# A 8–14 GHz Varactorless Current Controlled LC Oscillator in 16nm CMOS Technology

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***Abstract*—This paper presents a new concept of fine frequency tuning in a differential LC oscillator by injecting quadrature phase shifted current into the tank that can replace the use of a conventional varactor, eliminating all limitations associated with it, such as specific bias voltage requirement, limited Q-factor, technology/process uncertainties, model accuracy etc. Wide tuning range of 8-14 GHz is achieved by 5-bit switched capacitor array distributed across the exit legs of the inductor. The circuit is implemented in 16nm CMOS technology. Oscillator core consumes 2.85mW power from 0.9V supply at 10GHz frequency for 400mV single-ended swing. Phase noise is -108dBc/Hz at 1MHz offset for no tuning current injection.**

***Keywords—LC oscillators; current controlled oscillators; phase noise; common-mode feedback;***

## I. INTRODUCTION

Oscillators are one of the fundamental blocks in any Phase-Locked Loop (PLL) based system used for clock generation and data recovery of a communication system. In high performance application, LC oscillators are preferred over ring for their lower phase-noise and power consumption at operating frequencies of several GHz. Fine frequency tuning in LC oscillators is conventionally done using varactor. But varactor loss significantly reduces tank quality factor at higher frequencies. Also it requires a specific bias voltage at the oscillating nodes. Additional circuitry e.g. resistor at the tail, is needed and that affects oscillator performance. Gain of the VCO ($K_{VCO}$) is entirely decided by the voltage dependence of varactor capacitance, which is again technology dependent [1]. This $K_{VCO}$ is very high in advanced CMOS technologies. Higher $K_{VCO}$ is better for covering wide tuning range, but causes significant amount of AM (coming from PLL loop filter) to FM noise conversion that increases output jitter and phase noise [2]. Therefore finding an alternative of varactor tuning is important. One of the alternatives proposed in [3] is phase shifted tuning. But this suffers from higher phase noise and tuning range is also strongly dependent on tank quality factor. A varactorless frequency tuning technique is described in [4] by changing the bias of the coupling transistors in a quadrature VCO. But the idea is limited to QVCOs only. The technique proposed here is extended for any LC oscillator by a pair of quadrature phase shifter. The introduction of phase shifter gives the freedom to control $K_{VCO}$ by controlling phase shifter gain. A common-mode feedback (CMFB) circuit is designed to improve fine tuning range and also to make the circuit robust under process-voltage-temperature (PVT) variations.

Section II describes the concept of current controlled frequency tuning and derives the expression for oscillation frequency. Section III is about the design details of the oscillator. Simulation results are shown in section IV, followed by conclusion in section V.

## II. CURRENT CONTROLLED FREQUENCY TUNING

Fig. 1 explains the basic concept of current controlled frequency tuning with phase diagram. An LC oscillator consists of an LC tank and a cross-coupled pair that compensates the losses of the tank. Under resonant condition, the current flowing through the inductor ($i_L$) and capacitor ($i_C$) in LC tank is 90° phase shifted to the voltage at the oscillating nodes ($V_{osc}$). Also $i_L$ equalizes $i_C$ but flows in opposite direction.

Now if some additional current ($i_{inj}$) which is in phase with either $i_L$ or $i_C$, is injected into the tank, new oscillation condition will be established and the tank will oscillate at a different frequency from its free running oscillation frequency i.e. $1/(2\pi\sqrt{LC})$ depending on the strength of $i_{inj}$. This $i_{inj}$ can be generated from tank oscillation voltage ($V_{osc}$) itself using a $g_m$-cell (that works as a voltage-controlled-current-source). But this requires an additional ±90° phase shifter to make it phase aligned with $i_L$ or $i_C$. In addition, the rate of change of frequency i.e. the gain of the oscillator can be adjusted by controlling the gain of the $g_m$-cell and phase shifter.

