A 0.6 V 10 GHz CMOS VCO Using a Negative-Gm Back-Gate Tuned Technique

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***Abstract*—Without an extra on-chip accumulation-mode MOS varactor, a voltage-controlled oscillator (VCO) using a negative-transconductance back-gate tuned technique is demonstrated in a standard 0.18 μm CMOS process to achieve low-voltage, wide-range and high-frequency designs. Employing the varied p-n junction capacitance and the varied transconductance in the intrinsic-tuned regime, the VCO provides the tuning range of 9.95 to 11.05 GHz at a 0.6 V supply and dissipates below 4.35 mW. At 11 GHz carrier frequency, the measured phase noise is -110.4 dBc/Hz at a 1 MHz offset.**

***Index Terms*—Back-gate MOS, varied p-n junction capacitance, varied transconductance, voltage-controlled oscillator (VCO).**

I. Introduction

With the continuous growth of wireless communications, recently interest in LC-based voltage-controlled oscillators (VCOs) has surged for the demand of high performance solutions in the radio-frequency (RF) design. In addition to low-phase-noise considerations, two designed issues are more interesting: one is the low-voltage low-power design and the other is the high-frequency wide-range design. In order to maximize both the tuning range and the operating frequency, the capacitances existing in the LC tank become crucial design factors. Nevertheless, it suffers from a trade-off between the dynamic range and the operation frequency. To achieve higher operation frequency, the varied capacitance can be embedded in the cross-coupled transistors for LC-type VCOs. The back-gate tuned VCOs with an intrinsic p-n junction capacitor have been adopted to replace the use of conventional MOS varactors and achieve MMW operation [1], as shown in Fig. 1(a). This back-gate controlled technique can also dynamically change the threshold voltage of the MOSFET to lower the operating supply voltage [2]. However, the transconductance provided by the back-gate devices to sustain oscillation also vary with the control voltage, resulting in violent variations for the start-up condition of oscillation. In this letter, a 0.6 V 10 GHz mutual-coupled VCO using a negative-transconductance back-gate tuned technique is demonstrated 0.18 μm CMOS process, in which the tuning technique combines the varied junction capacitance with its varied transconductance on the back-gate transistors and the employed mutual-coupled transconductance can mitigate the variations for the start-up condition.

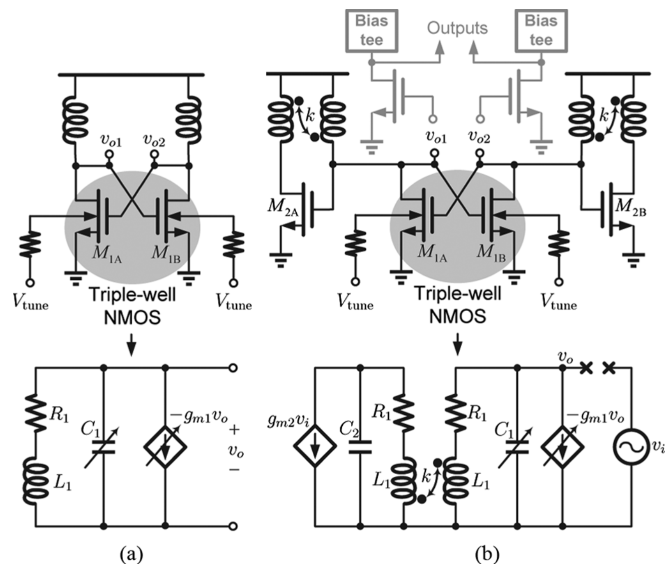

Fig. 1. Back-gate tuned VCOs and half circuits: (a) the conventional schematic in [1] and (b) the proposed scheme.

II. Circuit Design

A. Proposed Back-Gate Tuned VCO

In this work, we regard the parasitic capacitances as a varactor as well as its transconductance for frequency tuning in the cross-couple pair. As shown in Fig. 1(b), the negative resistance is generated by the circuit of the M_{2A} and M_{2B} pair in mutual feedback and the negative-transconductance circuit of the M_{1A} and M_{1B} pair in positive feedback. The tuned block made by the triple-well transistors M_{1A} and M_{1B} uses back-gate control to vary both the junction capacitances and the transconductances.

