V-Band Varactor-less Interpolative-Phase-Tuning Oscillators with Multiphase Outputs

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Abstract - An interpolative-phase-tuning technique is presented to tune the frequency of millimeter-wave LC-based ring oscillators without using varactors. Two oscillator prototypes are designed and implemented in a 0.13- μm CMOS process. The first oscillator operates at 50GHz with 8 output phases and measures phase noise of -127.8dBc/Hz at 10MHz offset, tuning range of 6.8%, and figure of merit (FOM) of 186.4dB while occupying chip area of 0.36mm². The second prototype oscillates at 60GHz with 4 output phases and measures phase noise of -120.6dBc/Hz at 10MHz offset, tuning range of 9%, and FOM of 180.6dB with chip area of 0.2mm².

I. INTRODUCTION

Aggressive scaling of CMOS technologies makes it possible to design and integrate voltage-controlled-oscillators (VCOs) at millimeter-wave (MMW) frequencies [1]. These MMW VCOs are preferred to generate multiphase outputs in order to support in-phase and quadrature-phase (IQ) modulation and demodulation in modern wireless transceivers. In addition, multiphase LO signals are also required for the phase-array applications [2] as well as the half-rate clock-and-data recovery (CDR) in wire-line systems [3].

On the other hand, compared to the RF VCOs, the spectrum purity of MMW VCOs are much degraded due to the inferior Q of the varactors at high frequencies and the serious AM-PM noise transformation caused by the large VCO gain on the order of GHz/V [4]. Moreover, the serious trade-off between varactor's tuning ratio and Q [5] and the reduced supply voltage in deep sub-micron CMOS technologies make the varactor-tuning method less effective for MMW VCOs.

An interpolative-phase-tuning (IPT) technique is proposed in this work to tune the frequency of multiphase MMW LC-based ring oscillators without using varactors. Section II describes the topology and the operation principle of the proposed IPT in a 4-stage current-controlled-oscillator (CCO) with 8 output phases. In Section III, the proposed phase tuning technique is demonstrated in a commonly-used 2-stage oscillator with IQ outputs. Section IV presents the experimental results, and conclusions are drawn in Section V.

II. PROPOSED IPT 8-PHASE CCO

Fig. 1(a) shows the schematic of a conventional 4-stage LC-based ring oscillator. To ensure stable oscillation, the voltage gain of each stage needs to be larger than 1, and the phase shift provided by each stage, which is induced by the LC tank, has to be $\pi/4 + (\pi/2) \cdot N$. If a second-order LC tank is used in each

Fig. 1 Conventional 4-stage LC ring oscillator and the frequency responses of the LC tank with: (a) ideal parallel LC tank, (b) Q_L dominating tank's Q, and (c) significant phase delay due to the transconductor.

stage, N can be either 0 or -1. For an ideal parallel LC tank with symmetrical magnitude and phase responses at frequencies ω_{high} and ω_{low} , as shown in Fig. 1(a), where $\omega_{\text{high}} > \omega_0 > \omega_{\text{low}}$ with ω_0 being the peak frequency of the LC tank, the tank can provide the required phase shift $-\pi/4$ and $\pi/4$ with the same magnitude. Thus, the oscillator can possibly operate at both frequencies. In practice, the oscillator would preferably operate at one frequency rather than the other, because of two effects. At radio frequencies below 10GHz, the integrated inductor typically limits the tank's Q, and the phase shift of the LC tank at ω_0 is actually not zero as shown in Fig. 1(b). Consequently, the oscillator tends to operate at ω_{high} with larger tank impedance [6]. At higher frequencies, in the MMW range, the tank capacitor's Q becomes much smaller, and it is possible that $Q_L > Q_C$. Nevertheless, the non-ideal transconductor devices could introduce significant negative phase shift $\Delta\theta$ when the operation frequency is close to the cut-off frequency ω_T of the transistors. As a result, as shown in Fig. 1(c), the required

Fig. 2 Proposed 4-stage LC ring oscillators with: (a) tunable phase shift, (b) interpolative phase tuning at each stage.

phase shift of the LC tank becomes $\pm\pi/4-\Delta\theta$. It is still typically true that $|Z(\omega_{\text{high}})| > |Z(\omega_{\text{low}})|$ and the circuit would still prefer to oscillate at ω_{high} .

