for LC oscillators):

$$\omega_0 = \frac{1}{\sqrt{L\left(\frac{C_1C_2}{C_1 + C_2}\right)}}. \tag{4}$$

FO and CO is not dependent on input resistance (R_i in our case) of used current amplifier in ideal case (other sources of non-idealities are neglected) which is important in case of this type of the oscillator.

3. Fully Electronically Controllable Solution Employing Current Amplifier and VDTA

A classical wire inductor is not very popular circuit component due to its dimension, weight, dependence on mechanical influences (vibration, change of shape, change of coil distance, requirements for mechanical fixing, etc.). We selected known and popular method how to create synthetic equivalent of inductance [8] consisting of two OTAs [1], [8] that were also used in our case (in VDTA implementation [21]). The VDTA consists of two transconductance sections with auxiliary z terminal. Detailed principle of VDTA was discussed for example in [1], [20]. Resulting circuit is presented in Fig. 3 and FO is now:

$$\omega_0 = \frac{1}{\sqrt{\left(\frac{C_1C_2}{C_1 + C_2}\right)\frac{C_3}{g_{m1}g_{m2}}}}, \tag{5}$$

which leads to

$$\omega_0 = \frac{g_m}{C}\sqrt{2}, \tag{6}$$

if equality of all capacitors ($C_1 = C_2 = C_3 = C$) and both transconductances ($g_{m1} = g_{m2} = g_m$) is ensured. Used solution allows favorable features of proposed modification i.e. simply electronic and linear control of FO by simultaneous change of both transconductances.

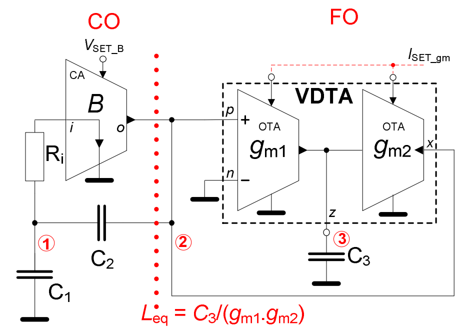

Fig. 3. Oscillator employing synthetic inductor

Relation between nodal voltages V_2 and V_3 has form:

$$\frac{V_3}{V_2} = \frac{g_{m1}}{sC_3}. \tag{7}$$

Substitution of $s = j\omega_0$ gives:

$$\frac{V_3}{V_2} = \frac{1}{j}\sqrt{\left(\frac{C_1C_2}{C_1 + C_2}\right)\frac{1}{C_3}\frac{g_{m1}}{g_{m2}}}, \tag{8}$$

that reduces to (supposing above discussed equality of capacitors and transconductances) numerical relation between generated amplitudes and phases:

$$\frac{V_3}{V_2} = \frac{\sqrt{2}}{2}\exp\left(\frac{\pi}{2}j\right), \tag{9}$$

which proves that phase shift between both produced signals is 90 degrees and amplitudes are constant during the tuning process (independent on FO).

Relation between V_1 and V_2 is given by:

$$\frac{V_1}{V_2} = \frac{sC_2R_i}{1 + s(C_1 + C_2)R_i}. \tag{10}$$

Supposing simplification (equal capacitors and transconductances) and mathematical arrangements we reached (at frequency of oscillation):

$$\frac{V_1}{V_2} = \sqrt{\frac{\left(g_mR_i\sqrt{2}\right)^2}{1 + \left(g_mR_i2\sqrt{2}\right)^2}}\exp\left[\arctan\left(\frac{1}{g_mR_i2\sqrt{2}}\right)j\right]. \tag{11}$$

Amplitude of V_1 is approximately 1/2 of V_2 for $g_m.R_i = 1$ and higher values. Amplitude V_1 of steady state oscillations changes (decreases) significantly for product $g_m.R_i < 1$ as we can see from (11). Fortunately, high $g_m.R_i$ ensures almost unchangeable level of V_1 also if oscillator is tuned (products in numerator and denominator of (11) are almost equal and higher than 1 in denominator). However, this node (V_1) is not suitable as output of multiphase type because phase shift between V_2 and therefore also V_3 changes if FO is tuned. Nodes 2 and 3 are useful to obtain quadrature phase shift with constant amplitude level during the tuning process.

4. Simulation and Measurement Results

The oscillator in Fig. 3 was simulated in OrCAD (PSpice). Completed circuit (Fig. 4) includes low cost automatic amplitude gain control circuit (AGC). The time constant of the AGC (R_f, C_f) was chosen adequately to speed of processed signal (stabilization of oscillations takes many times longer time than period of observed signal). In practice, several times higher value of C_f should be used (minimization of small fluctuances) but amplitude stabilization is sufficient after tens-hundreds of milliseconds. The PSpice macromodels of commercially available devices were used in presented results. Current amplifier (CA) was realized by current-mode multiplier EL2082 [23], VDTA section by two diamond transistors (DTs, OPA660/860 [24]) and low-cost opamps TL072 [25] were utilized in AGC.

