A Low Phase Noise and Wide Tuning Range Millimeter-Wave VCO Using Switchable Coupled VCO-Cores

Qiong Zou, Student Member, IEEE, Kaixue Ma, Senior Member, IEEE, and Kiat Seng Yeo, Senior Member, IEEE

Abstract—This work presents a millimeter-wave (mm-wave) dual-mode voltage-controlled oscillator (VCO) topology with switchable coupled VCO-cores for wide frequency tuning range and low phase noise application. By taking advantage of the different parasitic capacitance of cross-coupled pair when the VCO-core operates in ON and OFF states, the dual-mode operation of VCO can be realized, and the oscillations for both modes can be excited at the lower resonant frequency of tank, such that tank $Q$ and phase noise performance could be improved for both modes. Strongly coupled transformer with large coupling coefficient ($k$) is utilized to increase the oscillation stability at the desired resonant frequency for both modes. The large $k$ transformer will also facilitate the enhancement of tank $Q$ at the lower resonant frequency. Frequency tuning range of the VCO is increased by properly designing the VCO-cores and combining the frequency bands of the two modes. In addition, the cross-coupled pair of VCO-core at OFF state is able to act as high $Q$ active capacitor, which can further increase the tank $Q$ and thus reduce the phase noise. Fabricated in a 0.18 $\mu m$ BiCMOS process, the VCO exhibits a wide tuning range of 17.2% from 55.7 GHz to 66 GHz, and low phase noise from $-87.5$ dBC/Hz to $-93.5$ dBC/Hz at 1 MHz offset over the entire tuning range.

Index Terms—Active capacitor, coupled VCO-cores, low phase noise, millimeter-wave (mm-wave), voltage-controlled oscillators (VCO), wide frequency tuning range.

I. INTRODUCTION

The unlicensed 60 GHz band from 57 to 66 GHz backed by international standards such as IEEE802.15.3c, IEEE802.11ad, and ECMA-387 becomes one of the most exciting opportunities to develop the next generation gigabit-data-rate wireless terminals due to its large bandwidth. As a key building block of transceiver, voltage-controlled oscillator (VCO) operating at 60 GHz with wide tuning range and low phase noise is highly demanded [1]–[5].

Conventionally, wide tuning range and low phase noise of VCO can be obtained by employing a parallel combination of switched capacitor array for coarse frequency tuning and MOS varactors for fine frequency tuning [6]–[8]. However, when operation frequency increases to the millimeter-wave (mm-wave) region, the quality factor ($Q$) of $LC$-tank will be significantly reduced due to the low $Q$ of capacitor/varactor and the high loss incurred in switches. As a result, the phase noise of VCO will be degraded. On the other hand, to sustain oscillation at mm-wave, transistor size is increased which will inevitably increase the parasitic capacitance and further limit the tuning range.

Transformer-based coupled $LC$–tanks with two resonant frequencies can be used to realize dual-mode VCOs [9]–[12] and increase frequency tuning range [13]–[15]. In those dual-mode VCOs, the oscillation of one mode is excited at the lower resonant frequency of the tank, and the other mode is excited at the higher resonant frequency of the tank. This method can extend frequency tuning range, and when the loss of the tank capacitor is much smaller than that of the inductor, the tank $Q$ at lower resonant frequency can be enhanced compared to single $LC$-tank with the same component $Q$. However, the tank $Q$ at higher resonant frequency will be decreased [11], resulting the phase noise being improved for one mode but degraded for the other mode. What is worse, when the loss of the tank capacitor is comparable with loss of the inductor at mm-wave frequency, as will be shown in this work, the tank $Q$ at the lower resonant frequency can only be enhanced when the coupling coefficient ($k$) is larger than a certain value. But, at the same time, to ensure stable oscillation at higher resonant frequency of the tank, small $k$ of the transformer is required [16] (e.g., 0.2 in [14]), which will in return limit the enhancement or even degrade the tank $Q$ at the lower resonant frequency.

To achieve low phase noise and stable oscillation for both modes, a new dual-mode mm-wave VCO topology involving switchable coupled VCO-cores is proposed in this work. Taking advantage of different parasitic capacitance of cross-coupled pair as the corresponding VCO-core in ON and OFF states, the dual-mode operation in mm-wave frequency can be realized, and the oscillations for both modes are allowed to be excited at the lower resonant frequency of the tank, thus the tank $Q$ and phase noise performance of the VCO can be improved for both modes. Simultaneously, the stability of oscillation at desired resonant frequency can be improved for both modes by using large $k$ transformer, which in return facilitates the enhancement of tank $Q$ for both modes. Besides, the frequency tuning range of the VCO can be increased by properly designing the VCO-cores and combining the frequency bands of the two modes. Moreover, the cross-coupled pair of the VCO-core operating in the OFF state is able to act as high $Q$ active capacitor to further improve the tank $Q$ and phase noise performance of VCO. Using the proposed topology, a dual-mode mm-wave VCO with the wide tuning range of 17.2% and low phase noise from $-97.5$ dBC/Hz to $-92$ dBC/Hz at 1 MHz offset over the
entire tuning range has been successfully demonstrated in 0.18 μm SiGe BiCMOS process.

The paper is organized as follows. Section II presents the theoretical analysis of the proposed VCO including the operation mechanism, tank Q, oscillation stability, and high Q active capacitor. Section III gives the details of circuit design and implementation. The experimental results are shown in Section IV, and the conclusion is given in Section V.

II. ANALYSIS OF THE PROPOSED VCO

Fig. 1(a) shows the schematic of the proposed dual-mode VCO which is composed of two switchable coupled VCO-cores (i.e., Core1, Core2) and a frequency doubler. Bipolar transistors Q1 ~ Q4 form two cross-coupled pairs of the two VCO-cores. C_{o1} and C_{o2} are varactors for frequency tuning. I_{B1} and I_{B2} are the bias currents of Core1 and Core2. A triple-coils strongly coupled transformer is adopted to couple two VCO-cores and also couple the oscillated signal to the input of frequency doubler, such that the output of the two operation modes can be obtained at the same port. L1, L2, and L3 are self-inductance of the primary, secondary, and tertiary coils of transformer.

