A 24-GHz and 60-GHz Dual-Band Standing-Wave VCO in 0.13 μ m CMOS Process

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Abstract — By exploiting the intrinsic multiple oscillation modes of a standing-wave oscillator, a dual-band millimeter-wave VCO is designed. Implemented in 0.13 μ m CMOS with an area of 0.05mm², the VCO prototype measures a dual-band operation at 24 GHz and 60 GHz with tuning range of 10.8% and 7.2%, phase noise of -120dBc/Hz and -114dBc/Hz at 10MHz offset, power consumption of 11mW and 24mW, corresponding to FoM of -177dB and -176dB, respectively.

Index Terms — Voltage controlled oscillator, millimeter-wave, standing-wave, dual-band, CMOS, 24 GHz, 60 GHz.

I. INTRODUCTION

With recent advances in CMOS technology, millimeter-wave (MMW) circuits become more and more attractive for many interesting applications such as vehicle radars at 77 GHz and 24 GHz and communication systems at 60 GHz and 24 GHz. Moreover, to support multiple-band systems and to increase the level of integration without increasing the chip area and thus the fabrication costs, multiple-band MMW VCOs are highly desired. However, existing multiple-band VCO techniques, including tapped-inductor based [1] and transformer-based [2], are not suitable for MMW frequency bands. Prediction of their performance is difficult due to the inaccurate modeling of both passive and active devices at the MMW frequencies. In addition, the performance of these existing VCOs is typically not good because of the low quality factor of inductors and transformers. Wave-based oscillators such as standing-wave oscillators (SWO) can reach operation frequencies approaching the devices' transition frequency f_T with proper distribution of active gain elements along a transmission line. Given the quasi-transverse electromagnetic mode of propagation, transmission lines are capable of realizing precise values of small reactance and are inherently scalable in length. Also, the well-defined ground return path significantly reduces magnetic and electric field coupling to adjacent structures [3].

In this paper, a dual-band MMW SWO at 24 GHz and 60 GHz is demonstrated by exploiting and switching the intrinsic multiple standing-wave modes. Employing only one differential transmission line, this dual-band VCO occupies an area of only 0.05mm² and achieves figures of

merit (FOMs) in the two frequency bands comparable with that of existing single-band MMW VCOs.

II. MULTIPLE-MODE OSCILLATION OF SWO

In a $\lambda/4$ SWO, the boundary conditions allow several standing-wave modes at $l_0 = \lambda/4 \times n$ ($n=1, 3, 5 \dots$) where l_0 is the length of transmission line. The corresponding frequency can be expressed as:

$$f_n \approx \frac{n}{4l_0\sqrt{LC}} \quad (1)$$

where L and C are the inductance and capacitance per unit length.

In the fundamental mode ($n=1$), the voltage amplitude exhibits monotonic variations as a function of the position with an oscillation frequency f_L as depicted in Fig. 1(a). On the other hand, in the third-order mode ($n=3$), the voltage amplitude exhibits periodic variations along the transmission line with an oscillation frequency f_H as shown in Fig. 1(b). Oscillations in the fifth mode and higher modes are insignificant and negligible due to the substantial high-frequency loss.

Fig. 1 Voltage amplitudes of a $\lambda/4$ SWO operating in (a) the fundamental mode and (b) the third-order mode

The multiple modes of SWO imply multiple impedance peaks of the transmission line. Using the S-parameters of the transmission line obtained from EM simulations in

ADS Momentum and the capacitor models from the foundry, the small-signal model is built and simulated as shown in Fig. 2, where C_p is the parasitic capacitance. The impedance seen from Node T_3 exhibits two distinctive peaks with $|Z_{3L}| > |Z_{3H}|$, while the impedance seen from Node T_1 also exhibits two distinctive peaks: $|Z_{1H}| > |Z_{1L}|$, where $f_H \approx 3f_L$. As a result, mode-switching for dual-band operation can be achieved by injecting energy at different nodes, i.e. low-band and high-band oscillations are excited by injecting energy at Nodes T_3 and T_1 , respectively.

