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New Realization of Quadrature Oscillator using OTRA

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ABSTRACT

In this paper a new, operational transresistance amplifier (OTRA) based, third order quadrature oscillator (QO) is presented. The proposed structure forms a closed loop using a high pass filter and differentiator. All the resistors employed in the circuit can be implemented using matched transistors operating in linear region thereby making the proposed structure fully integrated and electronically tunable. The effect of non-idealities of OTRA has been analyzed which suggests that for high frequency applications self-compensation can be used. Workability of the proposed QO is verified through SPICE simulations using 0.18μm AGILENT CMOS process parameters. Total harmonic distortion (THD) for the proposed QO is found to be less than 2.5%. The sensitivity, phasenoise analysis is also discussed for the proposed structure.

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1. INTRODUCTION

In last few decades current-mode (CM) processing has evolved as a promising design technique to provide efficient solutions to circuit design problems. This evolution has resulted in emergence of numerous CM analog building blocks [1].The operational trans resistance amplifier (OTRA) is one among these blocks. It is a high gain current input, voltage output amplifier [2] and uses current feedback technique which makes its bandwidth almost independent of the closed loop gain. Additionally it is free from the effect of parasitic capacitances at the input due to virtually internally grounded input terminals [2] and hence non-ideality problem is less in circuits implemented using OTRA.

Quadrature oscillators (QO) are an important class of circuits and find wide application in communication, power electronics and instrumentation. This has led a consistent research effort towards second order QO design using wide variety of active blocks, as is evident from vast literature available [2-14] .It is well known that higher order networks, provide better accuracy, frequency response and distortion performance [15-17] as compared to lower order circuits. Owing to this in recent past few third order QO designs [15], [16], [18-32] have been reported. Realizing sinusoidal oscillator using closed loop with positive feedback is a well-established method. A careful observation suggests that all the reported third order QO designs are based on forming closed loop using combinations of lossy and/or lossless integrators. In this paper a new OTRA based third order QO is proposed that adapts the scheme of using second order high pass filter and a differentiator in a feedback loop [32].A comparative statement of the proposed structure with previously reported QO circuits is recorded in Table 1. It may be observed from the table that the available topologies-presented in [19] are realized using op-amps however, the constant gain-bandwidth product and lower slew rate of the op-amps limit their high frequency operations. Additionally these circuits use more number of active components as compared to proposed circuit

a. Lack electronic tunability [15], [19], [20], [24]
b. Use mix of active blocks such as DDCC and OTA [28],CCCDTA and OTA [29],CCCCTA and OTA [30]

c. Provide voltage output at high impedance [15], [16], [20], [22], [24], [25], [28], [29], [31] making a buffer necessary to drive the voltage input circuits
d. Provide current output [18], [20-22]which need to be converted to voltage for circuits requiring voltage inputs and would considerably increase the component count

Table 1. A comparative statement of the proposed structure with previously reported QO circuits

Ref.	Active blocks	No of active blocks	Close loop formation scheme	Passive elements C + R	Electronic tuning	Output Type	Output Impedance	Floating passive Compnents
[15]	CCII	3	Fig. 1, two lossy and one lossless integrators	3 +5	N	Voltage	High	N
		3	Fig.2, -do-	5+3	N	Voltage	High	N
		3	Fig. 3, Intuition	3+5	N	Voltage	High	N
[16]	OTA	3	Fig. 7, two lossy and one lossless integrators	3+0	Y	Voltage	High	N
		4	Fig. 9, -do-	3+1	Y	Voltage	High	N
[18]	CDTA	3	-do-	3+0	Y	Current	High	N
[19]	Op-amp	3	-do-	3+5	N	Voltage	Low	Y
		3	-do-	5+3	N	Voltage	Low	Y
[20]	CCII	2	A second order low pass filter and an integrator	3+3	N	Both	High	Y
[21]	CCCII	4	-do-	3+0	Y	Current	High	N
[22]	CCCII	3	one lossy and two lossless integrators	3+0	Y	Both	High	N
[23]	CCCII	4	Cascade of three LPFs with gained feedback	3+1	Y	Voltage	High	N
[24]	DVCC	4	A second order LPF and an integrator	3+3	N	Voltage	High	Y
[25]	CDTA	3	-do-	3+0	Y	Both	High	N
[26]	OTRA	2	-do-	3+4	Y	Voltage	Low	R1+C
[27]	CCCCTA	2	-do-	3+0	Y	Current	High	N
[28]	DDCC and OTA	3	-do-	3+1	Y	Voltage	High	N
[29]	CCCDTA and OTA	2	-do-	3+0	Y	Both	High	N
[30]	CCCCTA and OTA	2	-do-	3+0	Y	Current	High	N
[31]	DVCCTA	2	Fig. 5, HPF and differentiator	3+1	Y	Both	High	N
		2	Fig. 6, -do-	3+2	Y	Both	High	N
[32]	OTRA	2	Fig.2,LPF and integrator	3+3	N	Voltage	Low	R2,R3,C1,C2
		2	Fig.3,HPF and differentiator	3+3	N	Voltage	Low	R1,R2,C2,C3
Proposed	OTRA	2	Fig.2HPF and differentiator	3+4	Y	Voltage	Low	R1+C

