

Electronically Controllable Fully-Uncoupled Explicit Current-Mode Quadrature Oscillator Using VDTAs and Grounded Capacitors

Dinesh Prasad^{1*}, Mayank Srivastava², Data Ram Bhaskar¹

¹Department of Electronics and Communication Engineering, Faculty of Engineering and Technology, Jamia Millia Islamia, New Delhi, India

²Department of Electronics and Communication Engineering, Amity School of Engineering and Technology, Amity University, Noida, India

Email: *dprasad@jmi.ac.in, mayank2780@gmail.com, dbhaskar@jmi.ac.in

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ABSTRACT

An electronically controllable fully uncoupled explicit current-mode quadrature oscillator employing Voltage Differencing Transconductance Amplifiers (VDTAs) as active elements has been presented. The proposed configuration employs two VDTAs along with grounded capacitors and offers the following advantageous features 1) fully and electronically independent control of condition of oscillation (CO) and frequency of oscillation (FO); 2) explicit current-mode quadrature oscillations; and 3) low active and passive sensitivities. The workability of proposed configuration has been demonstrated by PSPICE simulations with TSMC CMOS 0.18 μm process parameters.

Keywords: VDTA; Quadrature Oscillator; Current-Mode Circuits

1. Introduction

Among various kinds of oscillators, the quadrature oscillators (QO) are widely used because they can provide two sinusoids with $\pi/2$ phase difference, for example, in telecommunication systems, for quadrature mixtures and single-sideband generators or for measurement purposes in vector generators or selective voltmeters [1,2]. Therefore the QO play an important role in many communication systems, instrumentation systems and signal processing see [3-11]. Recently, a CMOS realization of VDTA and its applications as 1) RF filter and 2) double tuned amplifier have been presented in [12]. In [13], an electronically controllable explicit current-output sinusoidal oscillator has been reported. Another application as a single input five output voltage-mode universal filter using VDTAs has been presented in [14]. The purpose of this communication is to introduce a new electronically controllable fully uncoupled explicit current-mode quadrature oscillator using two VDTAs and two grounded capacitors. The proposed configuration provides the advantageous features of: 1) completely and electronically independent control of condition of oscillation (CO) and

frequency of oscillation (FO); 2) explicit current-mode quadrature oscillations; and 3) low active and passive sensitivities. The workability of proposed configuration has been verified using SPICE simulation with TSMC CMOS 0.18 μm process parameters.

2. The Proposed Configurations

The symbolic notation of the VDTA is shown in **Figure 1**, where V_P and V_N are input terminals and Z , X^+ and X^- are output terminals. All terminals of VDTA exhibit high impedance values [12]. The VDTA can be described by the following set of equations:

$$\begin{bmatrix} I_Z \\ I_{X^+} \\ I_{X^-} \end{bmatrix} = \begin{bmatrix} g_{m1} & -g_{m1} & 0 \\ 0 & 0 & g_{m2} \\ 0 & 0 & -g_{m2} \end{bmatrix} \begin{bmatrix} V_{V_P} \\ V_{V_N} \\ V_Z \end{bmatrix} \quad (1)$$

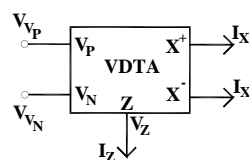


Figure 1. The symbolic notation of VDTA.

*Corresponding author.

The proposed configuration is shown in **Figure 2**.

Circuit analysis of **Figure 2** gives the characteristic equation (CE) as:

$$s^2 + s \frac{1}{C_2} (g_{m_4} - g_{m_3}) + \frac{g_{m_1} g_{m_2}}{C_1 C_2} = 0 \quad (2)$$

From Equation (2), the CO and FO can be expressed as CO:

$$(g_{m_4} - g_{m_3}) \leq 0 \quad (3)$$

and

FO:

$$\omega_0 = \sqrt{\frac{g_{m_1} g_{m_2}}{C_1 C_2}} \quad (4)$$

Therefore, it is seen from Equations (3) and (4) that the CO and FO are completely uncoupled and electronically tunable as g_{m_i} ; $i = 1 - 4$ are controlled by bias currents.

The current transfer functions obtained from **Figure 2** are given by:

$$\frac{I_{o2}(s)}{I_{o1}(s)} = \frac{g_{m_4}}{g_{m_1} g_{m_2} s C_1} \quad (5)$$

