



Low Power CMOS Oscillator-Mixer for Wireless Communication Applications

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Abstract— This paper describes design and simulation of a compact down-conversion oscillator-mixer for wireless communication applications. Simulation performed by Advanced Design System (ADS) in 0.18 μ m TSMC CMOS technology. Two different components, the mixer core, and the voltage-controlled oscillator (VCO) are assembled in a stacked structure. This permits the mixer current to be reused by the VCO cross-coupled pair to decrease the total current consumption of the VCO and mixer. The proposed oscillator -mixer operates over a radio frequency (RF) input range of 4.6 GHz to 5.6 GHz with a tunable local oscillator (LO) from 2.1 GHz to over 3.1 GHz. The intermediate frequency (IF) was kept constant at 2.5 GHz. The simulation results demonstrate 12.24 dB conversion gain at 4.6 GHz, 9.26 dB single sideband (SSB) noise figure, -115.6 dBc/Hz phase noise at 1 MHz offset from 2.5 GHz LO frequency and 3.49 mW total DC power consumption.

Keywords— Oscillator-mixer, Low Power, Low Noise, Conversion gain.

I. INTRODUCTION

Demand is increasing for high bit rate in wireless communications consumer products. Low power and highly integrated circuits (ICs) are important issues for developing components of wireless communications systems. Circuits combining oscillator and mixer using GaAs, Bi-CMOS, and CMOS technologies were produced [1], [2], for achieve a high degree of integration and reducing power dissipation. A number of different oscillator-mixer circuits have been presented in [3]-[7]. In the RFIC-based oscillator-mixers, the switching core of the Gilbert cell has been used because it is a double-balanced configuration and it inherently suitable for integration with differential LC-tank oscillators. The LC oscillator can be stacked either above [3] or below [4] the switching core due to particular design requirements. In some oscillator-mixers a harmonic frequency of the local oscillator is used in the mixing operation instead of the fundamental signal to enable the oscillator-mixer to work at higher frequencies [5], [6]. This paper presents a low power LC-tank double-balanced oscillator-mixer that employs an improved flicker noise and conversion gain. The proposed topology stacks the mixer on the VCO to do the total current consumption reduction by using the current reuse technique. In this structure the mixer current flows into VCO cross-coupled pair directly.

II. CIRCUIT DESIGN AND ANALYSIS

A. proposed oscillator-mixer scheme

Active mixer has a larger gain and relaxes noise performance than a passive mixer; most down-converter ICs use active mixer configurations. Most commonly used active mixers is the double-balanced Gilbert mixer which consist of an input transconductance stage, an LO switch stage, and output loads. Fig. 1(a) illustrates the schematic of a conventional down-conversion double-balanced mixer. The transconductances of the n-MOS transistors (M_5 and M_6) are used to convert input RF voltage signals to currents. The n-MOS devices ($M_1 - M_4$) are the time-variant section to mix the RF signals with the LO signals to create the IF signals. Fig. 1(b) illustrates a conventional differential VCO. The device M_3 is the current source, and devices ($M_1 - M_2$) are the negative- G_m pair. The oscillating frequency is obtained by the LC-tank. The proposed oscillator-mixer is shown in Fig. 2. The circuit consist of conventional mixer and regular mixer and conventional VCO. The devices (M_7 and M_8) play an important role in the operation of the oscillator-mixer to prepare the necessary current to the n-core VCO; therefore the switching pairs ($M_1 - M_4$) can be biased with a low overdrive voltage, which reduces the consumption power.

The mixer current flowing into the LC-tank VCO, flicker noise and thermal noise from mixer get translating into phase noise. Flicker noise or thermal noise increasing will deteriorate the phase noise performance. Decreasing transistor sizes ($M_1 - M_4$) improve the phase noise of the VCO core. But use of small-sized switching transistors will degenerate the conversion gain of the mixer. The devices (M_7 and M_8) reduce mixer flicker noise without degenerate the conversion gain of the mixer. The flicker noise of the switching pairs appears directly at the output without frequency translation, which is caused by the direct and indirect mechanism and is described in detail in [8]. To minimize the effect of the direct mechanism, the biasing current through the switching pairs ($M_1 - M_4$) should be reduced. Thus, the devices ($M_7 - M_8$) are used here to lower the DC current flow. The indirect flicker noise mechanism is caused by the tail-capacitance of the switching pairs.

