A Millimeter-Wave CMOS *LC*-Tank VCO With an Admittance-Transforming Technique

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Abstract—A novel circuit topology suitable for millimeter-wave voltage-controlled oscillators (VCOs) is presented. With the admittance-transforming technique, the proposed VCO can operate at a frequency close to the $f_{\max}$ of the transistors while maintaining remarkable circuit performance in terms of phase noise, tuning range, and output power. Using a standard micrometer CMOS process, a *U*-band VCO is implemented for demonstration. The fabricated circuit exhibits a frequency tuning range of 1.1 GHz in the vicinity of 50 GHz. The measured output power and phase noise at 1-MHz offset are -11 dBm and -101 dBc/Hz, respectively. Operated at a supply voltage of 1.8 V, the VCO core consumes a dc power of 45 mW.

Index Terms—Admittance transforming, coplanar waveguide (CPW), *LC*-tank oscillators, millimeter wave, phase noise, quality (*Q*) factor, voltage-controlled oscillators (VCOs).

I. INTRODUCTION

RECENTLY, with the emerging wireless applications such as the broadband wireless local area network (WLAN) at 60 GHz, the automatic cruise control at 76–77 GHz, and the advanced image sensors at 94 GHz, the development of low-cost and high-performance monolithic microwave integrated circuits have attracted great attention. Due to the unparalleled advantages in both fabrication cost and high-level integration, CMOS technology appears to be well suited for the circuit implementations. Unfortunately, the restricted cutoff frequency f_T and maximum oscillation frequency $f_{\max}$ have impeded the evolution of RF frontends for millimeter-wave applications. In order to enhance the high-frequency performance of the CMOS RF circuits, various design strategies have been proposed over the past few years [1]–[3].

Being a crucial component in the RF frontends, the voltage-controlled oscillator (VCO) provides the required local oscillator (LO) signal for frequency conversion. Conventionally, the cross-coupled *LC*-tank topology is widely used for high-frequency designs [4]–[9]. However, as the operating frequency approaches to the $f_{\max}$ of the active devices, the realization of the *LC*-tank CMOS VCOs encounters great challenges due

Fig. 1. (a) Equivalent half circuit of the conventional *LC*-tank VCO. (b) Simplified model for circuit analysis.

to the inherently low quality (*Q*) factors of the on-chip passive elements, making it extremely difficult to meet the stringent startup conditions for sustained oscillation. In order to alleviate the limitations on the operating frequencies, an admittance-transforming technique is presented. By inserting inductive elements in series connection with the varactor, the equivalent shunt conductance of the tank can be effectively reduced at higher frequencies, leading to a VCO design close to the $f_{\max}$ of the MOSFETs. In addition, with the nature of the tunable effective inductance, a reasonable frequency tuning range is also demonstrated. Using a standard 0.18- μ m CMOS process, a VCO prototype is successfully implemented at the 50-GHz frequency band.

The remainder of this paper is organized as follows. In Section II, the concept of the admittance-transforming technique is illustrated. The proposed VCO topology and the circuit implementations are presented in Sections III and IV, respectively. The experimental results are provided in Section V. Finally, a conclusion is given in Section VI.

II. ADMITTANCE-TRANSFORMING TECHNIQUE

Fig. 1(a) shows the equivalent half circuit of the conventional *LC*-tank VCO where the resonator is composed of an inductor L and a varactor C_{var} . The active devices are employed to provide a negative conductance $-G_m$, compensating for the losses from the passive elements. For the derivations of the oscillation frequency and the startup conditions, a simplified equivalent circuit, as shown in Fig. 1(b), is employed. According to

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the narrowband approximation, the shunt components including L_P , C_P , and G_P can be estimated as

$$L_P = L \quad (1)$$

$$C_P = C_{\text{var}} + C_{\text{par}} \quad (2)$$

$$G_P \approx G_{\text{var}} + G_{\text{par}} + R_s/(\omega L)^2. \quad (3)$$

The oscillation frequency is given by

$$f_0 = \frac{1}{2\pi\sqrt{L_P C_P}} \quad (4)$$

and the startup condition yields

$$G_m > G_P. \quad (5)$$

Typically, G_P is dominated by the loss of the on-chip inductors for conventional LC-tank CMOS VCO designs at an oscillation frequency below 10 GHz. However, for applications at millimeter-wave frequencies, G_P is mainly determined by the losses from the varactor, imposing a fundamental limitation on the maximum operating frequency of the LC-tank VCOs [4].