Fig. 2 Impedance seen at Nodes (a) T_3 and (b) T_1

III. ANALYSIS ON THE STABILITY ISSUE

In order to achieve stable oscillation of the VCO, it is critical to ensure that the impedance at the desired frequency is significantly larger than that at the other frequencies. If energy is injected at Node T_3 , the low-band oscillation is excited and can work stably since $|Z_{3L}|$ is much larger than $|Z_{3H}|$ as shown in Fig. 2(a). However, if energy is injected at Node T_1 , the high-band oscillation is potentially unstable and may jump to the low-band in the presence of process variations since $|Z_{1H}|$ is just slightly larger than $|Z_{1L}|$ as shown in Fig. 2(b). Therefore, a stabilization technique is indispensable to achieve stable and proper oscillation.

Analysis on SWO is performed with transmission line theory in the situation when the high-band oscillation is desired by injecting energy at Node T_1 . To simplify the calculation, it is assumed that parasitic capacitance is absorbed into the transmission line. By treating the SWO as a distributed oscillator, the start-up requirement can be derived. Fig. 3 shows a simplified model of the SWO [4] with a negative-conductance stage injecting energy to Node T_1 .

Starting from Node T_1 , the forward wave V_1 is amplified by the gain stage and travels toward the right end where it is completely reflected. The first reflected

wave travels along the transmission line to the left until it arrives at Node T_3 where it is partially reflected if Z_s is neither infinitely large nor 0. The second reflected wave propagates along the transmission line to the right and finally reaches T_1 where it is amplified to V_1' . The amplitude of V_1' can be expressed as:

$$V_1' = -\left(\frac{1}{2}g_m Z_0\right)^2 e^{-(\alpha+j\beta)2l_0} \Gamma_3 V_1. \quad (2)$$

where α is the attenuation constant and β is the phase constant of the transmission line. The reflection coefficient at Node T_3 can be express as:

$$\Gamma_3 = |\Gamma_3| e^{j\theta} = \frac{Z_s - Z_0}{Z_s + Z_0} = 1 - \frac{2Z_0}{Z_s + Z_0}. \quad (3)$$

To satisfy the phase condition for oscillation:

$$e^{-j\beta \cdot 2l_0} e^{j\theta} = -1. \quad (4)$$

Fig. 3 Simplified model of SWO operating in the high-band mode

The required normalized gain to sustain oscillation can be derived by setting $|V_1| = |V_1'|$, that is:

$$\left(\frac{1}{2}g_m Z_0\right)^2 = \frac{1}{e^{-\alpha 2l_0} |\Gamma_3|}. \quad (5)$$

As the attenuation α increases with frequency, the high-band oscillation would need more energy than the low-band oscillation. However, the gain condition is relaxed and the high-band oscillation is easily achieved if $|\Gamma_3|$ is increased with frequency, which can be implemented by adding capacitor C_s at Node T_3 in such a way that $C_s > l_0 C$, where C is the capacitance per unit length. This result is further verified by simulation. As shown in Fig. 4(b), the difference between $|Z_{1L}'|$ and $|Z_{1H}'|$ becomes more distinct if C_s is added. However, since C_s is in parallel with the transmission line, $|Z_{3L}|$ decreases to $|Z_{3L}'|$ as shown in Fig. 4(a), and thus more power is needed to sustain the low-band oscillation. The oscillation frequencies f_L and f_H are also shifted down due to the loading effect of C_s .

The mode-switching SWO can also be approximated by a lumped model which is simply a second-order LC tank. After calculating the impedance seen from Nodes T_1 and T_3 and utilizing the notch-peak cancellation concept [2], it can be proved that adding capacitance at Node T_3 can help suppress unwanted oscillation and hence stabilize desired oscillation, which is consistent with the results from the analysis above with a distributed model.

Fig. 4 Impedance with and without stabilization technique seen at Nodes (a) T_3 and (b) T_1