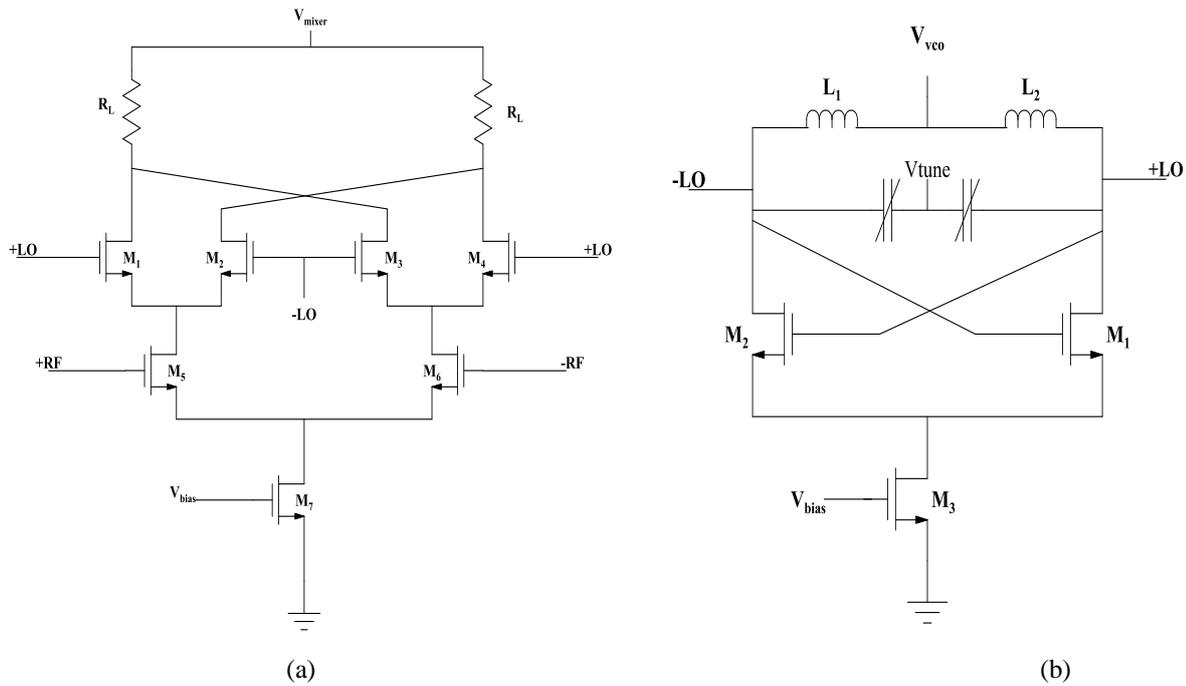


Fig. 1 (a) Conventional double-balance Gilbert mixer and (b) Conventional VCO

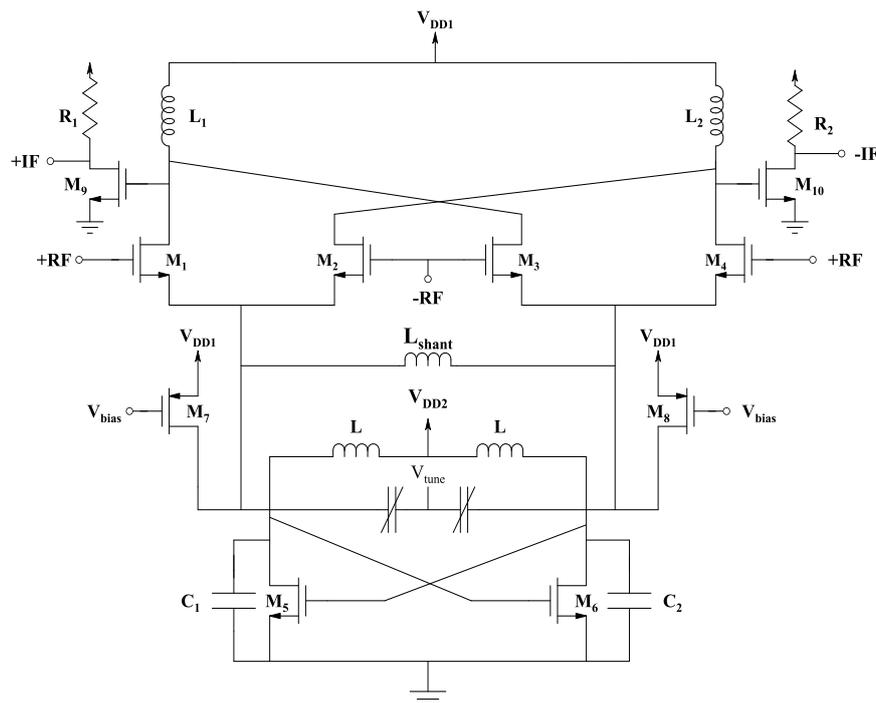


Fig. 2 The proposed oscillator-mixer.