The two operation modes of VCO are denoted as Mode A and Mode B. By controlling I_{B1} and I_{B2}, Core1 is switched on and Core2 is switched off for Mode A, while Core1 is switched off and Core2 is switched on for Mode B. As shown in Fig. 1(b), supposing the frequency tuning range of VCO is from f_0 to f_1 for Mode A, and f_0 to f_2 for Mode B, when the frequency tuning ranges of Mode A and Mode B have been carefully designed to make f_0 < f_2 ≤ f_1 < f_3, a continuous wide frequency tuning range of VCO from f_0 to f_3 can be achieved.

A. Dual-Mode Operation With Switchable Coupled VCO-Cores

As shown in Fig. 2, when the bias current of the cross-coupled pair is set to I_{on} or I_{off}, the value of the input parasitic capacitance of the cross-coupled pair, C_{on} or C_{off} are different. Denoting C_{on1}, C_{off1}, and C_{on2}, C_{off2} as the input parasitic capacitance of the cross-coupled pair in Core1 and Core2 in their ON and OFF state, the equivalent circuits of VCO (without doubler) for Mode A and Mode B can be shown as in Fig. 3(a) and (b). Using C_1 and C_2 to represent the capacitance loaded on the primary and secondary coil of transformer, then there are C_1 = C_{on1} + C_{off1}, C_2 = C_{off2} + C_{on2} for Mode A, and C_1 = C_{on1} + C_{off1}, C_2 = C_{on2} + C_{off2} for Mode B. R_{c1A}, R_{c2A} and R_{c1B}, R_{c2B} represent the overall resistive loss of C_1, C_2 for Mode A and Mode B. C_3 represents the input parasitic capacitance of the doubler and R_{c3} is its parasitic resistance. For simplicity of calculation, the coupling coefficients between any two windings of transformer are assumed to be identical and labeled as k. R_1, R_2, and R_3 are the parasitic resistance of the transformer.

The equivalent network of tank impedance (Z_{inA} and Z_{inB}) for Mode A and Mode B are also shown in Fig. 3(a) and (b). Ignoring R_1 ~ R_3 and R_{c1A} ~ R_{c2A} ~ R_{c2B}, R_{c3}, the resonant frequencies can be calculated. In this work, C_3 is much smaller than C_1 and C_2, and we can assume C_3L_3ω^2 << 1, which is reasonable in our target frequency range. Then, L_{eq1} and L_{eq2} can be derived and simplified as

\[
L_{eq1} \approx \frac{k^2ω^2L_1L_2C_2}{ω^2L_1C_3 - 1},
\]

\[
L_{eq2} \approx \frac{k^2ω^2L_1L_2C_1}{ω^2L_1C_3 - 1}.
\]
Therefore, the resonant frequencies of tank for Mode A can be solved by \( \omega = 1/\sqrt{L_{eq}C_1} \), and the resonant frequencies of tank for Mode B can be solved by \( \omega = 1/\sqrt{L_{eq}C_2} \). It can be found that the tank resonant frequency expressions for Mode A are the same as the expressions for Mode B. For each mode, the resonant frequencies are given by

\[
\omega_{H/L} = \sqrt{\frac{L_1C_1 + L_2C_2 \pm \sqrt{(L_1C_1 - L_2C_2)^2 + 4k^2L_1C_1L_2C_2}}{2L_1C_1L_2C_2(1 - k^2)}}
\]

(3)

To take advantages of the \( Q \) enhancement of coupled \( LC \)-tanks, as will be discussed in Section II-B, the oscillations of VCO are excited at the lower resonant frequency \( \omega_L \) for both modes. Though the expression of \( \omega_L \) are the same for Mode A and Mode B, the parameters \( C_1 \) and \( C_2 \) vary as mode changes, resulting in different value of \( \omega_L \) in Mode A and Mode B. Without loss of generality, the capacitance ratio is defined as \( \alpha = C_1/C_T - 1 = C_2/C_T(0 < \alpha < 1) \), where \( C_T = C_1 + C_2 \) is the total capacitance of \( C_1 \) and \( C_2 \). When the operation mode of VCO changes from Mode A to Mode B, the capacitance ratio \( \alpha \) and total capacitance \( C_T \) would vary from \( \alpha_A, C_{T.A} \) to \( \alpha_B, C_{T.B} \). Substituting \( \alpha \) and \( C_T \) to (3), the expression of \( \omega_L \) can be rewritten as a function of \( \alpha \) and \( C_T \), which is given by

\[
\omega(\alpha, C_T) = \sqrt{\frac{\alpha + n(1 - \alpha) - \sqrt{(\alpha + n + \alpha n)^2 + 4n\alpha(1 - \alpha)k^2}}{2(1 - k^2)\alpha n(1 - \alpha)L_1C_T}}
\]

(4)

where the \( n \) is inductance ratio \( n = L_2/L_1 \). Thereby, the operation frequencies for Mode A and Mode B can be expressed as \( \omega(\alpha_A, C_{T.A}) \) and \( \omega(\alpha_B, C_{T.B}) \). Assuming \( \omega(\alpha_A, C_{T.A}) < \omega(\alpha_B, C_{T.B}) \), if \( \Delta f = (\omega(\alpha_B, C_{T.B}) - \omega(\alpha_A, C_{T.A})) \) is properly designed, the frequency tuning ranges of the two modes are able to form a continuous wide frequency tuning range [Fig. 1(b)]. In order to obtain suitable \( \Delta f \), the design parameters \( \alpha_A, \alpha_B, C_{T.A}, \) and \( C_{T.B} \) need to be carefully chosen. In addition to the requirement of \( \Delta f \), the range of \( \alpha, \alpha_B, C_{T.A}, \) and \( C_{T.B} \) can be restrained by optimized power consumption and requirement of \( C_{x1} \) and \( C_{x2} \). Besides, the selection of \( \alpha_A \) and \( \alpha_B \) need consider the tank \( Q \) and stability of oscillation. The detail of parameter selections will be discussed in Section III-A.

