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<td><strong>Author(s)</strong></td>
<td>Mahalingam, Nagarajan; Ma, Kaixue; Yeo, Kiat Seng; Lim, Wei Meng</td>
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A $K$-band High PAE Wide Tuning Range VCO Using Triple-Coupled LC Tanks

Nagarajan Mahalingam, Student Member, IEEE, Kaixue Ma, Senior Member, IEEE, Kiat Seng Yeo, Senior Member, IEEE, and Wei Meng Lim

Abstract—This paper presents a fully differential $K$-band voltage-controlled oscillator (VCO) using commercial 0.18µm SiGe BiCMOS process. By using triple-coupled LC tanks with strong coupling and varactor controlled capacitor bank, the proposed VCO achieves wide tuning range, low phase noise, high output RF power and high power added efficiency (PAE) simultaneously. The implemented VCO demonstrates an oscillation frequency range from 22.50 GHz to 26.23 GHz, frequency tuning range of 15.3%, phase noise of -107.7 dBc/Hz at 1 MHz offset, output power of -3.1 dBm, power added efficiency of 5.97% and a figure-of-merit (FOM) of -186.9 dBc/Hz with a DC power of 8.2 mW.

Index Terms—24 GHz, Coupled LC tanks, Frequency Tuning Range (FTR), K-band, Low Power, Phase Noise, SiGe, VCO.

I. INTRODUCTION

RECENTLY, high data-rate communication systems with increased bandwidth requirements have drawn much attention. To realize wideband and high data rate systems with good performance, it is critical for VCO in the local oscillator to have a wide tuning range with low phase noise and high output power. However, the stringent constraint on low power consumption, severely limits the achievable phase noise, tuning range and especially the power added efficiency (PAE).

Transformer based VCOs with low power low phase noise are reported in [1] – [4] but these works have narrow tuning range and relatively lower output power. Improvement in tuning range is achieved in [5] with dual LC tanks at source and drain terminals but the VCO has a lower output power. High quality transmission line VCOs are reported in [6], [7], but these designs have narrow tuning range and higher phase noise respectively. Inductor based VCOs are reported in [8] - [10] but these designs have narrow tuning range and low output power. Although many VCOs addressed on different aspects like tuning range, power consumption, phase noise, etc., have been reported in literature, it still remains a challenge to design VCOs with optimized trade-off to simultaneously achieve a wide tuning range, low phase noise, high output power and good PAE ($P_{out}/P_{DC}$).

In this paper, we present a fully differential $K$- band Colpitts VCO extending the work in [11] to address the VCO operation based on the theoretical analysis. This design achieves a wide tuning range, low phase noise, high output power and good PAE simultaneously, by using the proposed triple-coupled LC tanks with varactor and capacitor banks. The strong coupling of LC tanks and proper choice of coupling coefficient increase not only the equivalent tank quality factor and VCO tuning range but also the delivered output power and PAE. The operation mechanism of the tank circuit is investigated theoretically using the equivalent circuit model and verified experimentally based on a commercial SiGe BiCMOS process.

This paper is organized as follows: In section II, analysis of triple-coupled LC tanks and the implementation of the proposed VCO is presented. Section III gives the measurement results and the conclusion is presented in Section IV.

II. CIRCUIT DESIGN AND ANALYSIS

The circuit schematic of the proposed Colpitts VCO is shown in Fig. 1. The negative resistance of the Colpitts oscillator is generated by the fixed capacitor $C_1$, transistor base-emitter parasitic capacitance $C_p$ and coarse tuning varactors in the emitter ($C_{eo} - C_{e3}$). In the proposed VCO, the triple-coupled LC
A. Analysis of triple-coupled LC tanks