The parasitic capacitances (C_{par}) between the switch's sources and ground, not only prepare leakage paths that degenerate Gilbert mixer's conversion gain, but also corrupt input third-order intercept point (IIP3) due to the nonlinear characteristics of the parasitic elements. The problem will become more serious when increase the circuit's operation frequency. By resonance the capacitance with a parallel inductor, L_{shunt} , the indirect flicker noise is reduced. Now, there is no reason for decrement the transistor sizes ($M_1 - M_4$), thus the conversion gain will not degenerate. The devices (M_9 and M_{10}) act as a buffer. The inductor ($L_1 - L_2$) is proposed for the free of voltage headroom and help to reduce flicker noise.

The input impedance of proposed oscillator-mixer shown in Fig. 2 is given by the following expression:

$$Z_{in} = \frac{-2}{g_m - j\omega C_{gs}} \tag{1}$$

with set capacitors($C_1 - C_2$), the input impedance can be written as:

$$Z_{in} = \frac{-2}{g_m - j\omega(C_{gs} + C)} \quad (2)$$

The Q factor of LC-tank VCO can be increased by decreasing the input impedance. Therefore according to the phase noise model proposed by Hajimiri and Lee, Eq. (15), the phase noise of LC-tank VCO will reduce.

The minimum supply voltage for the mixer and the VCO shown in Fig. 1 are:

$$V_{mixer} = (V_{GS7} - V_{TH}) + (V_{GS5} - V_{TH}) + (V_{GS1} - V_{TH}) + V_{load} \quad (3)$$

$$V_{vco} = (V_{GS3} - V_{TH}) + V_{GS1} \quad (4)$$

But the minimum supply voltage (V_{mixer}) for the mixer shown in Fig. 2 is:

$$V_{mixer} = V_{GS5} + [(V_{GS1} - V_{TH}) - A_{LO}] + V_{load} \quad (5)$$

Where A_{LO} is the magnitude of the oscillating signal from the cross-coupled pair($M_5 - M_6$). However, V_{GS1} is equivalent to A_{LO} for the large oscillating signal. Therefore:

$$V_{mixer} = V_{GS5} - V_{TH} + V_{load} \quad (6)$$

This supply voltage is lower than that of a conventional mixer, Eq. (3) by $2(V_{GS} - V_{TH})$, if V_{load} are the same. The minimum supply voltage (V_{vco}) for the VCO is given as follows:

$$V_{vco} = V_{GS5} \quad (7)$$

This is lower than that of conventional VCO, Eq. (4) by $2(V_{GS} - V_{TH})$, if the processes are the same.

B. Voltage Gain Analysis

As shown in Fig. 2, the differential RF currents are associated with the transconductances ($g_{m1} - g_{m4}$) and the RF input signal ($v_{RF} = \frac{1}{2} A_{RF} \cos(\omega_{RF} t)$). Since v_{RF} is small, the relationship of them can be approximated as:

$$I_{RF1}(t) \approx I_1 + g_{m1} \frac{1}{2} A_{RF} \cos(\omega_{RF} t) \quad (8)$$

$$I_{RF2}(t) \approx I_2 + g_{m2} (-\frac{1}{2} A_{RF} \cos(\omega_{RF} t)) \quad (9)$$

where I_1 and I_2 are the bias current of the transconductors ($g_{m1} - g_{m4}$). After the switching function provided by the VCO, the IF current generated from M_1 and M_2 can be written as:

$$I_{IF1,2}(t) = g_{m1,2} A_{RF} \cos(\omega_{IF} t) \quad (10)$$

where $g_{m1,2} = g_{m1} = g_{m2}$. For the same reason, the IF current produced from M_3 and M_4 is:

$$I_{IF3,4}(t) = g_{m3,4} (-A_{RF} \cos(\omega_{IF} t)) \quad (11)$$

where $g_{m3,4} = g_{m3} = g_{m4}$. Due to the differential relation of $g_{m1,2}$ and $g_{m3,4}$, shown in Fig. 2, the differential IF output voltage is:

$$v_{IF}(t) = R_L g_m A_{RF} \cos(\omega_{IF} t) \quad (12)$$

where $g_m = g_{m1} = g_{m2} = g_{m3} = g_{m4}$. Neglect of the body effect and channel length modulation in all devices, the transconductance (g_m) for n-MOS ($M_1 - M_4$) can be written as:

$$g_m = \frac{\partial}{\partial V_{GS}} \left\{ \frac{1}{2} \mu_n C_{ox} \frac{W}{L} [V_{GS} - V_{TH}]^2 \right\} = \mu_n C_{ox} \frac{W}{L} (V_{GS} - V_{TH}) = \mu_n C_{ox} \frac{W}{L} [V_{RF} - V_{DD2} - V_{TH} + A_{LO}] \quad (13)$$

where A_{LO} is the amplitude of the oscillating signal. If the sinusoidal oscillating signal has a 50% duty cycle, the voltage conversion gain of the oscillator mixer can be derived as:

$$CG_v = \frac{1}{2} \mu_n C_{ox} R_L \frac{W}{L} [V_{RF} - V_{DD2} - V_{TH} + A_{LO}] \quad (14)$$

C. Phase Noise Analysis

The phase noise is another important issue for designing the oscillator-mixer According to the phase noise model proposed by Hajimiri and Lee, the phase noise of LC- tank VCO operated in the current-limited regime is given by [9]-[11]:

$$L(\Delta\omega) = \frac{\gamma g_{n,p} + (1/Q_{ind} r_p)}{C_{var} I_B Q_{ind} r_p} \frac{2kT\Gamma_{rms}^2}{\Delta\omega^2} \tag{15}$$

Where r_p is the parasitic resistance of the inductor, $g_{n,p}$ is the conductance of the cross-coupled pairs, Q_{ind} and I_B are the Q factor of the inductor (L) and the bias current of the VCO, respectively. $\Delta\omega$ is the offset frequency from the carrier, and Γ_{rms} is the rms value of the effective impulse sensitivity function (ISF) which represents the time-varying sensitivity of the distributions of phase noise. The parameter γ has a value of unity at zero V_{DS} and 2/3 in saturation region with long channel devices. Eq. (15) shows that $L(\Delta\omega) = f(Q_{ind}, I_B, C_{var})$ can be improved by increasing Q_{ind} , or designing I_B near the maximum position in current limited regime.

D. Current Reuse in Oscillator Mixer

Fig. 3 shows the DC paths of the oscillator mixer. The DC currents from V_{DD1} and V_{DD2} are defined as I_{MIXER} and I_{DD2} , respectively. The DC current (I_{MIXER}) through the mixer ($M_1 - M_4$) flows only into the cross-coupled pair ($M_5 - M_6$), and finally sink into ground. The DC current (I_{VCO}) can be written as:

$$I_{VCO} = I_{MIXER} + I_{DD2} \tag{16}$$

By increasing the mixer supply voltage (V_{DD1}) with a constant supply voltage (V_{DD2}), the mixer current (I_{MIXER}) from V_{DD1} is increased and the current (I_{DD2}) from V_{DD2} is reduced with the same amount. Moreover, the total currents flow through the VCO devices (I_{VCO}) is constant. In other words, the mixer current (I_{MIXER}) can be reused by the VCO cross-coupled pair ($M_5 - M_6$).

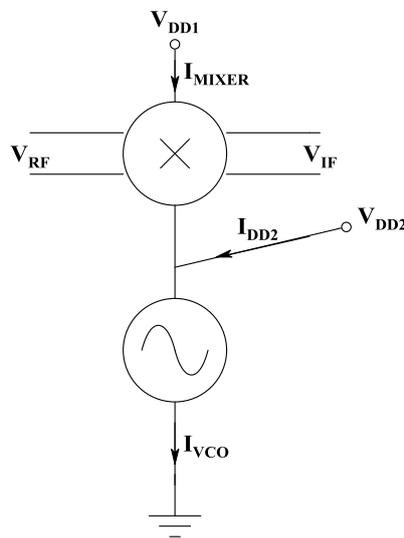


Fig. 3 Direct current (DC) path of the oscillator mixer.