IV. CIRCUIT IMPLEMENTATION

Fig. 5 shows the schematic of the proposed dual-band standing-wave VCO with the stabilization technique. The differential transmission line is implemented by a differential micro-strip line with the topmost and thickest Metal 8 as the signal line and Metal 1 as the ground plane. The width is optimized to be 16 μm to maximize the quality factor Q at the frequency in the middle of the range. Frequency tuning is done by two varactors C_{v1} ($\sim 40\text{fF}$) and C_{v3} ($\sim 300\text{fF}$) located at Node T_1 and Node T_3 , respectively. Large C_{v3} is used to ensure no frequency jumping while small C_{v1} and no varactor at Node T_2 are employed to reduce the effective loading capacitance at Node T_1 . Three equally-spaced NMOS cross-coupled negative gm cells g_{m1} , g_{m2} , and g_{m3} , with an additional g_{m1}' at Node T_1 are utilized to compensate the losses of the transmission lines and varactors. Each of the gm cells is a current-biased gm cell which can be fully turned on/off by switching the bias current. Series switches are avoided in the high-frequency signal path to minimize attenuation. In the presence of the capacitance contributed from varactors and parasitics, the length of the transmission line is chosen to be 260 μm to obtain the desired output frequency.

Dual-band operation is achieved by switching different combinations of negative gm cells at different nodes. When g_{m1} , g_{m2} , and g_{m3} are activated by turning on their associated bias currents I_{b1} , I_{b2} , and I_{b3} , energy is injected

to all the three nodes T_1 , T_2 , and T_3 . As a result, the fundamental-mode standing-wave is excited, and the low-band oscillation is obtained. In this mode, the three negative gm cells are distributed to reduce the effective parasitic capacitance at the output. They are also scaled to be $g_{m3} = 2g_{m2} = 2g_{m1}$ to save the power consumption [4]. On the other hand, when g_{m1} and g_{m1}' are activated by turning on the bias currents I_{b1} and I_{b1}' , energy is only injected to Node T_1 . As such, the third-mode standing-wave is excited, and the high-band oscillation is obtained. The loading capacitance at Node T_3 looks smaller at Node T_1 due to the loss of the transmission line [5], but the impedance is still low that both g_{m1} and g_{m1}' are needed for high-band oscillation.

Fig. 5 Proposed dual-band standing-wave VCO

V. EXPERIMENTAL RESULTS

A dual-band standing-wave VCO prototype operating at 24 GHz and 60 GHz in a 0.13 μm CMOS process for wireless communication systems is implemented with proposed stabilization technique. With 0.8V and 1.2V supply voltages, the frequency tuning curves are measured as shown in Fig. 6.

Fig. 6 Measured frequency tuning curves with 0.8V and 1.2V supply voltages

For low-band phase noise measurement, the output is measured directly with Agilent E4440A spectrum analyzer, and the phase noise is -120dBc/Hz at 10MHz offset while drawing a current of 14mA from a 0.8V supply, as shown in Fig. 7(a).

For high-band phase noise measurement, the output signal is first down-converted by Ducommun V-band balanced mixer with LO input at 50 GHz and then measured with spectrum analyzer. With current consumption of 20mA from a 1.2V supply, the phase noise is -114dBc/Hz at 10MHz offset, as shown in Fig. 7(b).

Fig. 7 Measured phase noise plots in (a) the low-band mode at 24 GHz and (b) the high-band mode at 60 GHz

The performance and comparison with recent published state-of-art dual-band VCOs and MMW VCOs are summarized in Table 1. The core area of the implemented VCO is 0.31mm×0.16mm, as shown in Fig. 8.

VI. CONCLUSION

In this paper, a MMW mode-switching technique with standing-wave architecture is presented. Based on the detailed stability analysis with a distributed model, a stabilization technique is proposed and demonstrated for a dual-band MMW VCO with good performance. The

oscillation frequencies are much higher than existing dual-band VCOs while the performance is comparable to existing single-band MMW VCOs.

Fig. 8 Die photo

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TABLE 1 PERFORMANCE SUMMARY AND COMPARISON

Ref.	This Work		[2]		[4]	[5]	[6]
Technology	0.13 μ m CMOS		0.18 μ m CMOS		0.18 μ m CMOS	90 nm CMOS	90 nm CMOS
f_{osc} [GHz]	24	60	4.2	10	40	75	58.4
FTR	10.80%	7.20%	42.00%	18.00%	20.00%	1.30%	9.32%
PN [dBc/Hz]	-120@10M	-114@10M	-116@1M	-112@1M	-100@1M	-108@10M	-91@1M
Supply [V]	0.8	1.2	1.0	1.0	1.5	1.45	0.7
P_{diss} [mW]	11	24	6	10	27	8	8.1
FoM [dB]	-177	-176	-181	-182	-178	-176	-177