The proposed VCO takes advantages of parasitic capacitance variation of cross-coupled pair at ON and OFF states to change the frequency band. At mm-wave frequency, the required tank capacitance is relatively small and the variation of parasitic capacitance of cross-coupled pair can be comparable with other capacitors in LC tank, therefore its effect on frequency would be more obvious. By controlling the bias current of cross-coupled pair, the mode switching between Mode A and Mode B is achieved without using any switches directly connected to \( LC \)-tank, avoiding the degradation of tank \( Q \) from lossy switches. In addition, for both modes, the oscillation of VCO is at the lower resonant frequency of the tank, as will be shown in the following, the strongly coupled transformer in the proposed VCO will not only better enhance the tank \( Q \) of modes but also mandate more stable oscillation on the desired resonant frequency of tank.

**B. Tank Q**

The tank \( Q \) at the resonant frequency \( \omega_L \) and \( \omega_H \) can be calculated using the open loop quality factor definition from Razavi in [17],

\[
Q(\omega_H/L) = \frac{\omega_{H/L}}{\frac{d \varphi_{A/B}}{d \omega} \mid_{\omega = \omega_{H/L}}} = \frac{\omega_{H/L}Q_{A/B}(2)}{d \varphi_{A/B} / d \omega} \mid_{\omega = \omega_{H/L}}
\]

(5)

where \( \varphi_{A/B} \) denotes the phase shift of tank impedance \( Z_{in,A} \) or \( Z_{in,B} \). Based on the equivalent network of \( Z_{in,A} \) or \( Z_{in,B} \) in Fig. 3(a) and (b), the tank \( Q \) for Mode A and Mode B can be calculated respectively. Since \( C_3 \) is much smaller than the \( C_1 \) and \( C_2 \), the effects of \( R_5 \) and \( R_4 \) on tank \( Q \) are negligible. Then, the expression of equivalent inductance \( L_{eq1} \) and \( L_{eq2} \) considering parasitic resistance are derived as

\[
L_{eq1} = L_1 + \frac{C_2k^2L_2C_2\omega^2(1 - C_2L_2\omega^2)}{(1 - C_2L_2\omega^2)^2 + C_2^2\omega^2(R_2 + R_{c2A})^2}
\]

(6)

\[
L_{eq2} = L_2 + \frac{C_2k^2L_1C_2\omega^2}{(1 - C_1L_1\omega^2)^2 + C_1\omega^2(R_1 + R_{c1B})^2}
\]

(7)

and the equivalent resistance \( R_{eq1} \) and \( R_{eq2} \) are given by

\[
R_{eq1} = R_1 + \frac{C_2^2k^2L_2L_1\omega^4(R_2 + R_{c2A})}{(1 - C_2L_2\omega^2)^2 + C_2^2\omega^2(R_2 + R_{c2A})^2}
\]

(8)

\[
R_{eq2} = R_2 + \frac{C_2^2k^2L_1L_2\omega^4(R_1 + R_{c1B})}{(1 - C_1L_1\omega^2)^2 + C_1\omega^2(R_1 + R_{c1B})^2}
\]

(9)

Then, the phase shift of \( Z_{in,A} \) or \( Z_{in,B} \) can be calculated and expressed as

\[
\varphi_A = \tan^{-1} \left( \frac{\omega(L_{eq1} - C_1R_{eq1}^2 - C_1L_2^2\omega^2 + C_1^2L_{eq1}^2R_{c2A}^2)}{R_{eq1} + C_1^2R_{c1A}R_{eq1}\omega^2(R_{eq1} + R_{c1A}) + C_1^2L_{eq1}^2R_{c1A}\omega^4} \right)
\]

\[
\varphi_B = \tan^{-1} \left( \frac{\omega(L_{eq2} - C_2R_{eq2}^2 - C_2L_2^2\omega^2 + C_2^2L_{eq2}^2R_{c2B}^2)}{R_{eq2} + C_2^2R_{c2B}R_{eq2}\omega^2(R_{eq2} + R_{c2B}) + C_2^2L_{eq2}^2R_{c2B}\omega^4} \right)
\]

(10)

Fig. 4(a)–(d) plot the calculated and simulated tank \( Q \) at resonant frequencies \( \omega_L \) and \( \omega_H \) for Mode A and Mode B with a set of normal tank parameters at target frequency range: \( L_1 = 140 \) pH, \( C_T = 230 \) fF, \( L_3 = 126 \) nH, \( C_3 = 20 \) fF, \( \Delta f = 3Q_L = 20 \), \( \varphi_{L1} = \varphi_{Q_2} = \varphi_{Q_3} = 15 \) at 31 GHz, and \( n = 0.9 \). For comparison, \( Q \) of conventional single \( LC \)-tank with same component \( Q \) (i.e. \( Q = 20 \), \( Q = 15 \) at 31 GHz is plotted in dashed lines. The trend of calculated tank \( Q \) matches well with simulated tank \( Q \) at the target resonant frequency \( \omega_{H/L} \) which validates the derivation. The discrepancy between calculated and simulated tank \( Q \) at \( \omega_{H/L} \) is mainly due to the approximation procedure during the deduction of resonant frequencies. Compared to conventional single \( LC \)-tank, the tank \( Q \) at lower resonant frequency \( Q_A(\omega_L) \) and \( Q_B(\omega_L) \) can be both enhanced when \( k \) and \( \alpha \) are properly selected, while the tank \( Q \) at the higher resonant frequency \( Q_A(\omega_H) \) and \( Q_B(\omega_H) \) are always degraded. Since the target resonant frequency for both Mode A and Mode B of the proposed VCO is the lower resonant frequency \( \omega_L \) of tank, the tank \( Q \) and phase noise of VCO can be improved for both of modes. Furthermore, for diverse values of \( \alpha, \alpha_A(\omega_L) \) and \( \alpha_B(\omega_L) \) are all increased with the increasing of \( k \), therefore the strongly coupled transformer with large \( k \) in the proposed VCO is able to better enhance the tank \( Q \) at the
C. Oscillation Stability

For a VCO with transformer-based LC-tanks which has more than one resonant frequency, it needs to avoid the potential stability problem that the oscillation could jump from one equilibrium oscillation frequency to the other with some disturbance [11]. Stable oscillation at only the target frequency is highly desired.