If tank oscillation voltage is $V_{osc} = V \sin \omega t$, LC tank currents can be written as: $i_L = (V / L\omega)\cos \omega t$, $i_C = V\omega C \cos \omega t$ and $i_{inj} = \pm g_m KV \cos \omega t$. $g_m$ is the trans-conductance of $g_m$-cell and K is the gain of 90° phase shifter.

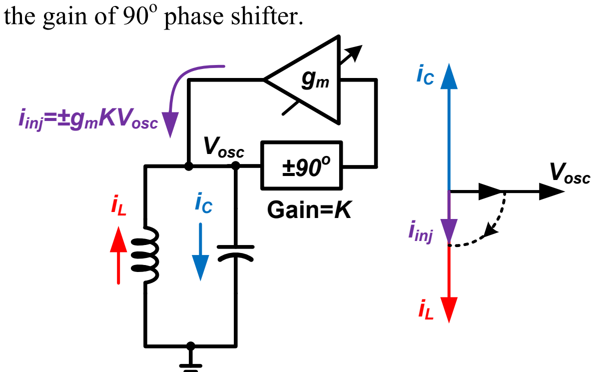



Fig. 1. Current controlled frequency tuning concept with phase diagram

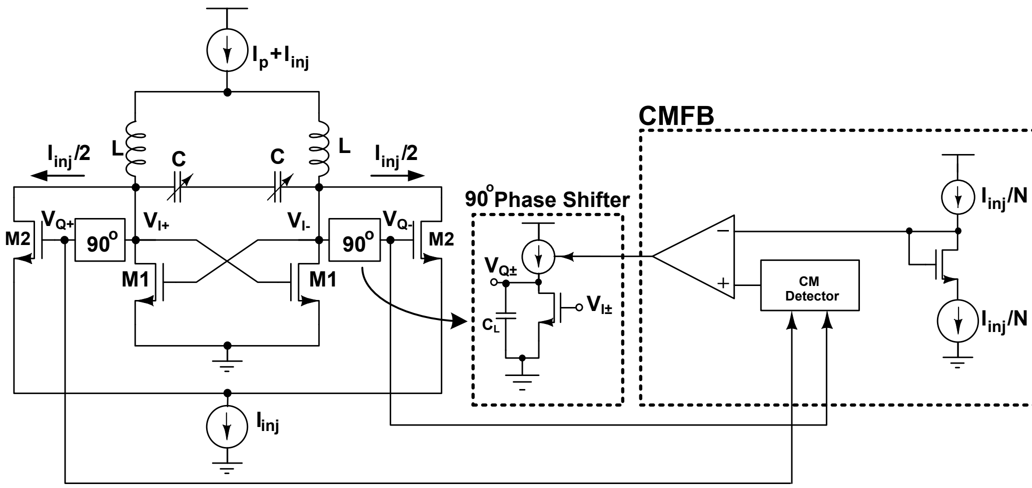



Fig. 2. Circuit schematic of current controlled LC oscillator

Therefore KCL at oscillation node, depending on the polarity of injection, is given by:

$$-\frac{V}{L\omega}\cos\omega t + V\omega C\cos\omega t = \pm g_m K V\cos\omega t$$

$$-\frac{1}{L\omega} + \omega C = \pm g_m K$$

$$-\omega^2 LC \pm g_m K L\omega + 1 = 0$$

$$\boxed{\omega \approx \frac{1}{\sqrt{LC}} \pm \frac{g_m K}{2C} = \frac{1}{\sqrt{LC}} \pm \frac{K\sqrt{\mu C_{ox}\frac{W}{2L}I_{inj}}}{C}}$$

Above equation shows a linear relationship between oscillation frequency and $g_m$ i.e. $\sqrt{I_{inj}}$ . Frequency can be tuned in higher or lower side of tank resonant frequency depending on the polarity of current injection. In this implementation lower side is used. $K_{VCO}$ can be well controlled by controlling phase shifter gain (K) and $g_m$.