To alleviate the frequency limitations due to presence of the varactors, an admittance-transforming technique is proposed. A conceptual illustration of this technique is shown in Fig. 2(a). With a series connection of the varactor C_{var} and an additional inductor L_1 , the input admittance of Y_{in} can be expressed as

$$Y_{\text{in}} = \frac{1}{j\omega L_1 + (G_{\text{var}} + j\omega C_{\text{var}})^{-1}} = G_T + (j\omega L_T)^{-1} \quad (6)$$

where

$$G_T = \frac{G_{\text{var}}}{L_1^2 C_{\text{var}}^2 \omega^4 + (L_1^2 G_{\text{var}}^2 - 2L_1 C_{\text{var}}) \omega^2 + 1} \quad (7)$$

$$L_T = \left(1 - \frac{L_1 C_{\text{var}} \omega^2 - 1}{L_1^2 C_{\text{var}}^2 \omega^4 + (L_1^2 G_{\text{var}}^2 - L_1 C_{\text{var}}) \omega^2} \right) L_1. \quad (8)$$

As indicated in (7), G_T is smaller than G_{var} provided that

$$\omega > \sqrt{\frac{2}{L_1 C_{\text{var}}} \cdot \frac{Q_{\text{var}}^2}{Q_{\text{var}}^2 + 1}} \approx \frac{\sqrt{2}}{\sqrt{L_1 C_{\text{var}}}} = \sqrt{2}\omega_R \quad (9)$$

where Q_{var} is the Q factor of the varactor and ω_R is the resonant frequency of L_1 and C_{var} . With proper choice in the value of the inductance L_1 , the high-frequency losses from the varactors can be effectively reduced by the admittance-transforming technique. Moreover, as the value of C_{var} varies, the imaginary part of Y_{in} can be treated as a tunable inductance L_T at a frequency higher than ω_R , enabling the frequency tuning mechanism for VCO operations. Note that, in order to simplify the analysis, the effective series resistance of L_1 is neglected in the above derivations.

To further investigate the influence of the loss from L_1 , a series resistance R_{s1} is included in the equivalent circuit, as shown in Fig. 2(b), and the input admittance Y'_{in} is given by

$$Y'_{\text{in}} = G'_T + (j\omega L'_T)^{-1}. \quad (10)$$

Fig. 2. Equivalent circuits of the series LC-resonator for the admittance-transforming technique with: (a) lossless and (b) lossy inductor L_1 .

Provided $R_{s1} G_T \ll 1$ and $(\omega L_T/R_{s1})^2 \gg 1$ in typical design cases, G'_T and L'_T in (10) are approximated by

$$G'_T \approx G_T + \frac{1}{Q_{L1}\omega L_1} \cdot \left(\frac{L_1}{L_T} \right)^2 \quad (11)$$

$$L'_T \approx L_T \quad (12)$$

where Q_{L1} is the Q factor of the inductor L_1 . From (11), it is apparent that the value of G'_T depends on R_{s1} . Fortunately, the second term in (11) is typically smaller than G_{var} by one order of magnitude at millimeter-wave frequencies. As a result, the proposed admittance-transforming technique is still effective with the inductor losses taken into account.

In order to have a better understanding of the admittance-transforming technique, simulations on the series resonator is performed by taking the parasitics and losses into consideration. For a 0.2-pF varactor with various values of inductance L_1 , the resulting G_T is shown in Fig. 3(a). As predicted in (7), the shunt conductance can be effectively reduced at higher frequencies. On the other hand, with a fixed L_1 of 0.25 nH, the equivalent inductance L_T is demonstrated in Fig 3(b) as the value of C_{var} varies, featuring a frequency tuning mechanism for the VCO circuit.