As shown in Fig. 5, the open loop transfer functions of the proposed VCO are \( H(s) = g_{m1}Z_{in,A} \) and \( H(s) = g_{m2}Z_{in,B} \) for Mode A and Mode B. To start-up oscillation, the barkhausen criteria should be satisfied which is the open loop gain \( |H(s)| = |g_{m1}/2Z_{in,A}| > 1 \) and \( |H(s)| = 0 \) or 180° [18]. In this work, \( |H(s)| = 0 \) at both \( \omega_I \) and \( \omega_H \). Therefore, the magnitude of tank impedance \( |Z_{in,A}| \) will dominate the start-up of oscillation. According to [16], one necessary condition for the oscillation only occurring at \( \omega_L \) is that the magnitude peak at \( \omega_L \) is larger than that at \( \omega_H \). Specifically, it is \( |Z_{in,A}(\omega_L)| > |Z_{in,A}(\omega_H)| \) for Mode A and \( |Z_{in,B}(\omega_L)| > |Z_{in,B}(\omega_H)| \) for Mode B. If the oscillation is free to choose equilibrium point, when \( |Z_{in,A}(\omega_L)| > |Z_{in,A}(\omega_H)| \) and \( |Z_{in,B}(\omega_L)| > |Z_{in,B}(\omega_H)| \), the oscillations for Mode A and Mode B are more probable at \( \omega_L \) rather than at \( \omega_H \). In another word, large ratio of the impedance magnitude peaks at \( \omega_L \) and \( \omega_H \) can help to increase oscillation stability at \( \omega_I \).

Fig. 6(a) and (b) plot the ratios of tank impedance magnitude peaks \( |Z_{in,A}(\omega_L)|/|Z_{in,A}(\omega_H)| \) and \( |Z_{in,B}(\omega_L)|/|Z_{in,B}(\omega_H)| \) in 10log scale, using the same component values as in Fig. 4. Apparently, both \( |Z_{in,A}(\omega_L)|/|Z_{in,A}(\omega_H)| \) and \( |Z_{in,B}(\omega_L)|/|Z_{in,B}(\omega_H)| \) are increased with the increasing of \( k \), which implies that transformer with large \( k \) is able to facilitate the stability of oscillation for both modes at the lower resonant frequency \( \omega_L \).

For Mode A, \( |Z_{in,A}(\omega_L)|/|Z_{in,A}(\omega_H)| \) is increased with the increasing of \( k \) for a fixed \( \alpha \), and \( |Z_{in,B}(\omega_L)|/|Z_{in,B}(\omega_H)| \) for all \( k \) values are always larger than 1 after \( \alpha \). On the contrary, for Mode B, \( |Z_{in,B}(\omega_L)|/|Z_{in,B}(\omega_H)| \) is reduced with the increasing of \( \alpha \), and when \( \alpha < 0.5 \), \( |Z_{in,B}(\omega_L)|/|Z_{in,B}(\omega_H)| \) are still larger than 1. Although both \( k \) and \( \alpha \) could affect the ratios of tank impedance magnitude so as to influence the stability of oscillations, since the different values of \( \alpha \) in Mode A and Mode B are related to the design of cross-coupledpair, it is better to reserve more freedoms for the choice of \( \alpha \) by adopting large \( k \). For instance, suppose the requirements of tank impedance ratios for stable oscillation at \( \omega_L \) are \( |Z_{in,A}(\omega_L)| > 4|Z_{in,A}(\omega_H)| \) and \( |Z_{in,B}(\omega_L)| > 4|Z_{in,B}(\omega_H)| \), if \( k = 0.2 \), it requires \( 0.6 < \alpha < 1 \) for Mode A and \( 0 < \alpha < 0.4 \) for Mode B, however, if \( k = 0.8 \), the range of \( \alpha \) for Mode A and Mode B are enlarged to \( 0.3 < \alpha < 1 \) and \( 0 < \alpha < 0.6 \) respectively.

In previous transformer-based dual-mode VCOs [9], [11]–[15], to cope with the potential stability problem calls for low \( k \) transformer, however, low \( k \) would limit the enhancement or even degrade the tank quality \( Q \) at lower resonant frequency of tank. Differently, in the proposed VCO, the strongly coupled transformer with large \( k \) is able to simultaneously increase the
oscillation stability and tank Q at the target resonant frequency for both modes [Fig. 4, Fig. 6].

D. High Q Active Capacitor

Active capacitor which provides negative input series resistance was introduced into active filter designs to compensate the loss of other components [19], [20]. High Q active capacitor based on cross-coupled pair was reported in [21]. In the proposed VCO, the special configuration of switchable coupled VCO-cores and alternative operation of two cores not only provide dual-mode operation but also enable the utilization of active capacitor. When one VCO-core is switched on to sustain oscillation, the cross-coupled pair of the switched off VCO-core is able to act as a high Q active capacitor. Different from the active capacitor used in filter, the current consumption of active capacitor in this work is reduced by biasing the cross-coupled pair in the region where the input conductance is still positive and in the vicinity of zero. In that case, the Q of active capacitor can be greatly enhanced while consuming very small current.