As we keep on increasing $g_m$, noise produced by $g_m$-cell and phase shifter will affect the oscillator phase noise performance. But tank Q is higher than conventional VCO because of the absence of varactor. So noise filtering will be better and overall phase noise will still be lower than varactor tuning case for a narrow fine frequency tuning range. Phase noise comparison results are shown later in section IV (Fig.5(b)).

## III. LC CCO DESIGN

Fig. 2 shows the schematic of current controlled LC oscillator designed in 16nm CMOS technology. Inductor L, capacitor C and cross-coupled pair M1 form conventional differential LC oscillator. Here two additional 90° phase shifter along with a pair of transistor M2 that serves as a $g_m$-cell, are introduced for current controlled frequency tuning.

As we are dealing with sinusoidal signals, 90° phase shifter is nothing but an integrator and it can be made by a capacitive load common-source (CS) amplifier. If $g_{mi}$ is the trans-conductance of CS amplifier and $C_L$ is load capacitance, integrator will operate accurately as long as $\omega >> 1/R_{out}C_L$. Gain of the integrator, $K = g_{mi}/\omega C_L$, is nearly constant over a small fine tuning range.

A common-mode feedback (CMFB) circuit is required to adjust the gate bias voltage of M2 ($V_{Q\pm}$) to make sure that additional dc injected tuning current ($I_{inj}$) flows through M2 only for large variation of $I_{inj}$ without modifying actual tank operating condition. This effectively increases the fine tuning range and make the circuit robust over PVT variations. CMFB circuit compares the CM outputs of phase shifter with the required bias voltage and adjusts the bias current of the integrator. The stability of CMFB loop is also taken care of by setting a phase-margin of 45° for a loop gain of 70dB (typical condition).

Inductor is symmetric two-turn stacked octagonal shape. It uses top two metal layers (metal 12, metal 11 in parallel and metal 13) to reduce the parasitic capacitance to ground. Outer diameter of the inductor is only 110µm. Differential inductance and quality factor are approximately 350pH and 15 respectively at 10GHz.

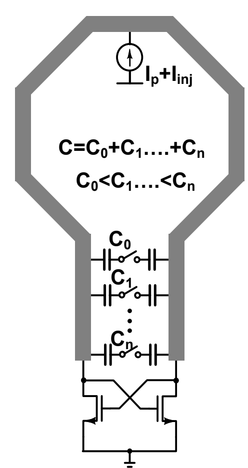



Fig. 3. Tuning range extension technique [5]

Frequency coarse tuning is performed with 5-bit binary weighted switched-capacitor bank. These capacitors are distributed across the exit legs of the inductor with LSB closest and MSB furthest to the coil (as shown in Fig.3) [5]. When MSB turns ON, current flows through the longest path and effective inductance increases. Opposite thing happens when LSB turns ON. Therefore we can get higher inductance when capacitance is highest and vice versa. That effectively increases the tuning range.

## IV. Simulation Results

All circuit simulations are performed in SpectreRF tool. Inductor model is extracted using EMX simulator. Oscillator core requires 3mA bias current from a supply of 0.95V to maintain output swing of 400mV (single ended) under typical condition.

### A. Transient Response

Fig. 4 shows the transient response of the oscillator. Oscillator output voltage ($V_{osc}$) and the inductor ac current ($i_L$) are 90$^o$ phase shifted to each other. Injected ac current ($i_{inj}$) in the tank is in opposite phase (i.e. 180$^o$ ) of $i_L$ as it is generated from the quadrature shifted version of output voltage. The amplitude of $i_{inj}$ is controlled by dc injected bias current ($I_{inj}$).

### B. Fine Frequency Tuning

Fig. 5(a) shows simulated oscillation frequency for different injected currents (0 to 2.4mA) at three different PVT corners. Frequency reduces continuously as $I_{inj}$ increases. Gain of the oscillator calculated from the plot is typically 150MHz/mA. Fig. 5(b) is the plot of phase noise at different frequencies obtained by tuning $I_{inj}$. Phase-noise increases with the increase in injected current. This is expected, as the gain of $g_m$-cell stage increases and the amplified noise of 90$^o$ phase shifter that appears at differential node also increases. Noise is maximum in fast corner (FF) as frequency is highest, but it requires less $I_{inj}$ to cover a given fine tuning range. Phase noise values for a conventional varactor based tuning (control voltage variation 0 – 0.95V) are also plotted to compare noise performances.