III. PROPOSED TOPOLOGY OF VCO

Fig. 4 shows the schematic of the proposed VCO, where the cross-coupled pair M_1 and M_2 is used to provide the required negative conductance and the resonator is established by the inductor L_2 and the varactor C_{var} . In consideration of the flicker noise contributed by the active devices, a pMOS transistor is employed as the tail current such that lower phase noise can be achieved. Meanwhile, the proposed admittance-transforming technique is introduced by inserting the inductors L_1 in the series path with the varactors. As a result, the VCO can operate at

Fig. 3. (a) Simulated equivalent conductance G_T for various values of L_1 with $C_{var} = 0.2$ pF. (b) Simulated equivalent inductance L_T for various values of C_{var} with $L_1 = 0.25$ nH.

Fig. 4. Circuit topology of the proposed millimeter-wave VCO.

much higher frequencies than the conventional ones while maintaining desirable circuit performance. To gain more insight into the proposed VCO topology, detailed circuit analysis is given as follows.

A. Startup Condition and Oscillation Frequency

For the derivations of the oscillation frequency and the startup conditions, the equivalent half circuit of the proposed VCO is shown in Fig. 5, where G_2 and G_{par} represent the losses from the on-chip inductor L_2 and the active devices, respectively. Note that the loss of the inductor is typically modeled by a series resistance. In the equivalent-circuit presentation, the nar-

Fig. 5. Equivalent half circuit of proposed VCO.

rowband approximation is employed, and the shunt conductance G_2 is given by

$$G_2 \approx R_{s2}/\omega^2 L_2^2 \quad (13)$$

where R_{s2} is the equivalent series resistance of L_2 . As for the inductance L'_T and shunt conductance G'_T , the analytic expressions in terms of L_1 and C_{var} are given in (11) and (12), respectively. To initiate the oscillation, the negative conductance should be sufficiently large to cancel out the total parallel conductance

$$G_m > G'_T + G_{par} + G_2. \quad (14)$$

On the other hand, the oscillation frequency is given by

$$f_0 = \frac{1}{2\pi\sqrt{(L_2//L'_T) \cdot C_{par}}}. \quad (15)$$

From (14) and (15), it is apparent that the use of the admittance-transforming technique leads to an alleviated startup condition and a higher oscillation frequency compared with the conventional circuit topology. In addition, by varying the capacitance of the varactor from $C_{var,min}$ to $C_{var,max}$ through the controlled voltage, the tuning range of the VCO is defined by

$$\Delta f_0 = \frac{1}{2\pi\sqrt{C_{par}(L_2//L'_{T,min})}} \cdot \left(1 - \sqrt{\frac{L'_{T,min}}{L'_{T,max}} \cdot \frac{L_2 + L'_{T,max}}{L_2 + L'_{T,min}}}\right). \quad (16)$$

Note that $L'_{T,max}$ and $L'_{T,min}$ are the maximum and minimum values of the tunable inductance L'_T , respectively, which can be calculated by (17) and (18), shown at the bottom of the following page. From (16)–(18), it is clear that the tuning range in this study is related to the values of L_1 , L_2 , and C_{var} . In order to evaluate the tuning characteristics, the simulated tuning range of the proposed VCO as a function of the inductance L_1 is shown in Fig. 6(a), where the value of L_2 is determined by L_1 such that the central oscillation frequency of the VCO remains unchanged. Based on the simulation results, the tuning range degrades as the value of L_1 increases. Meanwhile, the transconductance required for the VCO startup is effectively reduced. A design tradeoff is thus provided between the startup condition and the tuning range for the circuit implementations.

One interesting phenomenon appearing in Fig. 6(a) is that the tuning range increases as the size of the varactor decreases.

Fig. 6. Simulated: (a) tuning range and (b) phase noise of the 50-GHz VCO as a function of L_1 for various C_{var} values.