The simplified circuit schematic of the triple-coupled coils is shown in Fig. 2. The self inductance of the primary and secondary coils of the triple-coupled LC tanks is given by $L_{pi}$ and $L_{si}$ respectively; $R_p$ represents the loss in the primary coil, $C_{si}$ and $C_{sc}$ represent the total effective capacitances connected to the secondary coils formed by capacitance of varactors $C_{v1}$ and $C_{v2}$ respectively; $R_{v1}$ and $R_{v2}$ represent the losses in secondary coils respectively. Based on circuit theory, the impedance matrix of the triple-coupled LC tanks is:

$$
\begin{bmatrix}
V_1 \\
V_2 \\
V_3
\end{bmatrix} = \begin{bmatrix}
\frac{P}{s} - j\omega M_{12} & -j\omega M_{13} & i_1 \\
-j\omega M_{21} & S_1 - j\omega M_{23} & i_2 \\
-j\omega M_{31} & j\omega M_{32} & S_2
\end{bmatrix} \begin{bmatrix}
i_1 \\
i_2 \\
i_3
\end{bmatrix}
$$

(1)

where, $P = R_p + j\omega L_p$, $S_1 = R_{v1} + j\omega L_{v1}$, $S_2 = R_{v2} + j\omega L_{v2}$. Considering $M_{ij}=M_{ji}$ due to symmetry, the input impedance of the triple-coupled LC tanks can be derived as given in (2), where, the first term represents the primary impedance; the other terms represent the coupling between the primary and the two secondary coils; and $M_{ij} = K_i \sqrt{L_i L_j}$. From (2), the input impedance of the triple-coupled LC tanks can be simplified as shown in (3) – (6) assuming $R_{di} = R_{d2} = R_d, L_{si} = L_d = L$ and $C_{di} = C_{d2} = C$. To gain more design insight, the input impedance of triple-coupled LC tanks can be further simplified neglecting higher order terms as

$$Z_{triple, simplified} = R_{triple, simplified} + j\omega L_{triple, simplified}
$$

(7)

where,

$$R_{triple, simplified} = \left[ R_p + \frac{R_{eff}}{R_p + R_{eff}} \right] = \left[ R_p + \frac{R_{eff}}{R_p + R_{eff}} \right]
$$

(8)

$$L_{triple, simplified} = \left[ \frac{L_p + \frac{L_{eff}}{\omega}}{\omega M_{12}^2 (L_{sc1} M_1 + 2\alpha M_2) - \frac{L_{sc1} M_1 + 2\alpha M_2 L^2}{R_p + L_{eff}}} \right]
$$

(9)

where, $M_1 = \omega^2 (M_{12}^2 + M_{13}^2)$, $M_2 = (\omega^2 M_{12} M_{13})$ and $L_{sc} = (\omega L_p) - (\omega C)^2$. The quality factor of the triple-coupled LC tanks when $\omega$ equals VCO operating frequency $\omega_0$ can be derived as:

$$Q_{triple} = \frac{\omega_0 (R_p + L_{eff})}{R_p + R_{eff}}
$$

(10)

B. Layout and validation of triple-coupled LC tanks

The layout of the triple-coupled LC tanks is shown in Fig. 3 and the simulated self inductance, quality factor and coupling coefficient based on EM simulation is shown in Fig. 4. To validate the effect of the coupling coefficient on the quality factor of the triple-coupled LC tanks, Equation (10) is plotted as shown in Fig. 5. For the simulations, simulated parameters in Fig. 4(a) is used assuming equal coupling coefficients ($k_{12}=k_{13}=k_{23}=k$). $R_p = 0.65$ Ohm, $R_{v1} = 1$ Ohm, $L_{v1} = 100$ pH and $C_{v1} = 100$ IF at 24 GHz. From Fig. 5, it can be observed that the quality factor can be enhanced by providing a stronger coupling between the coils of the triple-coupled LC tanks. The increase in tank’s quality factor improves the energy transfer efficiency of the resonator and enhances the PAE and phase noise of the oscillator [12].

In triple-coupled LC tanks, the effect of quality factor of the secondary coils on $Q_{triple}$ cannot be neglected. To validate the
effect, secondary coil resistive loss ($R_s$) is varied and the resulting $Q_{\text{triple}}$ and $L_{\text{triple}}$ is plotted in Fig. 6. Simulated parameters in Fig. 4 are used by assuming a frequency of 24 GHz and a fixed secondary coil capacitance of 200 fF. It can be observed that, the increased resistive losses in the secondary coils degrade the quality factor of the coupled LC tanks and reduction of $R_s$ is desired to maintain the overall quality factor of the triple-coupled LC tanks ($Q_{\text{triple}}$). Also from Fig. 6, it can be seen that the variation in $L_{\text{triple}}$ is quite small as $R_s$ is increased.