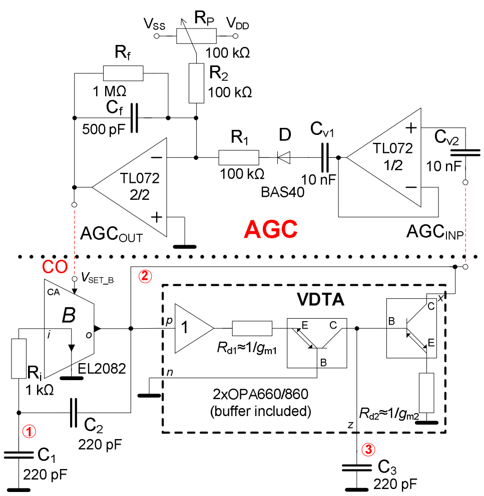

Fig. 4. Complete oscillator scheme – behavioral representation

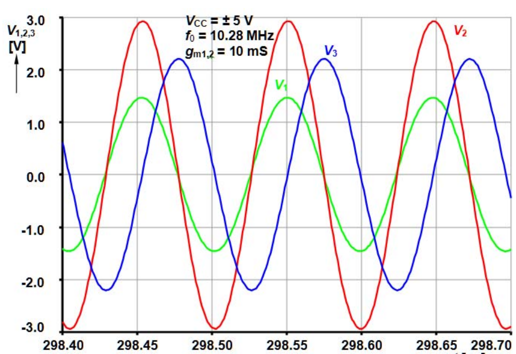

Fig. 5. Transient responses of voltages in nodes 1-3

Parameters of the design are following: $C_1 = C_2 = C_3 = C =$ 220 pF, $R_i = 1$ kΩ (external 910 Ω and internal 95 Ω [23]), $g_{m1,2}$ ($g_{m1} = g_{m2} = g_m$) adjustable from 0.3 to 10 mS. The rest of the parameters (AGC) is available in Fig. 4. Supply voltage was ± 5 V. The simulation results for stable FO are depicted in Fig. 5 (detail on all transient responses) and Fig. 6 (FFT spectrum).

Theoretical value of the frequency of oscillation 10.23 MHz for $g_{m1,2} = 10$ mS was obtained. Value 10.28 MHz was provided by simulations. Suppression of higher harmonic components was more than 40 dB (Fig. 6).

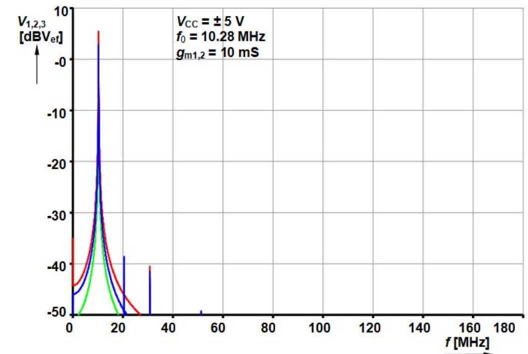

Fig. 6. FFT spectrum of generated signals

Simulations were focused mainly on verification and revision of tunable features of proposed modification. Figure 7 shows tuning in all available nodes (V_1, V_2, V_3) for three discrete values of $g_{m1,2}$ (0.5 mS, 1 mS, 10 mS). Obtained FO was 0.518, 0.99 and 10.28 MHz and signal levels reached values (peak-to-peak) of several volts. The relation between voltages quite corresponds to the theoretical prediction (9), (11).

The FO adjusting was tested in range from 0.34 MHz to 10.28 MHz and compared to the theoretical expectations (Fig. 8) that are given by equation (6). The amplitude level of output oscillations is quite stable for FO above 1 MHz to 10 MHz (see Fig. 9). Preliminary estimation of total harmonic distortion (THD) is shown in Fig. 10 for all signals (between 1 MHz and 10 MHz is maximally 1 %). Further improvements of THD are available by careful setting of the AGC but it is relatively time demanding process in transient simulations. Measurement of the real circuit representation (Fig. 4) allows fast and comfortable setting of the AGC if oscillator is under operation.

Automatic control of CO provided by the AGC system (shown in Fig. 4) is really necessary because it is not possible to operate with fixed gain $B = 0.5$ as we derived from ideal condition (3). Tuning of the frequency of oscillation causes gain changes in the loop that need to be compensated (this problem is discussed in more detail in [26]). Discontinuation of oscillation, high THD or even saturation of generated signals to supply corners is possible when operated without precise auto-compensating AGC systems. Changes of DC voltage, that control CO (current gain B), are noted in Fig. 11 (current gain is practically equal to V_{SET_B} in indicated range in accordance to [23]).

In order to confirm the theory, the functionality of the circuit in Fig. 4 was experimentally tested in solder-less breadboard ($f_0 = 1.107$ MHz for $g_{m1,2} = 1.3$ mS). Preliminary results are shown in Fig. 12. Detailed tests and verifications (Fig. 7 - Fig. 11) of this oscillator are assumed as our future tasks.