This observation conflicts with the design concept in conventional VCO circuits. By employing the admittance-transforming technique, the capacitance of the varactor is effectively transferred into a tunable inductance at the frequency of interest. Consequently, an enhanced tuning range can be achieved by decreasing the size of the varactor. However, for a given operating frequency, (9) indicates that the inductance L_1 is inversely proportional to the value of C_{var} , setting a lower bound for the varactor size in a practical design.

Another important feature of the admittance-transforming technique is the relation between the frequency tuning range and the phase noise. For conventional VCO circuits, the tuning range increases with the varactor size. As the Q factor of the varactors degrades significantly with the device size, especially at millimeter-wave frequencies, the wide tuning-range VCOs usually suffer from an inferior phase noise. To evaluate the noise performance of the proposed VCO, circuit simulations

are performed with identical design parameters in the investigation of the tuning range. The simulated phase noise is shown in Fig. 6(b). By employing the admittance-transforming technique, the shunt conductance of the varactor is suppressed, while the frequency tuning is achieved by the tunable effective inductance. As indicated in the simulation results, the VCO exhibits a phase noise virtually independent of the tuning range, which is a unique and desirable property for the circuit optimization.

B. Phase Noise

According to the Leeson's model [10], the phase noise of the proposed VCO in the $1/f^2$ region can be expressed as

$$L\{\Delta f\} = 10 \log \left\{ \frac{2FkT}{P_{sig}} \left[1 + \left(\frac{f_0}{2Q\Delta f} \right)^2 \right] \right\} \quad (19)$$

where F is the excess noise factor, Q is the Q factor of the LC-tank, and Δf is the frequency offset. As indicated in (19), the phase noise is strongly influenced by the Q factor of the tank, which can be expressed as

$$Q = \frac{1}{(G'_T + G_{par} + G_2)} \sqrt{\frac{C_{par}}{L_2/L'_T}}. \quad (20)$$

As indicated in (20), reduced phase noise can be achieved in the VCO design by suppressing the shunt conductance ($G'_T + G_{par} + G_2$) and by increasing the value of $C_{par}/(L_2/L'_T)$. However, an increasing ratio of $C_{par}/(L_2/L'_T)$ requires large cross-coupled transistors, which simultaneously increases the shunt conductance G_{par} and possibly the phase noise. Therefore, the values of L_1 , L_2 , and C_{par} should be chosen carefully in the proposed VCO for optimum phase-noise performance.

Fig. 7 shows the simulated phase noise of the VCO at 1-MHz offset with a fixed oscillation frequency of 50 GHz and a varactor value of 0.2 pF. As the values of L_1 and L_2 increase, the phase noise tends to decrease due to the smaller C_{par} required. However, further increase in L_1 and L_2 leads to higher phase noise due to the degraded Q factors of the inductors. In order to achieve the lowest phase noise possible, the Q factors of the varactors and inductors have to be taken into consideration. In this particular example, the optimum phase noise is approximately -100 dBc/Hz provided that L_1 and L_2 are 0.25 and 0.2 nH, respectively, with a Q factor of 10.

C. Output Buffers

In order to drive the 50- Ω impedance of the test instrument, an output buffer is required for the circuit design. For a VCO

$$L'_{T,max} = \left(1 - \frac{L_1 C_{var,max} \omega^2 - 1}{L_1^2 C_{var,max}^2 \omega^4 + (L_1^2 G_{var,max}^2 - L_1 C_{var,max}) \omega^2} \right) L_1 \quad (17)$$

$$L'_{T,min} = \left(1 - \frac{L_1 C_{var,min} \omega^2 - 1}{L_1^2 C_{var,min}^2 \omega^4 + (L_1^2 G_{var,min}^2 - L_1 C_{var,min}) \omega^2} \right) L_1 \quad (18)$$

Fig. 7. Simulated phase noise of the proposed VCO with an oscillation frequency of 50 GHz and a varactor of 0.2 pF.

Fig. 8. Circuit schematic of the output buffer.

Fig. 9. Structure of the CPW.

Fig. 10. Extracted equivalent inductance and Q factor of the CPW with various design parameters.

operating at a frequency close to the $f_{\max}$ of the transistors, the output buffer is also a great design challenge. In this study, a

Fig. 11. (a) Conventional and (b) inversion-mode pMOS varactors.