Based on the transistor model of BJT shown in Fig. 7, the input admittance of cross-coupled pair \( Y_{in,coupl} \) can be given by

\[
Y_{in,coupl} = \frac{1}{2\pi} + \frac{4Z_{m} + Z_{g} + r_{b} + g_{m}r_{b}Z_{g} - g_{m}Z_{g}Z_{m}}{2(r_{b}Z_{g} + r_{b}Z_{m} + Z_{g}Z_{m})} \tag{11}
\]

where \( Z_{g} = r_{g}(1 + j\omega C_{jg}) \) and \( Z_{m} = 1/j\omega C_{jm} \). \( C_{jg} \) is the junction capacitance between base and collector terminals, and \( C_{jm} \) is the junction capacitance between base and emitter terminal. With \( Y_{in,coupl} = G_{in,coupl} + j\omega C_{in,coupl} \), the input conductance \( G_{in,coupl} \) and input parasitic capacitance \( C_{in,coupl} \) are given by

\[
G_{in,coupl} = Re[Y_{in,coupl}] \tag{12}
\]

\[
C_{in,coupl} = \frac{Im[Y_{in,coupl}]}{\omega} \tag{13}
\]

The Q of \( C_{in,coupl} \) is the absolute value of image part of \( Y_{in,coupl} \) divided by the real part of \( Y_{in,coupl} \), which is given by

\[
Q_{in,coupl} = \frac{\omega C_{in,coupl}}{G_{in,coupl}} \tag{14}
\]

It is known that \( g_{m} \) and bias current \( (I_{c}) \) of BJT transistor have the relationship of \( g_{m} = qI_{c}/KT \), where \( KT/q = 26mV \) at 300 K. The \( I_{c} \) for \( G_{in,coupl} = 0 \) can be solved as

\[
I_{c} = \frac{KT}{q} \left( \frac{(r_{b}r_{f})(r_{b} + r_{f} + r_{e})}{r_{0}r_{f}} - C_{jg}(C_{jg} + C_{jm})r_{b}\omega^{2} \right) \frac{r_{b} \omega^{2}}{K_{T}} \frac{r_{b} r_{f} + \left( C_{jg} + r_{f} \right) r_{b} r_{f} + \left( C_{jg} + C_{jm} \right) r_{b} r_{f} + \left( C_{jg} + C_{jm} \right) r_{b} r_{f} \omega^{2}}{q} \tag{15}
\]

When \( I_{c} \rightarrow I_{co} \) and \( I_{c} < I_{co}, G_{in,coupl} \) can be reduced and approaches to zero. In that case, \( Q_{in,coupl} \) can be greatly enhanced, and the cross-coupled pair acts as a high Q active capacitor. Fig. 8 shows the simulated \( Q_{in,coupl} \) and \( C_{in,coupl} \) when \( I_{c} \) is increased from 0 to 0.5 mA at 30 GHz, where the emitter length of BJT transistor is 6 \( \mu \)m. The high Q value of more than 250 can be obtained when \( I_{c} = 270 \mu A \), and the input capacitance \( C_{in,coupl} \) for \( I_{c} = 270 \mu A \) is only changed by 2 fF (from 32.6 to 34.6 fF) compared to \( I_{c} = 0 \).

Therefore, to get high Q of active capacitor, the bias current of the VCO-core in OFF state \( (I_{off}) \) can be increased from 0 to 2\( I_{co} \). Since the change of capacitance as \( I_{off} \) increases from 0 to 2\( I_{co} \) is negligible, the VCO operation frequency is nearly the same with that when the transistor is completely off. Thus, we still denote this state as OFF state. The active capacitor is in parallel connection with the LC tank, thus the increased Q of active capacitor can help to enhance the tank Q and also improve the phase noise performance of VCO.

III. CIRCUIT DESIGN AND IMPLEMENTATION

From the above analysis, it can be seen that selecting proper values for \( \alpha_{A}, \alpha_{B}, C_{T,A}, C_{T,B}, \) and \( k \) is essential to design a VCO with the proposed topology. The next step involves the design of the triple-coils transformer with proper coupling coefficient. Then, the cross-coupled pairs, such as the size, bias current can be designed and optimized with certain requirement of current density.

A. Considerations of Parameters Selection

Based on Fig. 3, the equations about tank capacitors and design parameters can be given by

\[
C_{v1} + C_{e1} + C_{v2} + C_{off2} - C_{T,A} \tag{16a}
\]

\[
C_{v1} + C_{e1} - \alpha_{A} C_{T,A} \tag{16b}
\]

\[
C_{v1} + C_{off1} + C_{v2} + C_{on2} - C_{T,B} \tag{17a}
\]

\[
C_{v1} + C_{off1} - \alpha_{B} C_{T,B} \tag{17b}
\]

We also can define the difference of \( C_{on} \) and \( C_{off} \) as \( \Delta C_{p1} = C_{off1} - C_{on} \) and \( \Delta C_{p2} = C_{off2} - C_{on} \). From (18) and (19), we can get

\[
\Delta C_{p1} - C_{T,A} - C_{T,B} \tag{18}
\]

\[
\Delta C_{p1} = C_{T,A} - C_{T,B} \tag{19}
\]

Therefore, if \( \alpha_{A}, \alpha_{B}, C_{T,A}, C_{T,B} \) are known, \( \Delta C_{p1} \) and \( \Delta C_{p2} \) can be calculated through (18) and (19). Then, the transistors of cross-coupled pairs and bias current can be designed according to \( \Delta C_{p1} \) and \( \Delta C_{p2} \). First of all, the values of design parameters \( \alpha_{A}, \alpha_{B}, C_{T,A}, C_{T,B} \) need to satisfy requirement of \( \Delta f \), where \( \Delta f = \left| f(\alpha_{B}, C_{T,B}) - f(\alpha_{A}, C_{T,A}) \right| \). Then, since \( \Delta C_{p1} \) and \( \Delta C_{p2} \) are determined by bias current of cross-coupled pair, the ranges of \( \alpha_{A}, \alpha_{B}, C_{T,A}, C_{T,B} \) can be restrained for optimized power consumption. Finally, these ranges...
can be further restrained to satisfy the requirement of $C_{v1}$ and $C_{v2}$.

To cover the 60 GHz unlicensed band which is from 57 GHz to 66 GHz, the oscillation frequency range of the VCO before the frequency doubler should be from 28.5 to 33 GHz. Suppose the frequency tuning ranges of the two modes are equal, then it will require $\Delta f \geq 22.5$ GHz, and $f(\alpha_A, C_{T,A})$ needs to be smaller than 29.625 GHz, $f(\alpha_B, C_{T,B})$ need to be larger than 31.875 GHz. In our design, we make our targets as $f(\alpha_A, C_{T,A}) = 29.5$ GHz, $f(\alpha_B, C_{T,B}) = 32$ GHz to provide more tolerance to the circuit against the PVT variation.