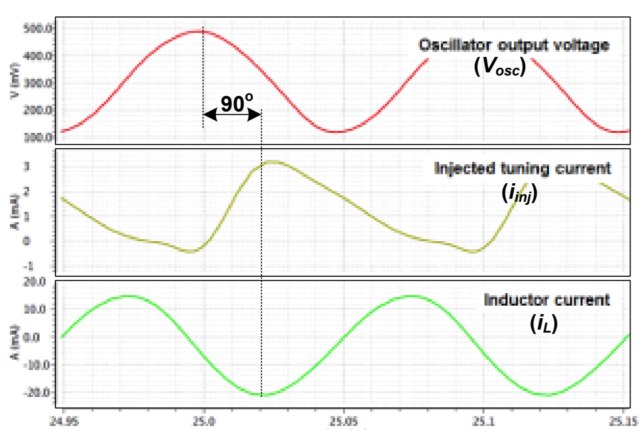
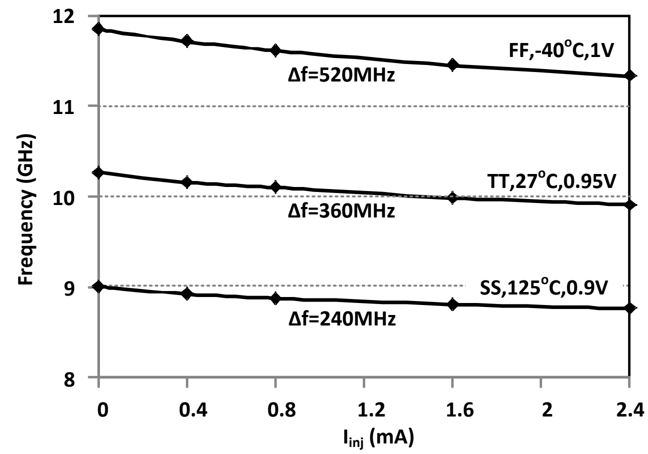


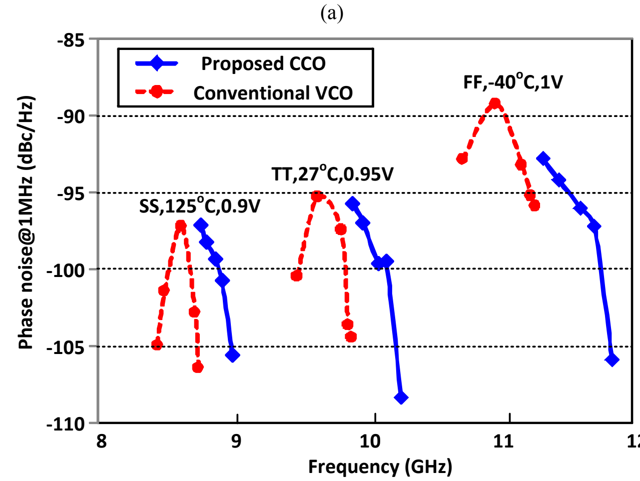

Fig. 4. Transient response (single-side) showing current and voltage phases (i.e. $V_{osc}$, $i_{inj}$, $i_L$ as in Fig. 1)

### C. Coarse Frequency Tuning

Frequency range (8.3 – 14.2 GHz in typical corner) achieved by tuning 5-bit switched capacitor bank is shown in Fig. 6(a). Phase noise corresponding to each frequency point is plotted in Fig. 6(b). Fine tuning current branch ($I_{inj}$) is turned OFF during coarse tuning simulation.