Fig. 12. Simulated capacitance and Q factor of the conventional and the inversion-mode varactor.

TABLE I
CIRCUIT PARAMETERS OF THE U -BAND VCO

Devices		Design Values	
M_1, M_2		$2.5 \times 14 \mu\text{m} / 0.18 \mu\text{m}$	
Buffers	M_{B1}	$2.5 \times 3 \mu\text{m} / 0.18 \mu\text{m}$	
	M_{B2}	$2.5 \times 4 \mu\text{m} / 0.18 \mu\text{m}$	
L_1	Inductance	0.24 nH	
	Q Factor	11.0	
L_2	Inductance	0.09 nH	
	Q Factor	14.0	
C_{var}	Capacitance	0.17~0.32 pF	
	Q Factor	4.8~7.8	
L_{B1}	Inductance	0.26 nH	
	Q Factor	8.2	
L_{B2}	Inductance	0.37 nH	
	Q Factor	6.5	
C_{B1}	Capacitance	0.8 pF	
	Q Factor	5.6	

two-stage buffer with common-source MOSFETs is employed [4], [11]. Fig. 8 shows the circuit schematic where L_{B1} , L_{B2} , and C_{B1} are on-chip passive components. The transistor sizes and inductance values are chosen such that the loading effect on

Fig. 13. Microphotograph of the fabricated circuit.

Fig. 14. (a) Oscillation frequency and (b) output power level of the VCO as the controlled voltage sweeps from 0 to 1.8 V.

the VCO core can be minimized while maintaining sufficient output swing.

IV. CIRCUIT IMPLEMENTATION

In the VCO design, it is desirable to reduce the shunt conductance for more relaxed startup conditions and enhanced phase noise. Due to the use of the admittance-transforming technique, the contribution from the varactor elements is thus minimized,

Fig. 15. Measured: (a) wideband spectrum, (b) narrowband spectrum, and (c) close-in phase noise of the fabricated VCO at 49 GHz.

especially at higher frequencies. Further performance improvement can be achieved by optimizing the Q factors of the on-chip passive components.

In consideration of the Q factor at millimeter-wave frequencies, the required inductors are implemented by the coplanar waveguide (CPW), which is constructed by one signal line surrounded by two adjacent grounded plates, as shown in Fig. 9. By properly choosing the dimensions of the linewidth w , the signal-to-ground spacing s , and the length l , the required inductance can be realized with sufficient accuracy. Based on the full-wave electromagnetic (EM) simulation, the extracted equivalent inductance and the Q factor of the CPW at 50 GHz are shown in

TABLE II
PERFORMANCE SUMMARY OF THE CMOS VCOs AT THE 50-GHz FREQUENCY BAND

	Unit	This Work	[2]	[5]	[14]	[9]
Technology	—	0.18- μm CMOS	0.18- μm CMOS	0.18- μm CMOS	0.18- μm CMOS	0.12- μm CMOS
Frequency	GHz	49	53	49	52	51
Tuning Range	GHz	0.8 1.1 ³	0.1	1.0 ²	0.6	1.0 ^{2,3}
Phase Noise @1-MHz	dBc/Hz	-101	-86	-96	-97	-85
V_{DD}	V	1.8	—	1.0	2.1	1.0
DC power	mW	45	41	4	27	1
Output Power Level	dBm	-11 ~ -12	-8	-26 ~ -27	-16	-30
Phase	—	Differential	Single	Differential	Single	Differential
FOM ¹	dBc/Hz	178	164	184	177	179

¹ FOM = $10 \log_{10} \left[\left(\frac{f_0}{\Delta f} \right)^2 \cdot \frac{1}{L(\Delta f) \cdot P_{dc}} \right]$. ² The bias conditions are changed to increase tuning range.

³ The controlled voltage is higher than supply voltage.

Fig. 10 for various w and s values. Note that, with the two separate ground plates in the CPW structure, underpasses using the second metal layer are also employed to suppress the unwanted odd-mode operations [12].