Above discussion suggests that OTRA based QO is most suitable choice for voltage output configurations.Rest of the paper is as organized as follows: in section 2 proposed circuit is described followed by Effect of nonideality of OTRA is dealt in section 3. Section 4 explains the MOS-C implementation details of proposed structure. Proposed structured is verified experietally by contructing OTRA using offshelf IC's AD844 [34], phase noise analysis using the method discussed in [35], [36] is presented in section 5. Sensitivity analysis are discussed in section 6. The simulation and experimental results are presented in section 7 and paper is concluded in section 8.

2. PROPOSED CIRCUIT

The circuit symbol of OTRA is shown in Figure 1 and its port characteristics are given by

$$\begin{bmatrix} V_p \\ V_n \\ V_0 \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ R_m & -R_m & 0 \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ I_0 \end{bmatrix} \quad (1)$$

The output voltage is the difference of two input currents multiplied by trans-resistance gain (R_m). Ideally the trans-resistance gain R_m approaches infinity and therefore the OTRA must be used in a negative feedback configuration [2]. The proposed QO topology is shown in Figure 2. It uses an OTRA based second order high pass filter [32] (C_1=C_3) and an inverting differentiator in the feedback forming a closed loop, which results in a third order characteristic Equation given by

$$s^3C_1C_3R_DC_D + s^2C_1C_2 + s\left(\frac{C_1}{R_2} + \frac{C_2}{R_1}\right) + \frac{1}{R_1R_2} = 0 \tag{2}$$

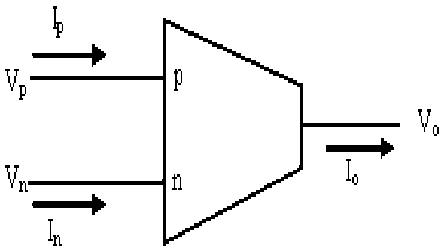


Figure 1. OTRA circuit

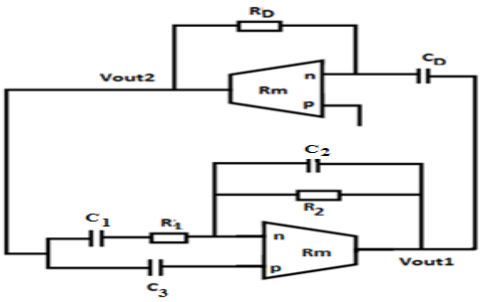


Figure 2. Proposed circuit

Assuming C_2=C_1 the condition of oscillation (CO) and frequency of oscillation (FO)

$$\text{F.O: } f = \frac{\sqrt{R_1 + R_2}}{2\pi\sqrt{C_3R_1R_2R_DC_D}} \tag{3}$$

$$\text{CO: } (R_1 + R_2)C_2 = R_DC_D \tag{4}$$

The FO can be adjusted to desired value through $R_{1,}$ R_2 and proper selection of resistor R_D would satisfy the CO.