On the other hand, this interesting phase relationship indicates that extra phase shift $\Delta\theta$ can be placed in the Gm cell to change the required phase shift θ provided by the LC-tank and thus the oscillation frequency. As shown in Fig. 2(a), the total phase shift ϕ including the intrinsic device phase shift $\Delta\theta$ is added in front of the transconductor. Based on the phase condition, each stage needs to provide a total phase shift of $\phi + \theta = \pi/4 + (\pi/2) \cdot N$. As plotted in the frequency response of the LC tank with assumption of symmetry for simplicity, if ϕ is a negative value close to zero, θ needs to be close to $-\pi/4$, and the oscillator would operate around the highest frequency ω_{high} . If $\phi = -\pi/4$, the LC tank doesn't need to provide any phase shift, and the oscillator would oscillate at the peak frequency of the LC tank. When ϕ is reduced and approaches $-\pi/2$, θ tends to increase to $\pi/4$, and the oscillation occurs at the minimum frequency limit ω_{low} . If ϕ is further reduced, the oscillator will oscillate again around ω_{high} where the tank impedance is larger.

The actual implementation of the proposed interpolative-phase-tuning (IPT) scheme to tune the phase shift ϕ and thus the oscillation frequency is shown in Fig. 2(b). A fixed phase shift β is used to introduce a delayed current i_1 via M_3 and M_4 , which is interpolated with the un-delayed current i_0 provided by M_1 and M_2 . By controlling the biasing DC current I_0 and I_1 , the delay of the total current i_t can be tuned from 0 to β . Taking into account the intrinsic phase shift of the transistors, β can be designed as $-\pi/2 - \Delta\theta_{M3/4}$ to obtain a maximum monotonous frequency tuning range. At high frequencies, the phase shift β can be simply implemented by another LC-based

Fig. 3 2-stage LC ring oscillators: (a) conventional Q-CCO, (b) IPT Q-CCO.

differential stage, where the inductance L_1 is designed to be much larger than L_0 to make β close to $-\pi/2$. The resistor $R_{\text{de-Q}}$ is added in parallel with the tank to flatten the magnitude response of the differential stage and reduce the variation of β within the tuning range.

III. PROPOSED IPT 4-PHASE CCO

The IPT technique can be also applied to tune the oscillation frequency of quadrature CCOs (Q-CCOs). Fig. 3(a) shows the schematic of the conventional 2-stage ring Q-CCO. As there are only two stages in the ring, each stage needs to provide a phase shift of $+\pi/2$ or $-\pi/2$. As shown in the phasor diagram of the current flowing into the LC tank, the current i_0 provided by M_1 and M_2 is combined with the current i_1 provided by the cross-coupled pair M_3 and M_4 to generate a total current i_t with a phase shift ϕ . Since the absolute value of the phase shift ϕ is smaller than $\pi/2$, the LC tank needs to provide an additional phase shift θ to satisfy the phase condition. As a result, the conventional Q-CCO cannot operate at the peak frequency of the LC tank. Because of the asymmetrical frequency response of the LC tank and the phase shift due to the transistors, in general the Q-CCO would prefer to oscillate at the frequencies higher than the peak frequency of the LC tank. Consequently, if i_0 is increased by increasing the biasing current I_0 , the phase shift $|\phi|$ would be reduced, and the LC tank needs to provide more negative phase shift θ , which would increase the oscillation frequency of the Q-CCO. The lower boundary of the frequency tuning is limited by IQ phase error while the upper boundary is limited by the phase noise performance.