III. SIMULATION RESULTS

The proposed has been simulated by 0.18- μm CMOS technology. The circuit operates over an RF input range of 4.6 GHz to 5.6 GHz with a tunable LO from 2.1 GHz to over 3.1 GHz. The IF was kept constant at 2.5 GHz in the simulations. From Fig. 4 the maximum conversion gain for the oscillator-mixer is 12.24 dB. This process was made by sweeping the RF and LO signals together in order to maintain a constant IF output frequency. Fig. 5 shows the conversion gain results as a function of RF input power. Fig. 6 shows the simulated output power versus input power, as it can be seen P_{1dB} is - 11.34 dBm.

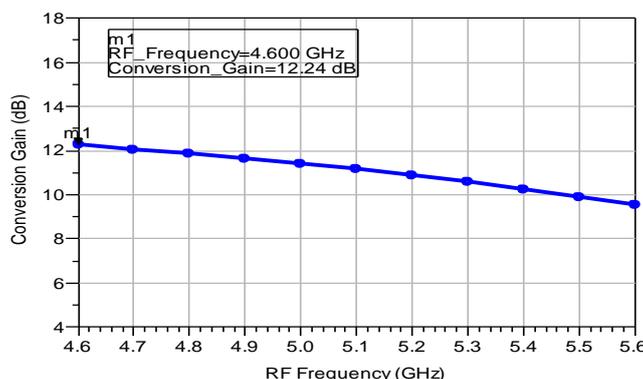


Fig. 4 Conversion gain versus input RF frequency with a constant 2.5 GHz IF.

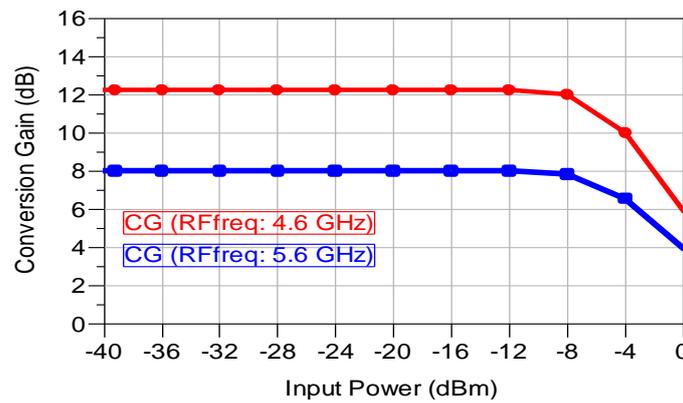


Fig. 5 Simulated conversion gain versus input RF power.

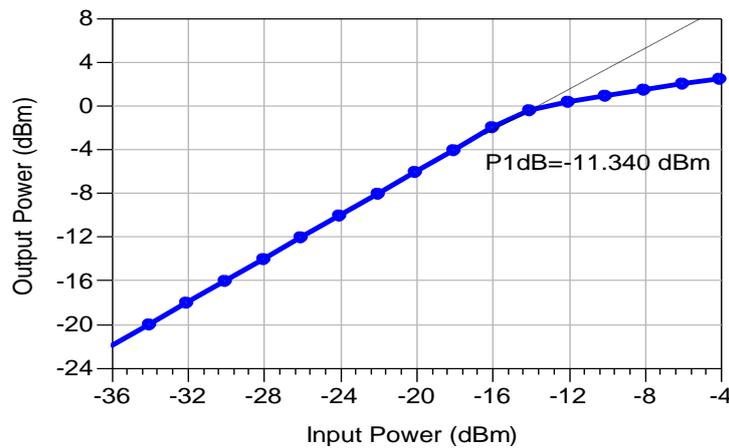


Fig. 6 Simulated output power versus input power.