3. NONIDEAL ANALYSIS

Ideally the transresistance gain R_m is assumed to approach infinity. However, practically R_m is a frequency dependent finite value. The output of the QO may deviate due to non-ideality of OTRA in practice. Considering a single pole model for the trans-resistance gain, R_m can be expressed as

$$R_m(s) = \left(\frac{R_o}{1 + \frac{s}{\omega_o}}\right) \tag{5}$$

Where R_O represents the dc transresistance gain. For high frequency applications the transresistance gain $R_m(s)$ reduces to

$$R_m(s) = \frac{1}{sC_P}, \qquad \text{where } C_P = \frac{1}{R_O\omega_O} \tag{6}$$

Taking this effect into account the characteristic Equation given by (2) modifies to

$$Ws^3 + Xs^2 + Ys + Z = 0 \tag{7}$$

Where the coefficients W, X, Y, and Z can be expressed as

$$W = CC_2C_PR_D + C_P^2CR_D + C^2C_DR_D$$

$$X = CC_2 + CC_P + CC_P\frac{R_D}{R_2} + C_PC_2\frac{R_D}{R_1} + C_P^2C\frac{R_D}{R_1} \quad ; Y = \frac{C}{R_2} + \frac{(C_2 + C_P)}{R_1} + \frac{R_DC_P}{R_1R_2} ; Z = \frac{1}{R_1R_2}$$

Due to parasitic effect the FO and CO also change and are given by (8) and (9) respectively

FO: $$f = \sqrt{\frac{\frac{R_DC_P}{R_1 + R_2} + \frac{C}{R_2} + \frac{C_2 + C_P}{R_1}}{C^2C_DR_D + CR_D(C_2 + C_P)}} \tag{8}$$

CO: $$(C_2 + C_P)R_2 + CR_1 + C_PR_D = \frac{R_1\,R_2\,R_D\{C(C_2 + C_P) + C^2C_D\}}{R_2(C + C_P)\left(R_1\,C + R_D\,C_P\right) + R_1R_D\,C_P\,C} \tag{9}$$

As the parasitic capacitance of the OTRA is very small, using approximation $(C_2 + C_p) \approx C$ the W, X, Y and Z coefficients can be simplified as

$$\begin{aligned} W &= CC_2C_PR_D + C_P^2CR_D + C^2C_DR_D \\ &= CR_D\left(C_P(C_2 + C_P) + CC_D\right) \\ &\approx CR_D\left(C_PC + CC_D\right) \approx C^2R_D\left(C_P + C_D\right) \approx C^2R_DC_D \end{aligned} \tag{10}$$

$$\begin{aligned} X &= CC_2 + CC_P + CC_P\frac{R_D}{R_2} + C_PC_2\frac{R_D}{R_1} + C_P^2C\frac{R_D}{R_1} \\ &= C(C_2 + C_P) + C_P(C_2 + C_P)\frac{R_D}{R_1} + \frac{R_D}{R_2}C_{\ P}C \quad \approx C^2 + CC_P\left(\frac{R_D}{R_1} + \frac{R_D}{R_2}\right) \\ &\approx C^2 \text{ as } CC_P \ll C^2 \end{aligned} \tag{11}$$

$$\begin{aligned} Y &= \frac{C}{R_2} + \frac{(C_2 + C_P)}{R_1} + \frac{R_DC_P}{R_1R_2} \approx Y = \frac{C}{R_2} + \frac{C}{R_1} + \frac{R_DC_P}{R_1R_2} \quad \text{as } C_P \ll C \\ &\approx \frac{C}{R_2} + \frac{C}{R_1} \qquad \text{and } Z = \frac{1}{R_1R_2} \end{aligned} \tag{12}$$

By substituting W, X, Y, Z from (10), (11), (12) in (7) the characteristic Equation and hence the FO and CO can be obtained which are same as given by (2), (3) and (4) respectively.

4. MOS-C IMPLEMENTATION

The differential input of OTRA allows the resistors connected to the input terminals of OTRA to be implemented using MOS transistors with complete non-linearity cancellation [33]. Each resistor implementation require two matched N MOS transistorsas shown in Figure 3.

Figure 4 shows a typical MOS based implementation of resistance connected at inverting terminal of OTRA where nodes X and Y need to be connected to inverting and non-inverting terminals of the OTRA respectively. The value of resistance so obtained is expressed as

$$R = \frac{1}{\mu_nC_{ox}(W/L)(V_a - V_b)} \tag{13}$$

Where μ, C_{OX} and W/L represent standard transistor parameters and V_a and V_b are the gate voltages. The MOS based implementation of the proposed circuit of Figure 2 is shown in Figure 4. The resistance value may be adjusted by appropriate choice of gate voltages thereby making oscillator parameters electronically tunable.