Based on (19), Fig. 9 plots the available $\alpha$ and $C_T$ satisfying $f(\alpha, C_T) = 32$ GHz and $f(\alpha, C_T) = 29.5$ GHz, where the required $\alpha_A$, $C_{T,A}$ can be chosen from the lines of $f = 29.5$ GHz, and $\alpha_B$, $C_{T,B}$ can be chosen from the lines of $f = 32$ GHz. With different $k$ of transformer, those lines are different, thus the range of $k$ should be specified firstly. Although large $k$ of transformer facilitate enhancement of tank Q and stable oscillation at target resonant frequency, the selection of $k$ also needs to consider $C_{T,A}$ and $C_{T,B}$. Since $\Delta C_{p1} - \Delta C_{p2} = C_{T,A} - C_{T,B}$, the increased value of $C_{T,A} - C_{T,B}$ implying larger parasitic capacitance of cross-coupled pair will be needed, thus requires more dc current and larger size transistor. Therefore, $C_{T,A} - C_{T,B}$ should be minimized. As shown in Fig. 9, as $k$ increases from 0.5 to 0.8, the lines of $f = 32$ GHz and $f = 29.5$ GHz both gradually become flat, resulting in decreased of maximum value of $C_{T,B}$. Since the minimum value of $C_{T,A}$ is not changed, the minimum value of $C_{T,A} - C_{T,B}$ will be increased. It can be seen in Fig. 9, $\min(C_{T,A} - C_{T,B}) < 0$ when $k = 0.7$, but $\min(C_{T,A} - C_{T,B}) > 0$ when $k = 0.8$. Therefore, in this case, $k$ needs to be smaller than 0.8.

For the selection of $\alpha_A$ and $\alpha_B$, the stability issue should be considered. Referring to Section II-C, the choosing of $\alpha$ will affect the ratios of tank impedance magnitude, and thus influence oscillation stability. If $k = 0.7$, then $0.4 < \alpha_A < 1$ and $0 < \alpha_B < 0.5$ are needed to make the ratios of tank impedance at $\omega_L$ to $\omega_H$ larger than 4. Besides, the range of $\alpha_A$ and $\alpha_B$ can be further restrained by $C_{v1}$ and $C_{v2}$. Suppose $C_{v1}$ and $C_{v2}$ need to be at least larger than 30 fF, which is a small capacitor value for this frequency, then from (18) to (19), we can get

$$\begin{align*}
\alpha_A C_{T,A} &> 30 fF \quad C_{v1} > 30 fF \\
(1 - \alpha_A) C_{T,A} &> 30 fF \quad C_{v1} > 30 fF \\
\alpha_B C_{T,B} &> 30 fF \quad C_{v1} > 30 fF \\
(1 - \alpha_B) C_{T,B} &> 30 fF \quad C_{v1} > 30 fF
\end{align*}$$

(20a) (20b) (21a) (21b)

For $k = 0.7, 0.4 < \alpha_A < 1, 0 < \alpha_B < 0.5$, minimum value of $C_{T,A}$ is 210 fF, and $C_{T,B}$ is 200 fF. Based on (20), (21), the ranges of $\alpha_A$ and $\alpha_B$ for smallest $C_{v1}$ and $C_{v2}$ are $0.14 < \alpha_A < 0.86$ and $0.15 < \alpha_B < 0.85$ with $C_{T,A} = 210$ fF and $C_{T,B} = 200$ fF. Therefore, by considering the requirement of $C_{v1}$ and $C_{v2}$ (e.g. $30$ fF), the ranges of $\alpha_A$ and $\alpha_B$ shrink to $0.4 < \alpha_A < 0.86$ and $0.15 < \alpha_B < 0.5$. In addition, another rule for choosing $\alpha_A$ and $\alpha_B$ is that $\alpha_A - \alpha_B$ needs to be as small as possible in order to minimize $\Delta C_{p1}$, and thus saving power consumption (19).

B. Triple-Coils Transformer

In this work, concentric coupling method is used for the coupling between primary and secondary coil of transformer for its moderate coupling coefficient compared to other coupling methods, such as stacked coupling and inter-wound coupling [22]. Moreover, the coupling coefficient of concentric coupled transformer can be easily optimized by varying the distance between metal traces.

The front view and side view of the triple-coils transformer is shown in Fig. 10. $L_1$ and $L_2$ are concentrically coupled using the top metal layer to keep far away from the lossy silicon substrate. Since the oscillation signals of the VCO-cores are coupled to the doubler by $L_3$, it is better to make the coupling strength between $L_1$ and $L_3$ same with that between $L_2$ and $L_3$. Thus, $L_3$ is stacked under the gap between $L_1$ and $L_2$. This structure also can reduce the coupling capacitance brought by stacked coupling, such that it increases the self-resonant frequency of transformer.

The outer dimension of transformer is 142 $\mu m$, and the metal trace widths of $L_1$ and $L_2$ are 12.5 $\mu m$ and 11.5 $\mu m$ respectively. The distance between metal trace of $L_1$ and $L_2$ is 2.3 $\mu m$. The simulated self-inductance for $L_1$, $L_2$, $L_3$ are 140 pH, 131 pH, and 137 pH. The coupling coefficient $k_{12}$, $k_{13}$, and $k_{23}$ are 0.59, 0.69, and 0.66 respectively. Although $k_{12}$, $k_{13}$, and $k_{23}$ are not identical, which is different from the assumption in deduction, the major coupling coefficient influencing the oscillation frequency is $k_{12}$. Assuming those coupling coefficients as identical simplifies the deduction and confers insight to circuit mechanism.