For sinusoidal steady state, Equations (5) becomes

FO:

$$\omega_0 = \sqrt{\frac{\left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)}{(C_1 C_2 + 2C_1 C_p + C_1 C_z + 2C_1 C_x + 2C_p C_z + C_2 C_z + C_z^2 + 2C_x C_z)}} \quad (8)$$

then the active and passive sensitivities of ω_0 can be found as:

$$S_{R_x}^{\omega_0} = \frac{-1}{R_x R_z \left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)}, \quad S_{R_z}^{\omega_0} = \frac{\left(-\frac{2}{R_p R_z} - \frac{2}{R_z^2} - \frac{2}{R_x R_z} - \frac{g_{m_4}}{R_z} + \frac{g_{m_3}}{R_z} \right)}{2 \left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)},$$

$$S_{R_p}^{\omega_0} = \frac{-1}{R_p R_z \left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)}, \quad S_{g_{m_1}}^{\omega_0} = \frac{g_{m_1} g_{m_2}}{2 \left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)} = S_{g_{m_2}}^{\omega_0},$$

$$S_{g_{m_3}}^{\omega_0} = \frac{g_{m_3}}{2R_z^2 \left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)}, \quad S_{g_{m_4}}^{\omega_0} = \frac{-g_{m_4}}{2R_z^2 \left(\frac{2}{R_p R_z} + \frac{1}{R_z^2} + \frac{2}{R_x R_z} + \frac{g_{m_4}}{R_z} - \frac{g_{m_3}}{R_z} + g_{m_1} g_{m_2} \right)},$$

$$S_{C_1}^{\omega_0} = \frac{-C_1 (C_2 + 2C_p + C_z + 2C_x)}{2(C_1 C_2 + 2C_1 C_p + C_1 C_z + 2C_1 C_x + 2C_p C_z + C_2 C_z + C_z^2 + 2C_x C_z)},$$

$$S_{C_2}^{\omega_0} = \frac{-C_2 (C_1 + C_z)}{2(C_1 C_2 + 2C_1 C_p + C_1 C_z + 2C_1 C_x + 2C_p C_z + C_2 C_z + C_z^2 + 2C_x C_z)},$$

$$\frac{I_{o2}(j\omega)}{I_{o1}(j\omega)} = \frac{g_{m_4}}{\omega C_1 g_{m_1} g_{m_2}} e^{-j90^\circ} \quad (6)$$

Thus, the phase difference ϕ between (I_{o2} and I_{o1}) is -90° . Hence, the currents (I_{o2} and I_{o1}) are in the quadrature form.

3. Parasitic Effects and Sensitivity Analysis

By considering the various VDTA non-ideal parameters like the finite P -terminal parasitic impedance consisting of a resistance R_p in parallel with capacitance C_p , the finite N -terminal parasitic impedance consisting of a resistance R_N in parallel with capacitance C_N , the finite X -terminal parasitic impedance consisting of a resistance R_X in parallel with capacitance C_X and the parasitic impedance at the Z -terminal consisting of a resistance R_Z in parallel with capacitance C_Z then the expression of CO and FO including the influence of parasitic are given by:

CO:

$$\left\{ \frac{2C_1}{R_p} + \frac{C_1}{R_z} + \frac{2C_1}{R_x} + \frac{2C_p}{R_z} + \frac{C_2}{R_z} + \frac{C_z}{R_z} + \frac{2C_x}{R_z} + \frac{2C_z}{R_p} + \frac{C_z}{R_z} + \frac{2C_z}{R_x} + C_1 (g_{m_4} - g_{m_3}) + C_z (g_{m_4} - g_{m_3}) \right\} \leq 0 \quad (7)$$

$$S_{C_x}^{\omega_0} = \frac{-2C_x(C_1 + C_z)}{2(C_1C_2 + 2C_1C_p + C_1C_z + 2C_1C_x + 2C_pC_z + C_2C_z + C_z^2 + 2C_xC_z)},$$

$$S_{C_p}^{\omega_0} = \frac{-2C_p(C_1 + C_z)}{2(C_1C_2 + 2C_1C_p + C_1C_z + 2C_1C_x + 2C_pC_z + C_2C_z + C_z^2 + 2C_xC_z)},$$

$$S_{C_z}^{\omega_0} = \frac{-C_z(C_1 + 2C_p + C_2 + 2C_z + 2C_x)}{2(C_1C_2 + 2C_1C_p + C_1C_z + 2C_1C_x + 2C_pC_z + C_2C_z + C_z^2 + 2C_xC_z)} \quad (9)$$

For $C_1 = C_2 = 0.5$ nF, $R_p = R_z = \infty$, $C_p = C_x = C_z = 0.15$ pF, $g_{m_1} = g_{m_2} = g_{m_3} = 1.5913$ mA/V and $g_{m_4} = 1.7916$ mA/V, the sensitivities are found to be 0, 0, 0, 0.5, 0, 0, -0.899, -0.499, -2.9e-4, -2.9e-4, -1.64e-3 for Equation (9). Thus, all the active and passive sensitivities of ω_0 with respect to each active and passive elements are low.

4. Simulation Results

To verify the theoretical analysis, the proposed ECMSO was simulated using CMOS VDTA from [12]. Power supply voltages were taken as $V_{DD} = -V_{SS} = 0.9$ V and $I_{B1} = I_{B2} = I_{B3} = I_{B4} = 2$ mA (for VDTA₁) and $I_{B1} = I_{B2} = 2$ mA, $I_{B3} = I_{B4} = 3.687$ mA (for VDTA₂) biasing currents are used. The transistor aspect ratios are same as in [12]. The passive elements of the configuration were selected as $C_1 = C_2 = 0.5$ nF. The transconductances of VDTA were controlled by bias currents. PSPICE generated output waveforms indicating transient and steady state responses are shown in **Figures 3(a)** and **(b)** respectively. These results, thus, confirm the validity of the proposed configuration. The total harmonic distortion (THD) of the proposed oscillator is found to be 3.00% (**Figure 4**). From **Figure 5** it is clear that the two currents are in quadrature.