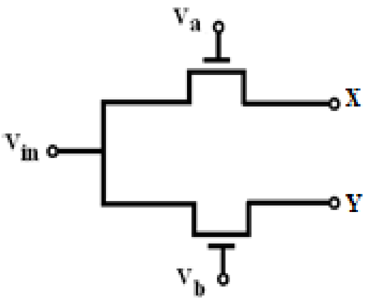


Figure 3. The MOS based resistor [33]

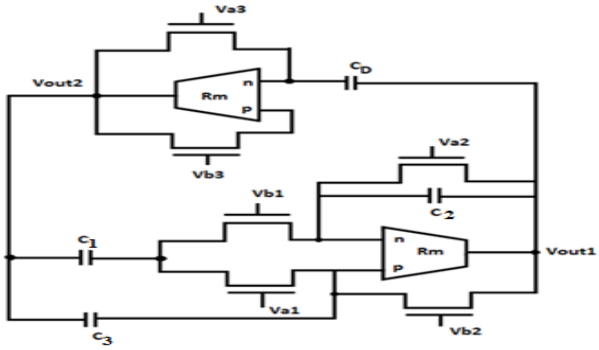


Figure 4. The MOS based implementation of QO Circuit

5. PHASE NOISE ANALYSIS

The random frequency fluctuations in a phase of a signal can be treated as a phase noise.To calculates the phase noise a procedure discussed in [35], [36] is adopted. The open loop transfer function H(s) of the oscillator circuit of Figure 2 is given by

$$H(s) = \frac{-s^3 C_1 C_3 C_D R_D R_1 R_2}{(1+sCR_1)(1+sCR_2)} \tag{14}$$

The H(s) given by (14) can also be expressed in terms of magnitude and phase as

$$H(j\omega) = A(\omega).e^{j\phi\omega} \tag{15}$$

From (15)

$$\frac{dH}{d\omega} = \left(\frac{dA}{d\omega} + \mathrm{jA}\frac{d\phi}{d\omega}\right) e^{j\phi\omega}$$

Substituting the CO and FO of the proposed oscillator in (14) the magnitude A(ω) can be written as

$$|A(\omega)| = \frac{\left(\frac{\omega}{\omega_o}\right)^2 \omega}{\sqrt{\left(-\frac{1}{C_D R_D}\left(\frac{\omega}{\omega_o}\right)^2 + \frac{1}{C(R_1+R_2)}\right)^2 + \omega^2}} \tag{16}$$

Determining $\left|\frac{dA}{d\omega}\right|$ from (16) results in $\left|\frac{dA}{d\omega}\right| = \frac{2}{\omega_o}$; $\angle H(\omega) = \phi = -Tan^{-1}\left(\frac{\frac{1}{\omega C_D R_D} - \frac{1}{C\omega\left(\frac{\omega}{\omega_o}\right)^2 (R_1+R_2)}}{\left(\frac{\omega}{\omega_o}\right)^{-2}}\right)$

(17)

Determining $\left|\frac{d\phi}{d\omega}\right|$ from (17) results in $\left|\frac{d\phi}{d\omega}\right| = \frac{2}{\omega_o^2 C_D R_D}$ (18)

From Equation (18) it is clear that Frequency stability of proposed quadrature oscillator decreases with increase of ω_o .

6. SENSITIVITY ANALYSIS

The sensitivity is an important performance criterion of any network. The sensitivity of FO (ω_o) with respect to a circuit parameters, say Y is given as

$$S_Y^{\omega_o} = \frac{\partial \omega_o}{\partial Y} \cdot \frac{Y}{\omega_o}$$

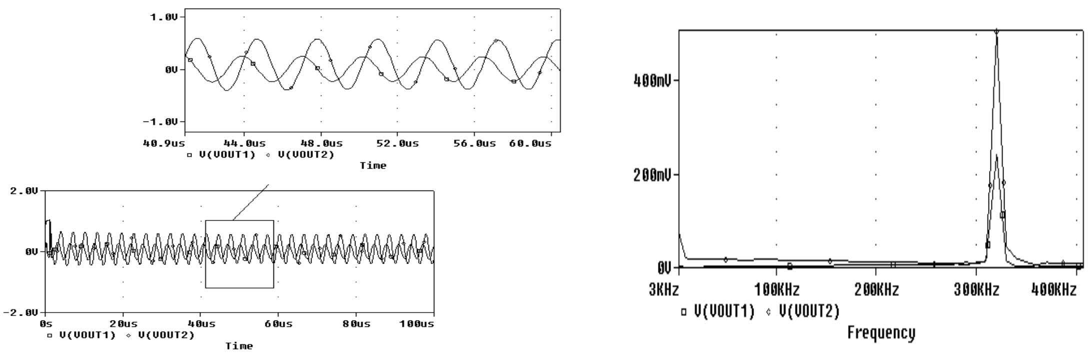
Using this definition, the sensitivity of FO (ω_o) for the circuit w.r.t R_1, R_2, C are given as