Based on the same phase-tuning concept, a novel Q-CCO can be built as shown in Fig. 3(b). Two fixed phase shifts β_1 and β_2 are introduced into each Gm cell. From the current phasor diagram, it can be seen that the phase shift ϕ provided by each Gm cell can be controlled from β_1 to $\beta_1 + \beta_2$. To satisfy

the phase condition, the phase shift provided by the LC tank needs to vary from $-\pi/2-\beta_1$ to $-\pi/2-(\beta_1+\beta_2)$ assuming β_1 and β_2 are negative, and thus the oscillation frequency would decrease accordingly. Assuming that the LC tank has symmetric frequency response for simplicity, β_1 should be designed as $(-\pi/2-\beta_2/2)$ to achieve a frequency tuning range symmetrically and optimally around the peak frequency of the tank with the maximum tank Q. On the other hand, β_2 determines the actual tuning range and can be ideally designed to be close to $(-\pi/2-\Delta\theta_{M3/4})$ with $\beta_1=0$. Ideally, an infinitely large tuning range could be obtained if the gain condition is not a problem. In practice, $|\beta_2|$ is limited by the power budget as well as the phase noise performance. As $|\beta_2|$ becomes larger and larger, the magnitude of i_t would become smaller and smaller due to the current interpolation, and it would become difficult to meet the gain condition. Similar to the 4-stage IPT CCO, the phase shift β_1 and β_2 can be simply implemented as differential pairs with low-Q LC tanks. In our design, β_1 and β_2 are designed to be around $-\pi/4$ and $-\pi/2$, respectively.

Compared to the conventional Q-CCO, the IPT Q-CCO can oscillate at a much larger frequency range around the peak frequency of the LC tank. Moreover, in the conventional Q-CCOs, only part of the biasing current is used for the coupling pairs. Consequently, at the edge of the tuning range when less current is available to the IQ coupling transistors, the oscillator becomes more sensitive to the mismatches. In contrast, in the proposed IPT Q-CCOs, because all the biasing currents are used for the coupling pairs, over the whole frequency tuning range no matter how the current is distributed, the coupling strength around the loop does not change much, and the loop can self-calibrate for the mismatches among the stages. As a result, the IPT CCO is much less sensitive to mismatches.

Assuming that the LC tank is symmetrical and that β of the 8-phase CCO and β_2 of the 4-phase CCO are around $-\pi/2$, the oscillation frequency of the two IPT CCOs can be estimated based on the phase condition as:

$$\omega_{\text{osc}} = \frac{\omega_0}{1 - \frac{1}{2Q_{\text{peak}}} \frac{1-a}{1+a}}$$

where $a=|i_t|/|i_0|$ is defined as the ac current ratio, ω_0 is the peak frequency of the LC tank, and Q_{peak} is the tank Q at the peak frequency. Numerically, Table I lists the achievable tuning range of the IPT VCOs with different tank Qs. When Q_{peak} is reduced from 20 to 3, the tuning range can be enlarged from 5% to 33.3%. For the proposed IPT CCOs, there is also a

TABLE-I ACHIEVABLE TUNING RANGE (TR) AT DIFFERENT Q_{PEAK}

Q_{peak}	$\omega_{\text{min}} (a=\infty)$	$\omega_{\text{max}} (a=0)$	TR
20	$0.976\omega_0$	$1.026\omega_0$	5.0%
15	$0.968\omega_0$	$1.034\omega_0$	6.6%
10	$0.952\omega_0$	$1.053\omega_0$	10.1%
7	$0.933\omega_0$	$1.077\omega_0$	14.3%
5	$0.909\omega_0$	$1.111\omega_0$	20.0%
3	$0.857\omega_0$	$1.200\omega_0$	33.3%

Fig. 4 Die photos of: (a) 8-phase CCO, and (b) 4-phase CCO.

Fig. 5 Measured tuning curves of: (a) 8-phase CCO, (b) 4-phase CCO.

trade-off between Q and tuning range. However, different from the capacitive tuning VCOs, this trade-off is independent of the operation frequency. As such, when the desired oscillation frequency becomes so high that capacitive-tuning method is no longer effective, the proposed phase-tuning technique would provide a good solution.