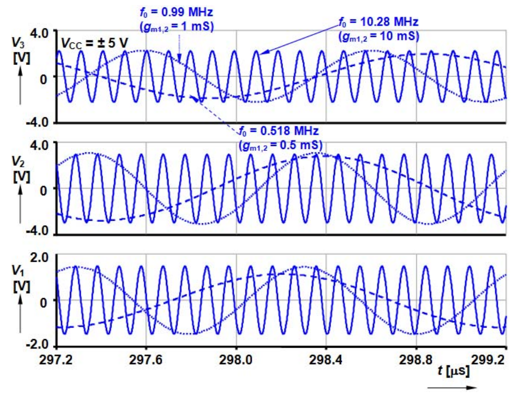

Fig. 7. Transient responses (V_1, V_2, V_3) in all nodes (example of tuning for three discrete frequencies - 0.52, 0.99, 10.28 MHz)

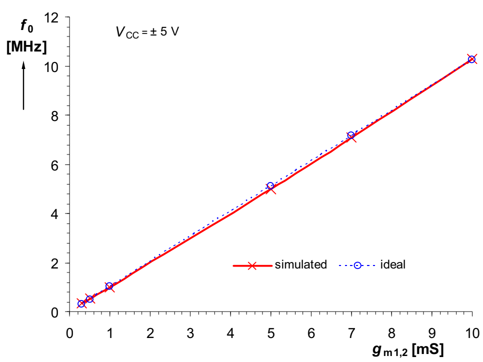

Fig. 8. Dependence of frequency of oscillation on transconductance control in VDTA based synthetic inductor

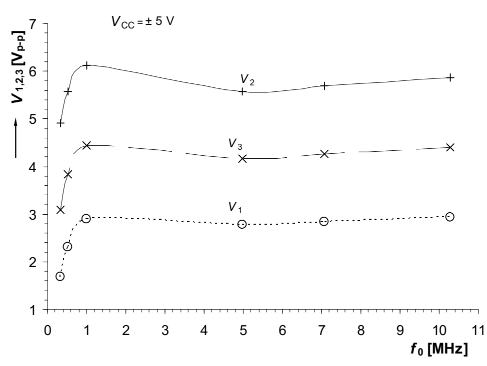

Fig. 9. Dependence of generated signals on frequency of oscillation

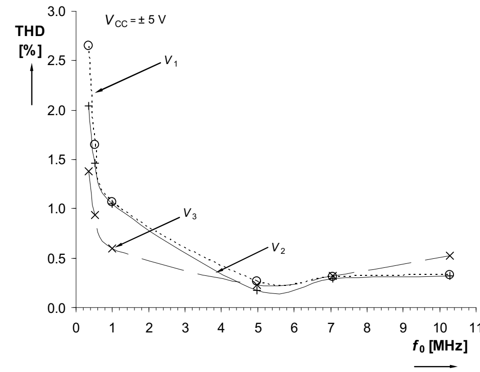

Fig. 10. Dependence of THD on oscillation frequency

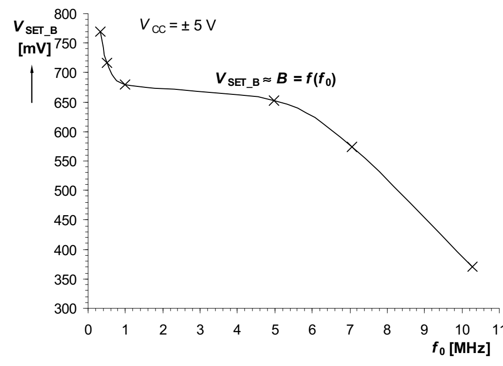

Fig. 11. Response of AGC on FO tuning

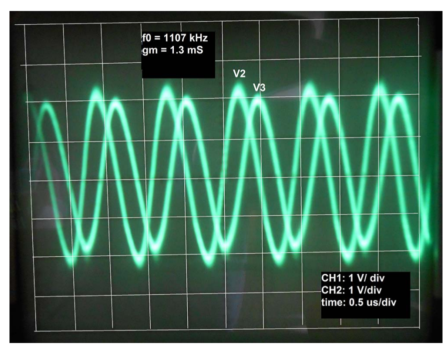

Fig. 12. Measured transient responses of voltages in nodes 2-3

5. Conclusion

Introduced derivation based on Colpitts circuit offers interesting features that are missing in classical version of Colpitts oscillator based on bipolar transistor and classical coil (inductance). These useful benefits can be summarized as: linear control of FO, low current gain required for fulfillment of CO (lower than 1, theoretical value given by (3) is 0.5), simple AGC control by B, quadrature outputs and unchangeable quadrature amplitudes during the tuning process. One floating capacitor (but it comes from principle of the Colpitts oscillator) is disadvantage of this solution. This contribution shows preliminary results. We suppose additional measurements with behavioral models, proposal of CMOS realization, study of parasitic influences and impact of R_i on levels and THD in the future.

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