$$\left| S_{C_D}^{\omega_o} \right| = \left| S_C^{\omega_o} \right| = \frac{1}{2} ; \left| S_{R_1}^{\omega_o} \right| = \frac{R_2}{2(R_1 + R_2)} ; \left| S_{R_2}^{\omega_o} \right| = \frac{R_1}{2(R_1 + R_2)}$$

From the above Equations it is observed that all passive sensitivities for both the circuits are lower than unity in magnitude. It ensures that the sensitivity performance is good.

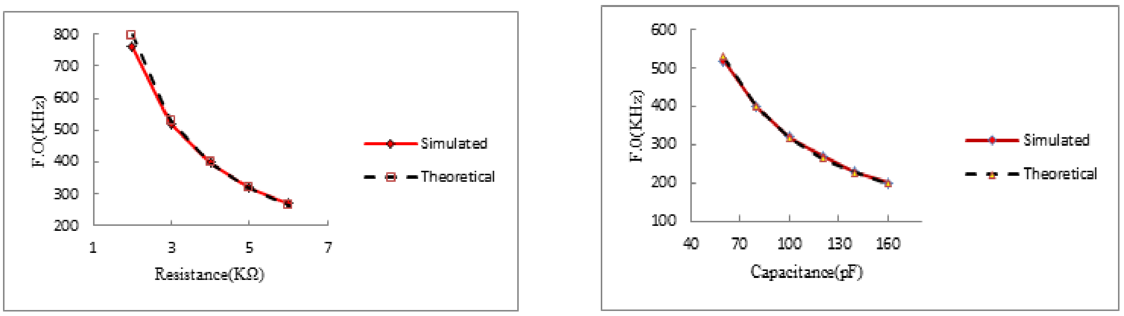
7. SIMULATION AND EXPERIMENTAL RESULT

The proposed QO is verified through simulations using the CMOS implementation of the OTRA [9]. The SPICE simulations are performed using 0.18µm CMOS process parameters provided by MOSIS (AGILENT). Supply voltages taken are ±1.5V. Component values are chosen as C_1=C_2=C_3=C_D=100pF and R_1=R_2=5KΩ, R_D=10KΩThe simulated FO was observed to be 320 KHz as against the calculated value of 318.47 KHz. The simulated transient output and corresponding frequency spectrum are shown in Figure 5(a) and 5(b) respectively.



(a) (b)

Figure 5. (a) Transient Output (b) Frequency spectrum of proposed QO circuit.

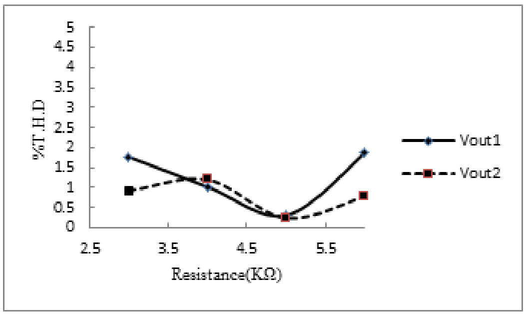


(a) (b)

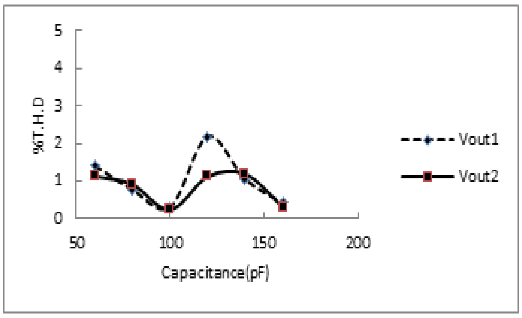
Figure.6. Frequency Tuning with (a) Resistance (b) with Capacitance for circuit

The FO of the proposed QO can be tuned through R or C variations, as suggested by (3).The FO tuning with R (varied from 3kΩ to 6kΩ) while keeping C fixed (100pF) is shown in Figure 6(a) whereas tuning with C (varied from 60pF to 140pF) with R fixed at 5kΩ is depicted in Figure 6(b). It may be observed that the simulated and theoretical values of FO are in close agreement.