C. Design of Cross-Coupled Pairs

The size and bias current of transistors can be optimized for a designed triple-coils transformer. When $k = 0.59, n = 0.93$, the relationship between $C_T$ and $\alpha$ for $f = 32$ GHz and $f = 29.5$ GHz is plotted in Fig. 11. Based on the guidelines for parameters selection in Section III-A, the ranges for $\alpha_A$ and $\alpha_B$, where $0.5 < \alpha_A < 0.86$ and $0.16 < \alpha_B < 0.5$, are marked. We can choose $\alpha_A = 0.6$ and $\alpha_B = 0.45$ as a good trade-off between small $\alpha_A - \alpha_B$ and $C_{T,A} - C_{T,B}$, because $C_{T,A}$ is decreased with the increasing of $\alpha_A$ at the target range, $\alpha_A = 0.6$ would be a moderate value. Meanwhile, $\alpha_B = 0.45$ corresponds to the peak value of $C_{T,B}$, such that both $\alpha_A - \alpha_B$ and $C_{T,A} - C_{T,B}$ can be very small. Then, from $\alpha_A = 0.6$ and $\alpha_B = 0.45, C_{T,A} = 264$ fF and $C_{T,B} = 231$ fF are obtained.

![Fig. 10. Structure of the transformer.](Image 77x613 to 251x726)
Substituting $\alpha_A$, $\alpha_B$, $C_{T,A}$, and $C_{T,B}$ to (18) and (19), there are $\Delta C_{p1} = 54$ fF, $\Delta C_{p2} = 21$ fF.

$\Delta C_{p1}$ and $\Delta C_{p2}$ are determined by the transistors size and bias current. Generally, for mm-wave VCO design, a proper current density $I_C/L_E$ of transistors need be chosen for a good $f_T$ and minimum noise figure (NF$_{min}$). Fig. 12 shows $f_T$ and NF$_{min}$ versus current density for the target process. The current density for best $f_T$ is $I_C/L_E = 1.4$ mA/μm, and for best NF$_{min}$ is $I_C/L_E = 0.4$ mA/μm. Thus, to get a good trade-off between $f_T$ and NF$_{min}$, $I_C/L_E = 1$ mA/μm and $I_C/L_E = 0.8$ mA/μm for the cross-coupled pairs in Core1 and Core2 are selected, where the NF$_{min} < 2.4$ dB, and $f_T > 165$ GHz.

Fig. 13(a) plots the contours of $\Delta C_p$ with the changing of bias current ($I_c$) and emitter length ($L_E$) of BJTs, where the line of current density $I_C/L_E = 1$ mA/μm and $I_C/L_E = 0.8$ mA/μm are marked. Fig. 13(b) shows $C_{off}$ with the changing of $L_E$. Thereby, from junction between $\Delta C_{p1} = 54$ fF and $I_C/L_E = 1$ mA/μm, along with $\Delta C_{p2} = 21$ fF and $I_C/L_E = 0.8$ mA/μm, the required $I_C = 6.4$ mA, $L_E = 6.4$ μm, $I_C = 3.1$ mA, and $L_E = 3.9$ μm can be obtained. Then, with $L_{E1}$ and $L_{E2}$, the corresponding $C_{off1} = 34$ fF and $C_{off2} = 22$ fF can be found from Fig. 13(b). Therefore, $C_{off1}$ and $C_{off2}$ can be solved as $C_{off1} = 70$ fF and $C_{off2} = 84$ fF by (18) and (19). It should be noted that parasitic capacitance of transformer is loaded on tank, and it can be absorbed into $C_{off1}$ and $C_{off2}$. The varactors with multi-fingers and minimum gate width, length are utilized in this design, and $C_{off}$ is formed by a MIM-cap in parallel connection with a MOS varactor to enhance the quality factor. With implemented tank, the simulated impedance magnitude ratios at $\omega_H$ and $\omega_L$ are $Z_{in,A}(\omega_L)/Z_{in,A}(\omega_H) = 8$, $Z_{in,B}(\omega_L)/Z_{in,B}(\omega_H) = 5$.

IV. MEASUREMENT RESULTS

The proposed VCO is implemented in a 0.18 μm SiGe BiCMOS process with six metal layers. The chip photomicrograph of VCO is shown in Fig. 14. Shielding is used under the transformer to enhance the isolation to substrate. The total chip size including the I/O pads is $0.503 \times 0.69$ mm$^2$. The bias current source of the cross-coupled pair is realized by MOS transistors to reduce required voltage headroom. With a supply voltage of 1.2 V, the two VCO-cores consumes 12.8 mA to 12.95 mA current for Mode A, and 6.24 mA to 6.39 mA for Mode B. The power consumption of the frequency doubler is 3.6 mW.

The measured and simulated oscillation frequency with the variation of tuning voltage ($V_T$) for Mode A and Mode B is shown in Fig. 15. The frequency tuning range of VCO is 10.3 GHz (17%), from 55.7 GHz to 66 GHz for Mode B, and from 59.9 GHz to 66 GHz for Mode B. Compared to simulated results, the measured oscillation frequency is shifted down by about 200 MHz (less than 0.4% over 60 GHz), and the measured frequency tuning range is increased by 500 MHz (4.85% of 10.3 GHz tuning range). This might be caused by inaccuracy of the process model or EM simulation of transformer.

The setup for phase noise measurement is shown in Fig. 16, where
two cascaded amplifiers V-LNA-50-75-4 are adopted to amplify the output signal and compensate the cable loss and harmonic mixer loss, and Signal Source Analyzer (SSA) E5052B is used to measure the phase noise. Fig. 17 shows the phase noise at 1 MHz offset versus the bias current of OFF state VCO-core. With \( I_{\text{off}} \) increased from 0 to 0.15 mA, the phase noise is reduced, and the improvement on phase noise is roughly increased with the increasing of 1. The reason should be that the \( Q \) of varactor at small \( V_T \) is much lower than that of active capacitance, so the low \( Q \) of varactor is the major factor to limit the phase noise at small \( V_T \). When \( V_T \) is increased, the \( Q \) of varactor is increased, thus the improvement of phase noise is also increased. Fig. 18 shows the measured phase noise and output power over the entire frequency tuning range when setting \( I_{\text{off}} = 0.15 \) mA. The measurement of output power is performed with a PNA-X network analyzer N5247A. The output power is from \(-32.8 \) dBm to \(-28.8 \) dBm over the frequency tuning range after calibration. The measured output power is low due to the frequency doubler. Using the frequency doubler can reduce VCO design frequency by half and allow better trade-off between tuning range and phase noise. Besides, the proposed VCO topology with frequency doubler omits 60 GHz frequency divider in PLL, which can significantly reduce the PLL design complexity. However, the frequency doubler has the drawback of large voltage conversion loss at mm-wave frequency due to Class-C operation [23]. In order to drive mixer/modulator in RF front-end, driven amplifier can be added.