To understand the effect of the coupling coefficient on the VCO frequency tuning range, the variation in $L_{\text{triple}}$ is plotted in Fig. 7 for different coupling coefficients between the primary and two secondary coils against the secondary coil capacitance $C_s$. From Fig. 7, it can be observed that, the maximum variation in $L_{\text{triple}}$ is achieved when strong coupling is present between the primary and the two secondary coils. From (9), intuitively, $L_{sc}M_1$ term dominates the effective inductance and $M_1$ increases with strong coupling between the primary and secondary coils. Therefore, with strong coupling and digitally tuning the secondary coil capacitance $C_s$, the tuning range of the VCO can be maximized by tuning the VCO operating frequency using the large variation in $L_{\text{triple}}$.

The phase noise of the Colpitts oscillator is given by [13]

$$S_{\text{phase}} = \frac{1}{2\pi f_0^2} \frac{1}{|\tan \phi|} \left| \frac{1}{\frac{1}{C_1 + C_{\pi}} \frac{1}{C_1 + C_2} \frac{1}{C_{\text{emitter}}}} \right| \frac{1}{Q_{\text{tank}}} \frac{1}{Q_{\text{emitter}}}$$

(11)

where, $S_{\text{phase}}$ denotes the phase noise spectral density, $\frac{1}{Q_{\text{tank}}} \frac{1}{Q_{\text{emitter}}}$ is the average input white noise power of the transistor, $V_{\text{tank}}$ is the tank swing and $\Delta f_0$ is the offset from the carrier angular frequency. From (11), it can be seen that, the phase noise is influenced by the quality factor of tank circuit and effect on capacitor ratio formed by $C_1$, base emitter capacitance $C_\pi$ and total capacitance connected to the emitter of transistors $Q_1$ and $Q_2$. To characterize the VCO phase noise performance due to capacitor and transistor mismatch, the proposed VCO was
simulated by using Monte-Carlo analysis. The analysis considers the process and mismatch models provided by the foundry including the self-heating in Hicum BJT models. With a total of 100 samples, the simulated VCO phase noise at 1 MHz offset over the entire frequency range and varying vtune is shown in Fig. 8. From the data it can be seen that the phase noise discrepancy of the VCO is within 3 dB over the entire VCO operating range. This change in the phase noise can be attributed to the change in the quality factor of the varactors used for digital coarse tuning in the VCO.

C. Improvement of negative resistance

Though Colpitts oscillator can achieve a lower phase noise, they consume high DC power to achieve the required negative resistance for reliable startup [12]. For the Colpitts oscillator, the negative resistance is given by

\[
\text{Re}[Z_{\text{in}}] = R_{\text{neg}} = \frac{s_m}{\omega^2 C_{IT} C_{\text{equi}}} (12)
\]

where, \( C_{IT} = C_j + C_x \) and \( C_{\text{equi}} \) is the equivalent capacitance connected across the transistor emitter terminals. For low frequencies, the effect of transistor’s collector-base capacitance \( C_{BC} \) is usually neglected and the negative resistance defined by Equation (12) is sufficient. For low power Colpitts VCO operating in K-band and above, the presence of \( C_{BC} \) makes negative resistance in (12) inadequate. This could pose a serious issue for VCO startup in the presence of PVT variations. To comprehend the effect of \( C_{BC} \) across the process corners, the negative resistance of the proposed VCO is simulated with different \( L_c \) values. From Fig. 9, it can be observed that, after 10 GHz, as \( L_c \) increases, the available negative resistance decreases and the usable range is reduced. In the proposed VCO, as shown in Fig 1, the capacitor \( C_M \) in combination with \( L_c \) improves the negative resistance in the frequency band of interest as shown in Fig. 10.