The %THD variation with R and C is also studied and is depicted in Figure 7.The %THD variation with R (C = 100pF) for both Quadrature outputs, is recorded in Figure 7(a) and the largest value observed is 1.87%. Similarly Figure 7(b) shows %THD variation with C (R= 5kΩ) where in the maximum observed value is well within 2.5%. The phase error plots between V_{out1} and V_{out2} are drawn in Figure 8. Variation of phase error with resistance and capacitance are depicted in Fig.ure 8(a) and 8(b) respectively.

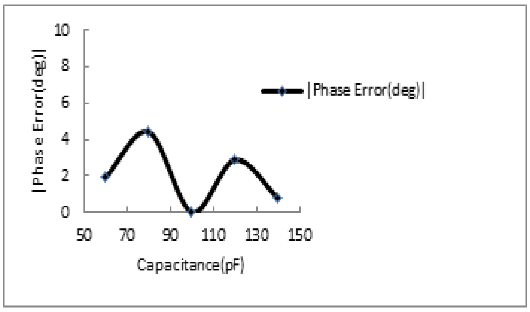


(a)

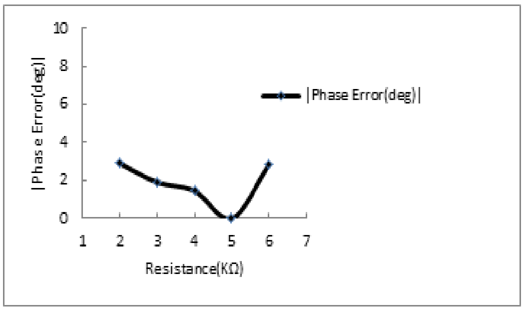


(b)

Figure 7. The % THD variation with (a) Resistance (b) Capacitance for circuit



(a)



(b)

Figure 8. Phase error between Vout1 and Vout2 with (a) Capacitance (b) Resistance for circuit

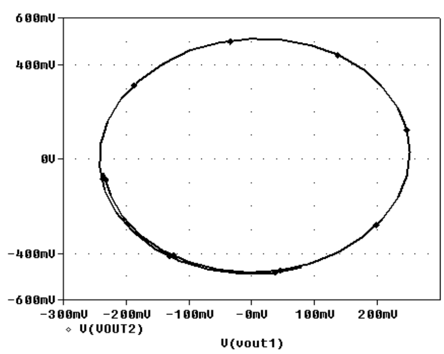


Figure.9. Plot of Vout1 vs Vout2

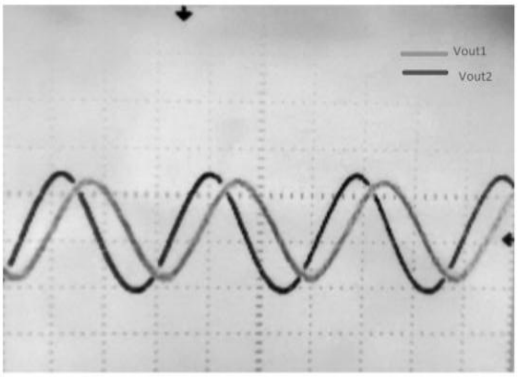


Figure.10. Transient Response of Proposed circuit on CRO

The plot of $Vout_1$ vs $Vout_2$ is shown in Figure 9. The proposed quadrature oscillator is also tested experimentally by bread boarding the circuit of Figure 2 and the corresponding transient response shown in Figure 10.The OTRA is realized using Current feedback operational amplifier (CFOA) IC AD844AN [34] with power supply of ± 8V.

8. CONCLUSION

New realization of OTRA based third order quadrature oscillator is presented in this paper using a high pass filter and a differentiator. The functionality of proposed structure is verified through SPICE simulations using 0.18 µm technology parameter. This topology is further tested experimentally where in the OTRA is realized using off the shelf CFOA IC AD844.The simulation and experimental results are found to be in close agreement with theoretical propositions. The simulated value of % THD is quite low. The phase noise analysis is also discussed for the proposed Q.O.The sensitivity of ω w.r.t passive components is also calculated and observed to be low.

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