The phase noise at 1 MHz is from \(-91.4 \) dBc/Hz to \(-93.45 \) dBc/Hz for Mode A, and from \(-87.6 \) dBc/Hz to \(-89.4 \) dBc/Hz for Mode B. The phase noise variation on the frequency tuning range for each mode is less than 2 dB. The worst and best phase noise over the whole frequency tuning range versus frequency offset are shown in Fig. 19(a) and (b). The overall performance of the proposed VCO is listed in Table I and compared to the recent mm-wave VCOs [1]–[3], [5], [8], [24]–[27]. To compare with these VCOs around 30 GHz [26], [27], the frequency doubler should be excluded. Then, the power consumption of our proposed work is 15.5 mW/7.6 mW for Mode A/B, and the phase noise would be 6 dB lower (i.e., \(-93.5 \sim -99.5 \) dBc/Hz). A commonly used figure of merit...
is calculated using the best phase noise case.

(1) Phase noise (PN) is at 10 MHz offset.

(2) FOMp is calculated using the best phase noise case.

(3) Average phase noise across the TR.

FOMT which takes into account phase noise and tuning range as well as power consumption is given as follows:

$$FOM_T = FOM_p - 20 \log \left( \frac{T/R \%}{10^3 f_m} \right) + 10 \log (P_{DC} \text{ mW})$$

where \(f_m\) is the frequency offset from carrier frequency. The proposed VCO achieve wide tuning range, and also competitive phase noise compared to other works in more advanced technologies, demonstrating the merits of the proposed topology. Although [1] and [3] have better FOMp, the variation of phase noise over the tuning range is much larger. Benefiting from the tank Q enhancement for both modes, good phase noise over the whole tuning range are achieved in our proposed work. Though the VCO output is single-ended at 60 GHz, we may use a balun to convert the single-ended output to a differential signal for transceivers requiring differential LOs. A passive balun based on transformer, or an active balun [28] which can also serve as a drive amplifier can be used.

V. CONCLUSION

In this paper, a new dual-mode VCO topology with switchable coupled VCO-cores are presented and successfully demonstrated. The proposed topology utilizes the parasitic capacitances of cross-coupled pair to realize dual-mode operation, making it particularly suitable for VCO design at mm-wave frequency. Compared to conventional transformer-based dual-mode VCOs, the proposed topology allows using large \(k\) transformer to simultaneously enhance tank \(Q\) and increase the oscillation stability for both modes. With the proposed VCO topology, better design compromise between wide frequency tuning range and low phase noise can be achieved.

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Qiong Zou received the B.Eng. degree in electronic science and technology from Huazhong University of Science and Technology (HUST), China, in 2008. She is currently working toward the Ph.D. degree in electrical and electronic engineering at Nanyang Technological University, Singapore. Her research interests include the RF and millimeter-wave front-end IC design.

Kaixue Ma (M’05–SM’09) received the B.E. and M.E. degrees from Northwestern Polytechnological Univ. (NWPU), China, and the Ph.D. degree from Nanyang Technological Univ. (NTU), Singapore. From August 1997 to December 2002, he was with China Academy of Space Technology (Xi’an), where he became Group Leader of millimeter-wave group for space-borne microwave and mm-wave components and subsystem for satellite payload and VSAT ground station. From September 2005 to September 2007, he was with MEDs Technologies as an R&D Manager. From September 2007 to March 2010, he was with ST Electronics (Satcom & Sensor Systems) as R&D Manager, Project Leader and Technique Management Committee of ST Electronics. In March 2010, he joined NTU as a Senior Research Fellow and Millimeter-wave RFIC team leader for 60GHz Flagship Chipset project. As a PI/Technique Leader, He did projects with funds more than $12 million (excluding projects done in China). He is a Senior Member of IEEE and His research interests include satellite communication, software defined radio, Microwave/MM-wave circuits and system using CMOS, MEMS, MMIC, and LTCC. He has eight patents, two patents pending, and authored/co-authored over 80 referable international journal and conference papers in the related area. He received best paper awards from IEEE SOCC2011, IEEE SOC Design Group Award, excellent paper award from International Conference on HSCD2010, chip design competition bronze award of ISIC2011. He is a reviewer of several international journals.

Kiat Seng Yeo received the B.Eng. (EE) and Ph.D. (EE) degrees from Nanyang Technological University (NTU), Singapore, in 1993 and 1996, respectively. Associate Provost (International Relations and Graduate Studies) at Singapore University of Technology and Design (SUTD) and Member of Board of Advisors of the Singapore Semiconductor Industry Association, Prof. Yeo is a widely known authority in low-power RF/mm-wave IC design and a recognized expert in CMOS technology. He has secured over $30 million of research funding from various funding agencies and the industry in the last 3 years. Before his new appointment at SUTD, he was Associate Chair (Research), Head of Division of Circuits and Systems and Founding Director of VIRTUS of the School of Electrical and Electronic Engineering at NTU. He has published 6 books, 5 book chapters, over 400 international top-tier refereed journal and conference papers and holds 35 patents. Dr Yeo served in the editorial board of IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES and hold/held key positions in many international conferences as Advisor, General Chair, Co-General Chair, and Technical Chair. He was awarded the Public Administration Medal (Bronze) on National Day 2009 by the President of the Republic of Singapore and was also awarded the distinguished Nanyang Alumni Award in 2009 for his outstanding contributions to the university and society.