B. Operating Analysis

In Fig. 1(b), the oscillator could be viewed as two independent VCOs with the cross-coupling mechanism to synchronize the two half-circuit VCOs into differential operation. Because of $v_{o2} = -v_{o1}$, the equivalent half circuit with open-loop analysis can be exploited to construct the down circuit in Fig. 1(b), where g_{m1} and g_{m2} are transconductances of the M_{1A} and M_{1B} and M_{2A} and M_{2B} transistors, respectively and C_1 and C_2 are

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the effective capacitances. In the symmetrical-structure transformer, L_1's and R_1's represent the inductances and the series parasitic resistances to account for the losses respectively associated with the primary and secondary coils and the inductive coupling coefficient between the two coils is modeled by k. With the mutual-coupled feedback, the open-loop transfer function can be found as

$$T(s) = \frac{v_o}{v_i} = \frac{kg_{m2}L_1 s}{\alpha_4 s^4 + \alpha_3 s^3 + \alpha_2 s^2 + \alpha_1 s + 1 - g_{m1}R_1} \quad (1)$$

where

$$\begin{aligned}
\alpha_4 &= \left(1-k^2\right) L_1^2 C_1 C_2 \\
\alpha_3 &= 2R_1 L_1 C_1 C_2 - \left(1-k^2\right) g_{m1} L_1^2 C_2 \\
\alpha_2 &= L_1(C_1+C_2) + R_1^2 C_1 C_2 - 2g_{m1} R_1 L_1 C_2 \\
\alpha_1 &= R_1(C_1+C_2) - g_{m1}\left(L_1 + R_1^2 C_2\right).
\end{aligned}$$

Substituting $s = j\omega_0$ and $k \approx 1$, the real part in the denominator of (1) must drop to zero when oscillation occurs and the oscillation frequency can be calculated by

$$\omega_0 \approx \frac{1}{\sqrt{L_1(C_1+C_2)}} \cdot \sqrt{\frac{1-g_{m1}R_1}{1+\frac{R_1^2}{L_1}\frac{C_1C_2}{(C_1+C_2)} - g_{m1}R_1\frac{2C_2}{(C_1+C_2)}}}. \quad (2)$$

With variables of C_1 and g_{m1}, (2) can be rewritten as: $\omega_0 = \omega_T(C_1)\cdot\gamma(C_1, g_{m1})$, where ω_T is the intrinsic resonance frequency with a lossless tank for $R_1 = 0$ and γ is used to represent the frequency-extended factor. Provided $C_2 < C_1$ and $g_{m1}R_1 < 1$ in the typical design case, we can find that the values of differential terms of ω_T and γ in (2), e.g., $\partial\omega_T/\partial C_1$, $\partial\gamma/\partial C_1$ and $\partial\gamma/\partial g_{m1}$, are all negative. It is apparent that the oscillation frequency range can be extended because the values of ω_T and γ both decrease as the variables of C_1 and g_{m1} increase.

In order to give a quick and intuitive understanding with mutual-coupled contribution, we assume C_2 is relatively small for simplicity and (1) can be calculated as

$$T(s) = \frac{kg_{m2}L_1 s}{L_1 C_1 s^2 + (R_1 C_1 - g_{m1}L_1)s + 1 - g_{m1}R_1}. \quad (3)$$

From (3), the oscillation frequency can be calculated by: $\omega_0 = \sqrt{(1-g_{m1}R_1)/(L_1C_1)}$. To ensure that oscillations will start, one should set the magnitude of the loop gain to slightly greater than unity. This can be achieved by

$$g_{m1} + kg_{m2} \geq \frac{R_1 C_1}{L_1}. \quad (4)$$

It implies that both g_{m1} and g_{m2} can provide the required condition for a sustained oscillation while the conventional back-gate tuned topology in Fig. 1(a) provides only g_{m1} for the start-up oscillation.

However, when the size of transistors in Fig. 1(a) needs to be enlarged for wide tuning, g_{m1} and the operating current exist violent variations in different start-up oscillations. As a result, the changes on g_{m1} and the operating current will affect the performance of the VCO, such as power dissipation, output amplitudes and even phase-noise characteristics. Comparatively, in this work the mutual-coupled g_{m2} can suppress the minimum value and variation of variable g_{m1} through small-size transistors. Table I shows the comparisons of the oscillation frequency and start condition between Fig. 1(a) and (b).