IV. EXPERIMENTAL RESULTS

The two IPT CCOs are designed and fabricated in a 0.13- μm CMOS process. Fig. 4(a) and 4(b) show the die photos of the 8-phase CCO and 4-phase CCO, which occupy chip area of 0.36mm^2 and 0.20mm^2 , respectively. Transmission lines are used to implement all the inductors to reduce the undesired mutual magnetic coupling. Fig. 5(a) and 5(b) show the measured frequency tuning curves of the two oscillators. When the differentially-controlled biasing current is tuning from -0.9mA to 0.9mA , the 8-phase CCO can be tuned continuously from 48.6GHz to 52GHz while consuming 32mW to 48mW from a 0.8V supply. The 4-phase CCO can be tuned continuously from 56GHz to 61.3GHz , with a total power of 30mW to 37mW from a 0.8V supply.

To measure the phase noise, the output signals of the oscillators are down-converted by an external V-Band mixer with LO signal around 50GHz generated by a signal generator. Assuming the phase noise contributed by the mixer and the signal generator is negligibly small, Fig. 6(a) and 6(b) plots the phase noise curves of the two IPT oscillators. The cyan, yellow, and purple curves are measured when the CCOs oscillate at the lowest, middle, and highest frequencies, respectively. The measured phase noise at 3MHz offset vary from -110.2 to -116.3dBc/Hz for the 8-phase CCO and from -104.8 to -110.1dBc/Hz for the 4-phase CCO. For both of the oscillators,

Fig. 6 Measured phase noise: (a) 8-phase CCO, (b) 4-phase CCO

as expected, the lowest phase noise is measured when they oscillate at the middle frequency, at which the tank Q is maximized.

Table II summarizes the performance of the proposed IPT oscillators and compares with that of the recently reported MMW oscillators. For n-stage LC oscillators, the phase noise contributed by each stage can be reduced by a factor n^2 while the number of stages is increased by n. As such, theoretically the phase noise can be improved by n. On the other hand, as the power consumption is also increased by n, the FOM as defined below becomes independent of the number of output phases [7]. From Table II, as compared to other MMW oscillators with varactor tuning, the proposed IPT oscillators

achieve much lower phase noise and better FOM and FOM_T while providing multiple output phases over comparable tuning range even in a less advanced process.

$$FOM = 10 \log \left[\left(\frac{f_0}{\Delta f} \right)^2 \frac{1}{L(\Delta f) \times P_{diss} |_{mW}} \right]$$

$$FOM_T = FOM + 20 \log[TR / 10]$$

V. CONCLUSIONS

A novel interpolative-phase-tuning technique is proposed in this work to implement varactor-less multi-phase LC oscillators with wide tuning range and low phase noise at MMW frequencies. Two phase-tuning CCO prototypes, one with 8-phase 50GHz outputs and another with 4-phase 60GHz outputs, implemented in a 0.13- μ m CMOS process and operated at 0.8-V supply, measure phase noise of -127.8dBc/Hz and -120.6dBc/Hz at 10MHz offset, FOMs of 186.4dB and 180.6dB, and FOM_T s of 183dB and 179.7dB, respectively, which are much better compared to other state-of-the-art MMW oscillators using capacitive tuning.

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TABLE-II SUMMARY AND COMPARISON OF MMW OSCILLATORS

Ref.	Phase No.	Freq. [GHz]	TR [%]	PN [dBc/Hz]	Power [mW]	FOM	FOM_T	Process
[1]	2	70.2	9.6	-106.1 @10MHz	5.4	175.8	175.4	65nm SOI
[4]	2	58.4	9.3	-91 @1MHz	8.1	177.2	176.6	90nm CMOS
[5]	2	59	9.8	-89 @1MHz	9.8	174.5	174.4	0.13 μ m CMOS
[5]	2	98.5	2.5	-102.7 @10MHz	7	174.1	162.2	0.13 μ m CMOS
[5]	2	105.2	0.2	-97.5 @10MHz	7.2	169.4	135.0	0.13 μ m CMOS
This work	8	50.3	6.8	-127.8 @10MHz	35	186.4	183.0	0.13 μ m CMOS
	4	58.5	9	-120.6 @10MHz	34	180.6	179.7	0.13 μ m CMOS