(a)




(b)

Fig. 5. Fine frequency tuning (a) frequency vs $I_{inj}$ plot (b) phase noise vs frequency plot

Table 1 compares the performance of this oscillator with other state of the art oscillator designs. The formula used for FoM calculation is:

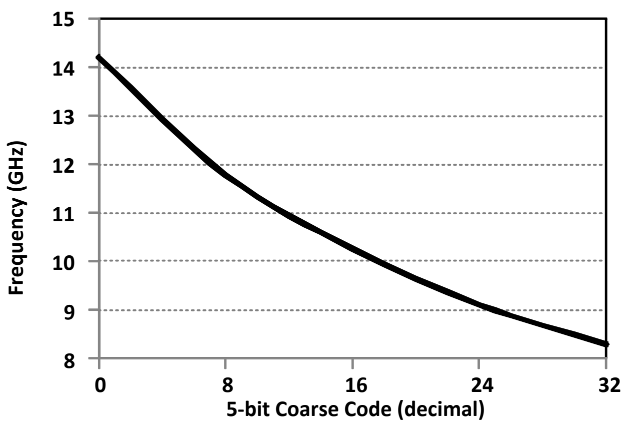
$$FoM = -10\log\left[\left(\frac{f_o}{\Delta f}\right)^2 \frac{1mW}{P_{diss}}\right] + L\{\Delta f\}$$

$L\{\Delta f\}$ is phase noise of the oscillator at $\Delta f$ offset frequency when oscillator is running at $f_o$ frequency. FoM is best compared to other designs in the list. Phase noise is also comparable to other designs. However, this phase noise value is for zero $I_{inj}$. For maximum $I_{inj}$, phase noise increases to -96dBc/Hz at 1MHz offset (as shown in Fig. 5(b)) and FoM becomes -171.5 dBc/Hz. Power consumption is lowest in the list.

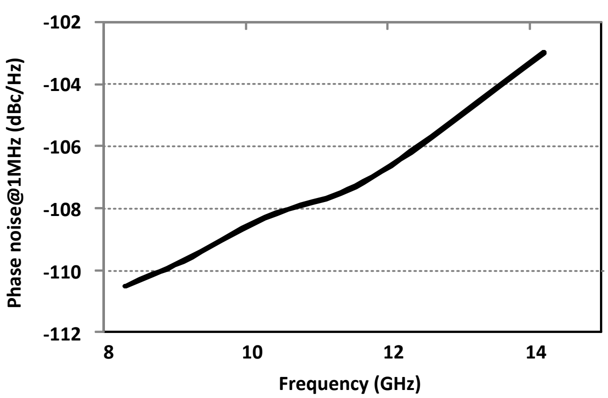
TABLE I. PERFORMANCE COMPARISON

| Reference | Process | Frequency range (GHz) | Output frequency, $f_o$ (GHz) | Phase-Noise @1MHz(dBc/Hz) | Core $P_{diss}$ (mW) | FoM (dBc/Hz) |
|---|---|---|---|---|---|---|
| [6] | 90nm-CMOS | 7 – 9.5 | 8 | -109 | 14.3 | -175.5 |
| [7] | 65nm-CMOS | 14.6 – 22.2 | 20 | -98 | 4 | -178 |
| [8] | 20nm-CMOS | 2- 16 | 15 | -115 | 46.2 | -182 |
| [9] | 28nm-CMOS | 9.9 – 14.6 | 11.75 | -102 | 15.5 | -171 |
| **This work*** | **16nm CMOS** | **8.3 – 14.2** | **10** | **-108** | **2.85** | **-183.5** |

*Typical corner simulation results




(a)




(b)

Fig. 6. Typical corner coarse frequency tuning (a) frequency vs coarse code plot (b) phase noise vs frequency plot

## V. CONCLUSION

In this paper a current controlled LC oscillator is designed in 16nm CMOS technology replacing the varactor with a $90^o$ phase shifter and a $g_m$-cell. There by design limitations associated with the varactor can be easily eliminated. DC power consumption of the oscillator core is 2.85mW typically at 10GHz frequency. Phase noise and FoM is comparable to other state of the art LC VCO designs.

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