TABLE I
COMPARISON OF FIG. 1(A) AND THIS WORK

Topology	Fig. 1(a)	This work
Oscillation frequency	$\frac{1}{\sqrt{L_1C_1}}$	$\sqrt{\frac{1-g_{m1}R_1}{L_1C_1}}$
Start condition	$g_{m1} \geq \frac{R_1C_1}{L_1}$	$g_{m1} + kg_{m2} \geq \frac{R_1C_1}{L_1}$

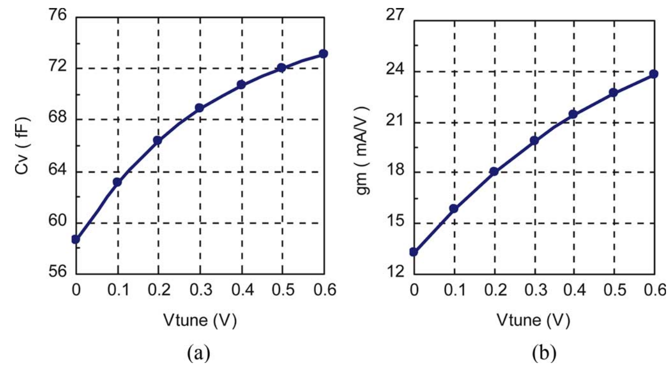

Fig. 2. Tuned characteristics of the triple-well transistor as a tuned device: (a) capacitance and (b) transconductance.

C. Tuned-Range Enhancement

Since the required minimum value of g_{m1} is reduced with small-size transistors, the tuned range of intrinsic capacitances will be reduced. In order to sustain a reasonable operating range, however, the tuned parameter g_{m1} can play a role of tuned-range enhancement. The effective capacitance C_v at the drain of $M_{1\mathrm{A}}$ (or $M_{1\mathrm{B}}$) increases as the tuned voltage V_{tune} increases, as shown in Fig. 2(a). Note that C_v also includes the gate capacitance of $M_{2\mathrm{A}}$ (or $M_{2\mathrm{B}}$). With the intrinsic-tuned technique, the tuning ratio of $C_{1,\max}/C_{1,\min} = C_v(V_{\mathrm{tune}} = 0.6\ \mathrm{V})/C_v(V_{\mathrm{tune}} = 0\ \mathrm{V})$ can be approximately 1.25. In addition to the intrinsic-tuned capacitance, the effective transconductance exhibits the tuning ratio of $g_{m1,\max}/g_{m1,\min}$ approximating to 1.8, as shown in Fig. 2(b). From device parameters, $L_1 = 2$ nH, $C_2 = 36$ fF and $R_1 = 18\ \Omega$ are given for calculation, then from (2) and Fig. 2(b) we find that $\omega_T(C_{1,\min})/2\pi = 11.57$ GHz, $\omega_T(C_{1,\max})/2\pi = 10.77$ GHz, $\gamma(C_{1,\min},\ g_{m1,\min}) = 0.96$ and $\gamma(C_{1,\max},\ g_{m1,\max}) = 0.89$; that is, $f_{0,\max} = 11.11$ GHz and $f_{0,\min} = 9.59$ GHz. As a result, the tuning range can be extended with a factor of 1.9 when capacitance-tuned and transconductance-tuned techniques are employed together. Note that the operating frequency would be reduced according to $\gamma < 1$.

III. CHIP IMPLEMENTATION AND MEASUREMENT

The proposed VCO was fabricated in standard 0.18 μm CMOS technology. The open-drain buffer at each output [as shown in Fig. 1(b)] was employed to drive the 50 Ω input-impedance of testing instruments. Fig. 3 shows the microphotograph of the test chip with an area of $550 \times 650\ \mu\mathrm{m}^2$ including output buffers and I/O pads. To evaluate the circuit performance, the VCO was tested on an FR-4 PC board using Agilent E4407B spectrum analyzer for measurement. The

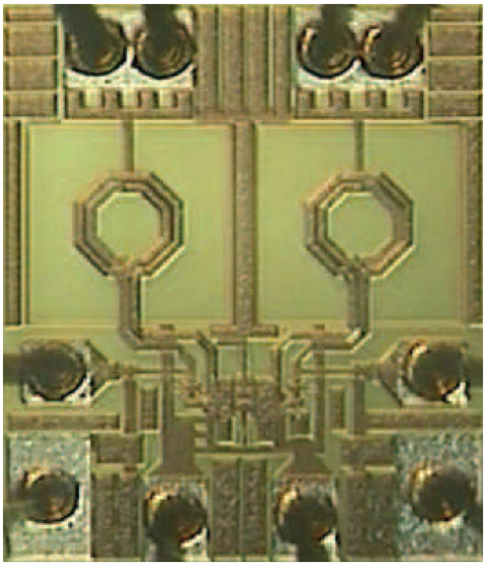

Fig. 3. Microphotograph of the fabricated chip.