Figure 5 shows that the two currents are in quadrature and the measured value of phase shift between two waveforms is = -89.98°.

5. Concluding Remarks

In this paper, an explicit current-mode quadrature oscillator using VDTAs has been presented. The presented circuit employs two VDTAs and two grounded capaci-

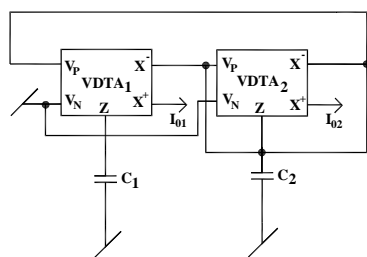
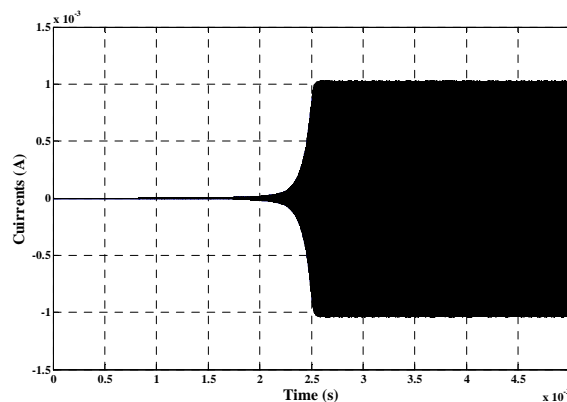
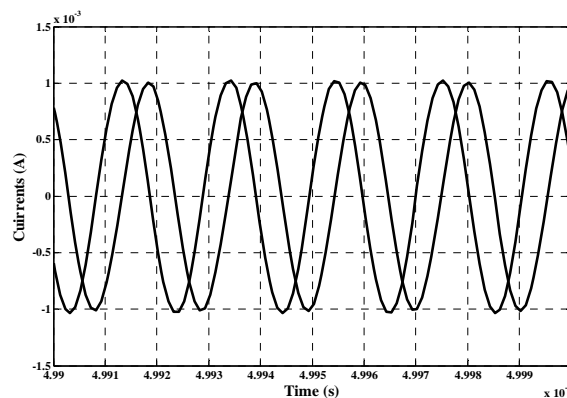


Figure 2. The proposed configuration.



(a)



(b)

Figure 3. (a) Transient response waveform; (b) Steady state response of the quadrature outputs.

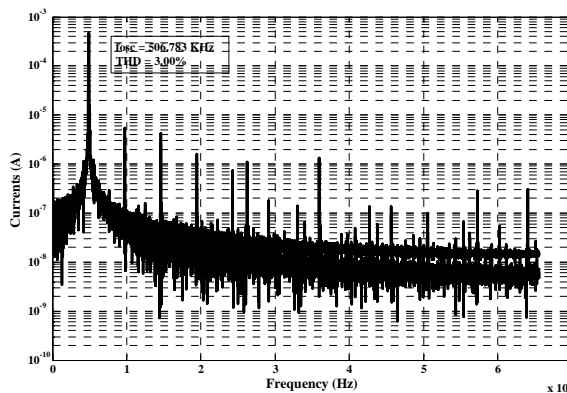


Figure 4. Simulation result of the output spectrum.

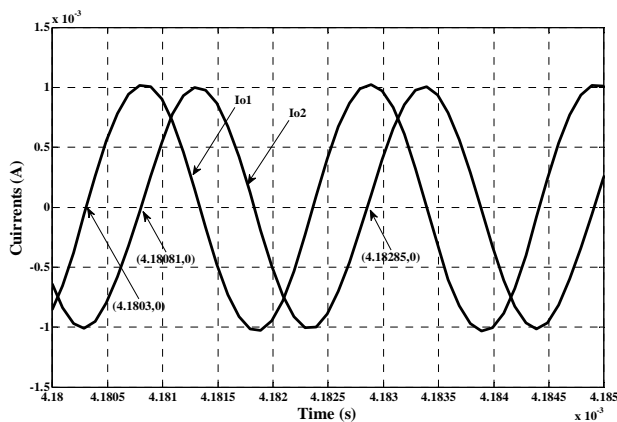


Figure 5. Steady state response of the quadrature outputs of I_{o2} and I_{o1} .

tors. The CO and FO of the proposed quadrature oscillator has the advantage of fully and electronically independent controllability. The proposed explicit current-mode quadrature oscillator also provides low active and passive sensitivities. The workability of proposed configuration has been verified using SPICE simulation.

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