A two-tone test was carried out to determine the third-order intercept (TOI) point of this oscillator-mixer. The two tone test for third-order intermodulation distortion (IIP3) is shown in Fig. 7 and it is observed that the oscillator-mixer has an IIP3 of -8.74 dBm. The SSB noise figure of the mixer is shown in Fig. 8. The minimum SSB NF for the proposed oscillator-mixer is 9.26 dB. The tuning range of the VCO was determined by the LO leakage signal at the IF port and the result shown in Fig. 9, that denotes the oscillator can be tuned from 2.1 GHz to 3.1 GHz when control voltage changes from 0 V to 1.5 V. The supply voltages for mixer and VCO are $V_{mixer} = 1.7$ V and $V_{vco} = 1.52$ V, respectively. The total DC power consumption is 3.49 mW for proposed structure. The phase noise of the VCO is -115.6 dBc/Hz at 1-MHz offset from the carrier as shown in Fig. 10 The figure-of-merit (FOM) of VCO core is given by the following expression:

$$FOM = L\{f_m\} + 10 \log \left[\left(\frac{f_m}{f_o} \right)^2 P_{DC} \right]. \tag{17}$$

where $L\{f_m\}$ is the phase noise, (f_o) the carrier frequency, and (P_{DC}) the DC power consumption. FOM of the proposed configuration is -180.9 dBc/Hz.

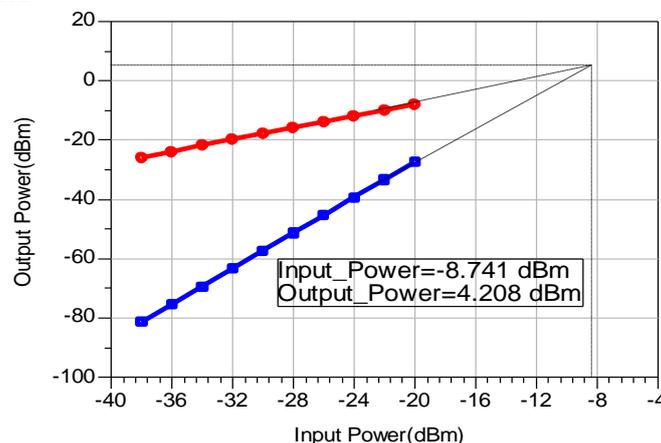


Fig. 7 Two tone test for third-order intermodulation distortion (IIP3).

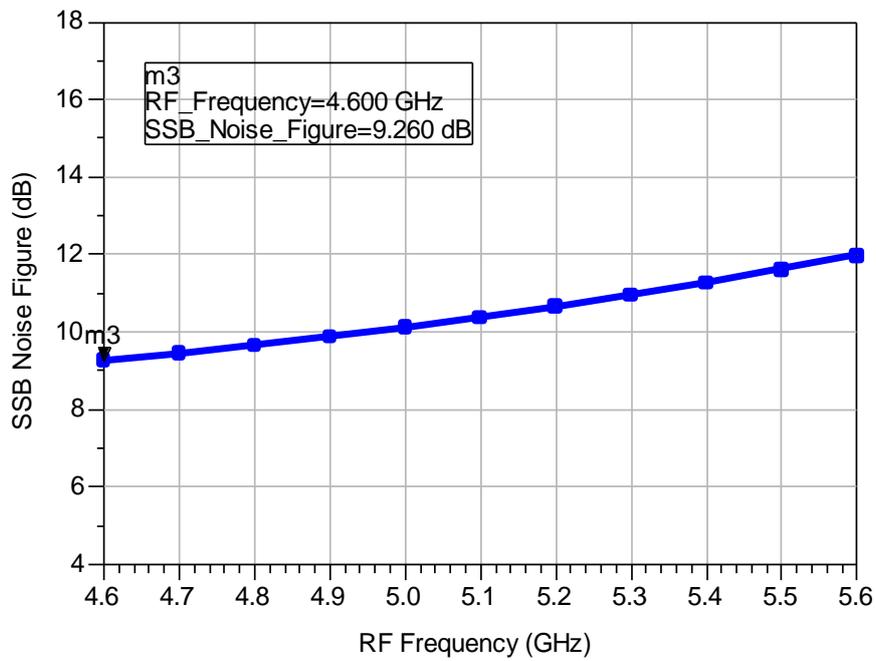


Fig. 8 SSB noise figure versus input RF frequency with a fixed 2.5 GHz IF .