III. EXPERIMENTAL RESULTS

The die-microphotograph of the proposed VCO implemented in Tower Jazz 0.18 µm SiGe BiCMOS process with \( f_T \) of 200 GHz is shown in Fig. 11. The VCO and buffer consume an area of 465 µm × 598 µm excluding the measurement pads. The triple-coupled coils are implemented with available top metal thickness of 2.81 µm and \( \kappa \) is around 0.65 (upper limit is process dependent). With a dc supply voltage of 1.8 V, the VCO core consumes 8.2 mW and the buffer consumes 5.4 mW. The VCO
is measured on-wafer varying the fine tuning voltage \( v_{tune} \) and coarse tuning bits \( B<0> - B<3> \) by digitally controlling \( v_{t1} \) and \( v_{t2} \) to 0 V or 1.8 V. As shown in Fig. 12, the proposed VCO covers frequencies from 22.50 GHz to 26.23 GHz, a frequency tuning range of 15.309%. Excluding the simulated buffer loss of 3.2 dB, the VCO measures an average single-end output power of -3.1 dBm for varying \( v_{t1} \) and \( v_{t2} \), as shown in Fig. 12. The phase noise of the proposed VCO across the operational frequency is shown in Fig. 13. The VCO achieves a lowest phase noise of \(-107.7 \text{ dBc}\/\text{Hz}\) at 1 MHz and \(-132 \text{ dBc}\/\text{Hz}\) at 10 MHz offset from 24 GHz carrier as shown in Fig. 13. A comprehensive summary and comparison with VCOs in the same frequency range is shown in Table I. For a fair comparison, the SiGe VCO in Table I have a similar performance with strong coupling among the triple-coupled LC tanks, the proposed VCO achieves a good FOM with an optimized trade-off among wide tuning range, low phase noise, high output power and PAE simultaneously.

### IV. CONCLUSION

In this work, a fully differential K-band VCO using the triple-coupled LC tanks is presented. The strong coupling of triple-coupled LC tanks in the proposed VCO is analyzed theoretically and verified experimentally. It is demonstrated that with strong coupling among the triple-coupled LC tanks, the proposed VCO achieves a good FOM with an optimized trade-off among wide tuning range, low phase noise, high output power and PAE simultaneously.

### ACKNOWLEDGMENT

The authors are thankful to Tower Jazz Semiconductors for their support on the fabrication.

### REFERENCES


### TABLE I

<table>
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<th>Ref.</th>
<th>( f_{om} ) (GHz)</th>
<th>( L_{1}[\text{f}(0.10 \text{dBm})] ) (dB/Hz)</th>
<th>( L_{1}[\text{f}(1.0 \text{dBm})] ) (dB/Hz)</th>
<th>( K_{VCO} ) (MHz/V)</th>
<th>( P_{dc} ) (mW)</th>
<th>( \text{FTR} ) (%)</th>
<th>( P_{out} ) (dBm)</th>
<th>( \text{PAE} ) (%)</th>
<th>( \text{FOM}^{1} ) (dBc/Hz)</th>
<th>Area (mm²)</th>
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<td>This Work</td>
<td>22.5 – 26.23</td>
<td>-75</td>
<td>-107.7</td>
<td>550</td>
<td>8.2</td>
<td>15.309</td>
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<td>0.465 × 0.6(^4)</td>
<td>0.18 μm SiGe</td>
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<td>[2]</td>
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<td>-65</td>
<td>-116.4</td>
<td>900</td>
<td>3</td>
<td>4.808</td>
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<td>1600</td>
<td>1.4</td>
<td>7.018</td>
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<td>550</td>
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<td>-80</td>
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<td>10</td>
<td>13.102</td>
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<td>1.731</td>
<td>-158.738</td>
<td>0.38 × 0.63</td>
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</table>

\[ FOM^{1} = \frac{P_{dc}}{P_{out} \times 10^{(10 \log \text{FTR} – 10 \log \text{PAE} – \text{FTR}) / 10}} \]

\(^1\) FOM is calculated with phase noise at 1 MHz offset.