Another important design issue of the millimeter-wave VCO is the realization of the varactors. Fig. 11(a) shows the terminal connection of a conventional pMOS varactor where the body terminal is shorted to the source and the drain of the device. In this VCO design, varactors with the inversion-mode topology, as shown in Fig. 11(b), are employed to enhance the Q factor at the higher frequency bands, while maintaining a relatively wide tuning range [13]. By connecting the body terminal to V_{DD} , the pMOS varactor can be operated in the accumulation region in addition to the strong, moderate, and weak inversion ones. As a result, the capacitance ratio of $C_{var,max}/C_{var,min}$ is effectively increased. Fig. 12 shows the simulated capacitance and Q factor of the pMOS varactors at 50 GHz for comparison.

Using a standard 1P6 M 0.18- μm CMOS process, a U -band VCO is implemented for demonstration. With a device layout optimized for the RF performance, the n -channel MOSFET in a deep n -well exhibits a maximum oscillation frequency f_{max} up to 60 GHz. As for the on-chip passive components, a top AlCu metallization layer with a thickness of 2 μm and the metal-insulator-metal (MIM) structure with an oxide intermetal dielectric are provided in this technology. Table I summarizes the design parameters of the proposed VCO.

In order to verify the derivations from (10) to (12) in this circuit implementation, the values of R_{s1} , G_{var} , G_T , L_T , G'_T , and L'_T are calculated from Table I. The resulting values of $R_{s1}G_T$ and $(\omega L_T/R_{s1})^2$ at the 50-GHz frequency band are 3.14×10^{-3} to 4.95×10^{-3} and 68.0 to 92.7, respectively, validating the assumptions of $R_{s1}G_T \ll 1$ and $(\omega L_T/R_{s1})^2 \gg 1$ in the theoretical derivation. Meanwhile, by using the admittance-transforming technique, the shunt conductance G_{var} with an original value from 6.8 to 20.9 mS is effectively suppressed to a value between 2.0–2.9 mS, leading to a relaxed startup condition for the 50-GHz VCO design.

V. EXPERIMENTAL RESULTS

Fig. 13 shows the die photograph of the fabricated circuit with a chip area of $0.83 \times 0.87 \text{ mm}^2$ including the RF pads. To evaluate the high-frequency performance, the output spectrum and

phase noise were characterized by a 50-GHz spectrum analyzer. The fabricated VCO was tested by on-wafer probing, while the losses from the measurement setup were calibrated and deembedded in the experimental results.

With a supply voltage of 1.8 V, the VCO starts to oscillate at a dc current level of 1 mA due to the use of the admittance-transforming technique. In order to obtain sufficiently high output power, the performance of the VCO is characterized under a core bias current of 25 mA. As the controlled voltage sweeps from 0 to 1.8 V, the tuning characteristics, including the oscillation frequency and output power, are shown in Fig. 14. Within the output frequency range from 48.4 to 49.2 GHz, the U -band VCO exhibits an output power in the vicinity of -11 dBm . By applying a controlled voltage up to 2.2 V, an enhanced tuning range of 1.1 GHz can be achieved. Fig. 15 shows the measured wideband and narrowband spectra and the close-in phase noise at 49 GHz, indicating a phase noise of -100.6 dBc/Hz at 1-MHz offset. Compared with the simulated result of -100.9 dBc/Hz at 1-MHz offset, good agreement is achieved due to accurate EM extraction. The performance of the VCO is summarized in Table II along with the results from previously published CMOS VCOs for comparison.

VI. CONCLUSION

In this paper, an admittance-transforming technique has been presented for the design of the LC -tank CMOS VCOs at higher frequencies. By effectively reducing the shunt conductance contributed by the varactors, VCO circuits with an oscillation frequency close to the f_{max} of the active devices can be realized while maintaining the desirable circuit performance. Using a standard 0.18- μm CMOS technology, a U -band VCO based on the proposed technique is successfully demonstrated. It is well suited for low-cost implementations of the RF systems at millimeter-wave frequencies.

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