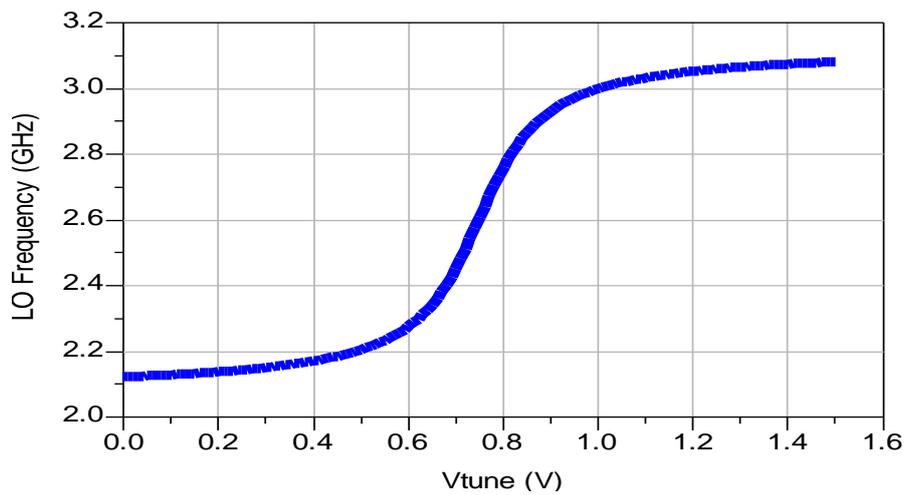


Fig. 9 LO frequency versus V_{tune} .

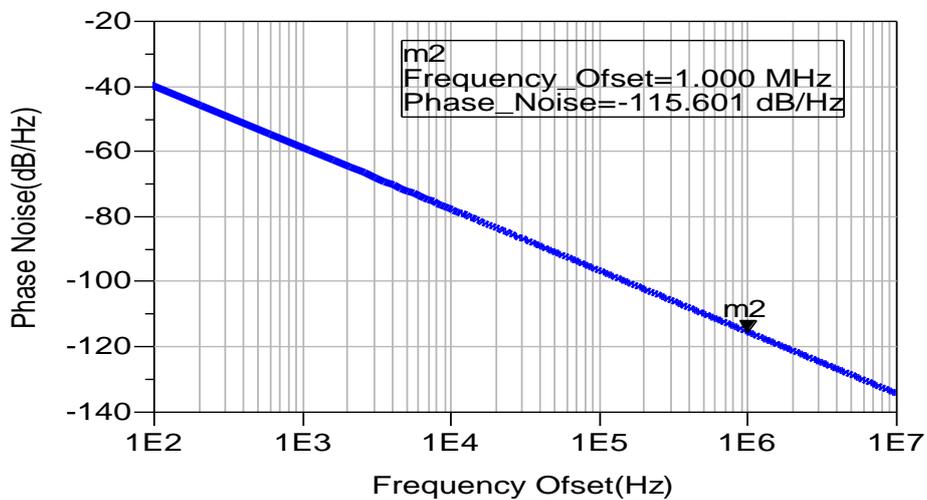


Fig. 10 Phase noise versus offset frequency

IV. Conclusion

The proposed low power CMOS oscillator-mixer presents wider tuning frequency range, good conversion gain, and improved phase noise. Table I compares the low power mixers in the CMOS technology [3-4, 12-14] with this structure. The low supply voltage and low power consumption is achieved by using the individual supply voltages and current reuse technique. By stacking the oscillator, and the mixing core on top of each other, the DC currents can be reused and this leads to moderately low power consumption. As a result, the oscillator-mixer combines the individual mixer and VCO to achieve low supply voltage, low current consumption and low power consumption with competitive conversion gain. For all that, proposed oscillator-mixer is particularly suitable for wireless communication applications.

TABLE I
Performance Comparison of Reported Low Power Mixers with This Work

	[3]	[4]	Mixer[12]	Mixer[13]	[14]	This Work
Technology	0.18 μ m-CMOS	0.18 μ m-CMOS	2 μ m-CMOS	0.18 μ m-CMOS	0.13 μ m-CMOS	0.18μm-CMOS
RF Freq.(GHz)	1.57	4.2	10	8	7.8-8.8	4.6-5.6
Conversion Gain(dB)	36	10.9	13	6.5	11.6	12.24
SSB Noise(dB)	4.8(DSB)	14.5	24	22	4.39(DSB)	9.26
IIP3(dBm)	-19	-11.8	-10.6	-11.2	-8.3	-8.74
1dB Comp.point(dBm)	-31	-	-15.72	-16.74	-13.6	-11.34
Phase Noise(dBc/Hz)	-108.8	-107.2	-111.6	-109.5	-103.5	-115.6
DC Power(mW)	5.4	3.14	6.6	6.9	12	3.49

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