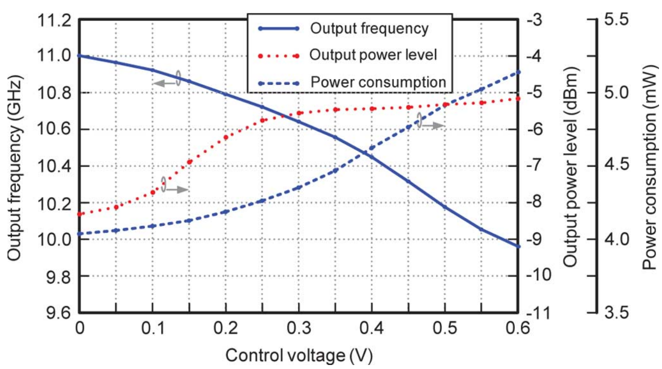

Fig. 4. Tuning characteristic, output power levels and power consumption of the tested circuit.

circuit operated at the supply voltage of 0.6 V and consumed the maximum current of 8.5 mA. The measured frequency tuning characteristic, buffered output power levels and power consumption with respect to the control voltage are shown in Fig. 4 and the tuning range is 10.5% (9.95 to 11.05 GHz) for the control voltage of 0 to V_{DD}. As can been seen, power dissipation and output power have the deviations of 1.2 mW and 3.1 dB, respectively, due to the operating current variations over the tuned range. The phase noise of the VCO at the maximum frequency is -110.42 dBc/Hz at 1 MHz offset, as shown in Fig. 5. The figure-of-merit (FOM) of VCOs is defined as

$$\mathrm{FOM} = 10\log\left[\left(\frac{\omega_0}{\Delta\omega}\right)^2 \cdot \frac{1}{L\{\Delta\omega\}\cdot P_{DC}}\right]. \quad (5)$$

Considering the tuning range, the FOM can be modified by: $\mathrm{FOM_T} = \mathrm{FOM} + 20\log(0.1\cdot\mathrm{FTR})$, where FTR is the frequency tuning range in percent. The worse FOM and $\mathrm{FOM_T}$ in this work are both 185.2 dB. Table II summarizes the overall specifications of the VCO with the prior works [3]–[8]. The proposed circuit can achieve only 0.6 V operation and its $\mathrm{FOM_T}$ is better than those in [3]–[6], [8].

IV. Conclusion

In conventional varactor-tuned VCOs, it may suffer from a trade-off between the tuning range and the operating frequency for low-voltage high-frequency designs. To satisfy both criteria, the designed VCO employs a negative-transconductance back-gate tuned technique, instead of the conventional accumulation-mode MOS varactor, which combines the varied p-n junction capacitance with its associated varied transconductance on the back-gate transistors. It is continuously tunable from 9.95 to 11.05 GHz at a 0.6 V supply.

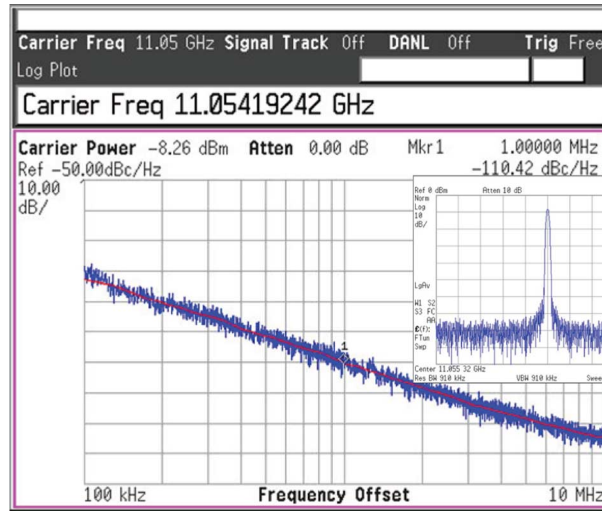

Fig. 5. Output spectrum and phase noise plot.

TABLE II
Comparison of VCO Performance

Ref.	Process (μm)	V_{DD}(V)/ P_{DC}(mW)	Freq. (GHz)	f_{tuned} (GHz)	FOM (dB)	$\mathrm{FOM_T}$ (dB)
[3]	0.18 CMOS	0.8/2.18	12.2	0.4	190.5	180.6
[4]	0.18 CMOS	1.8/8.1	11.55	0.63	183	177.7
[5]	p-HEMT	1/3	12.2	0.24	190.5	176.3
[6]	0.18 CMOS	1.8/6.84	11.22	0.3	182	170.5
[7]	0.18 CMOS	1.8/11.8	10.37	1.19	188	189.2
[8]	0.18 CMOS	1.8/5.8	9.83	1.1	165	165.9
This	0.18 CMOS	0.6/4.1	11.05	1.1	185.2	185.2

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