A High-Performance CMOS Voltage-Controlled Oscillator for Ultra-Low-Voltage Operations

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Abstract—In this paper, a novel circuit topology of voltage-controlled oscillators (VCOs) suitable for ultra-low-voltage operations is presented. By utilizing the capacitive feedback and the forward-body-bias (FBB) technique, the proposed VCO can operate at reduced supply voltage and power consumption while maintaining remarkable circuit performance in terms of phase noise, tuning range, and output swing. Using a standard $0.18\text{-}\mu\text{m}$ CMOS process, a 5.6-GHz VCO is designed and fabricated for demonstration. Consuming a dc power of 3 mW from a 0.6-V supply voltage, the VCO exhibits a frequency tuning range of 8.1% and a phase noise of -118 dBc/Hz at 1-MHz offset frequency. With an FBB for the cross-coupled transistors, the fabricated circuit can operate at a supply voltage as low as 0.4 V . The measured tuning range and phase noise are 6.4% and -114 dBc/Hz , respectively.

Index Terms—Capacitive feedback, forward body bias (FBB), ultra-low power, ultra-low voltage, voltage-controlled oscillators (VCOs).

I. INTRODUCTION

WITH THE emerging applications such as wireless personal area networks (WPANs), wireless sensor networks and RF identifications (RFIDs), the development of low-cost and low-power RF integrated circuits (RFICs) for short-range communications have attracted great attention over the past few years. For the circuit implementations, the CMOS technology appears to be particularly well suited due to its unparalleled advantages in the fabrication cost and system-level integration. Unfortunately, the inherently low transconductance of the MOSFETs at higher frequencies have impeded the evolution of low-power designs to RF front-ends. In order to overcome the limitations, various design methodologies and circuit techniques have been proposed [1]–[4]. Among these approaches, low-voltage circuit operation is one of the most promising solutions. In addition to the power considerations, a reduced supply voltage is also an inevitable trend for CMOS designs as well. With the continuous shrinking in the transistor feature size, a proportional down-scaling of the supply is required to ensure the gate-oxide reliability [5]. The forecast of the supply voltage for CMOS circuits within the next decade [6] is shown in Fig. 1, where a low-voltage operation of 0.4 V is anticipated by the end of 2020. Therefore, there exists an urgent need to develop low-voltage circuit techniques for high-performance RFICs at multigigahertz frequencies.

Manuscript received July 10, 2006; revised November 20, 2006. This work was supported in part by the National Science Council under Grant 94-2220-E-002-026 and Grant 94-2220-E-002-009.

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Digital Object Identifier 10.1109/TMTT.2006.891471

Fig. 1. Forecast of the CMOS supply voltage by International Technology Roadmap for Semiconductors (ITRS) 2004.

Being a crucial part in the RF front-ends, the voltage-controlled oscillator (VCO) is considered as one of the most power-consuming components. For low-power applications, CMOS oscillators operating at a supply voltage lower than 1 V were reported [7]–[11]. However, most of the circuits suffer from reduced output swing and degraded phase noise due to the limitations on the supply voltage. In this study, a VCO topology suitable for ultra-low-power and ultra-low-voltage operations is presented. By incorporating the capacitive feedback and the forward-body-bias (FBB) technique in the cross-coupled CMOS VCO, remarkable circuit performance in terms of the phase noise, frequency tuning range, and output swing can be achieved. Using a $0.18\text{-}\mu\text{m}$ CMOS technology, a 5.6-GHz VCO is implemented for demonstration.

The remainder of this paper is organized as follows. Section II describes the proposed circuit topology and the design considerations. The circuit design and the experimental results of the 5.6-GHz CMOS VCO are presented in Sections III and IV, respectively. Finally, a conclusion is given in Section V.

II. PROPOSED VCO TOPOLOGY

The schematic of the proposed VCO is shown in Fig. 2. In order to reduce the required supply voltage and to eliminate additional noise contribution, the tail current transistor in a conventional cross-coupled VCO is replaced by on-chip inductors [12]. For an enhanced voltage swing under an ultra-low supply voltage, the capacitive-feedback technique is employed [13]. Due to the use of the on-chip inductor and the feedback loop established by C_1 and C_2 , the drain and source voltages can swing above the supply voltage and below the ground potential. Consequently, the output swing of the VCO is enhanced, leading

Fig. 2. Schematic of the proposed VCO topology.

Fig. 3. Simplified half-circuit model of the proposed VCO.

to a superior close-in phase noise. Since the varactors are employed in the source terminals of the cross-coupled transistors, a more effective controlled mechanism of the tank resonance is presented. Therefore, a reasonable frequency tuning range can be achieved even with a reduced voltage range for the controlled signal V_{ctrl} . To further investigate the proposed VCO, detailed circuit analysis is provided as follows.

A. Startup Conditions

In order to derive the startup conditions and the oscillation frequency, the equivalent half-circuit of the proposed VCO is shown in Fig. 3, where R_1 and R_2 represent the losses of the on-chip inductors L_1 and L_2 , respectively. Note that the losses of the inductors are typically modeled by a series resistance. In the equivalent circuit, the narrowband approximation is employed to simplify the analysis, and the shunt resistance R_1 and R_2 can be estimated by

$$R_1 \approx \omega^2 L_1^2 / R_{s1} \quad (1)$$

$$R_2 \approx \omega^2 L_2^2 / R_{s2} \quad (2)$$

where R_{s1} and R_{s2} are the equivalent series resistances of L_1 and L_2 , respectively. Besides, the transistor parasitic capacitances, which are much smaller than the values of C_1 and C_2 , are neglected. From the small-signal analysis, the transfer function between V_o and V_i is given by

$$\frac{V_o}{V_i} = \frac{-g_m(L_1 L_2 R_1 R_2 C_2 s^2 + L_1 L_2 R_1 s + L_1 R_1 R_2) s}{a_4 s^4 + a_3 s^3 + a_2 s^2 + a_1 s + a_0} \quad (3)$$

where

$$a_0 = R_1 R_2$$

$$a_1 = L_1 R_2 + L_2 R_1 + g_m L_2 R_1 R_2$$

$$a_2 = L_1 L_2 (1 + g_m R_2) + R_1 R_2 (L_1 C_1 + L_2 C_1 + L_2 C_2) \\ \approx R_1 R_2 (L_1 C_1 + L_2 C_1 + L_2 C_2)$$

$$a_3 = L_1 L_2 (R_1 C_1 + R_2 C_1 + R_2 C_2)$$

$$a_4 = L_1 L_2 R_1 R_2 C_1 C_2.$$

The circuit oscillates if the loop gain is unity, which corresponds to a voltage gain $V_o/V_i = -1$ at the oscillation frequency ω_0

$$\frac{-j g_m \omega_0 (-L_1 L_2 R_1 R_2 C_2 \omega_0^2 + j L_1 L_2 R_1 \omega_0 + L_1 R_1 R_2)}{a_4 \omega_0^4 - j a_3 \omega_0^3 - a_2 \omega_0^2 + j a_1 \omega_0 + a_0} = -1. \quad (4)$$

With proper arrangement, (4) yields

$$L_1 L_2 C_1 C_2 \omega_0^4 - \left[L_1 C_1 + L_2 (C_1 + C_2) - \frac{L_1 L_2 g_m}{R_2} \right] \omega_0^2 + 1 = 0 \quad (5)$$

and

$$[L_1 L_2 (R_1 C_1 + R_2 C_1 + R_2 C_2) - g_m L_1 L_2 R_1 R_2 C_2] \omega_0^3 \\ - [g_m R_1 R_2 (L_2 - L_1) + L_1 R_2 + L_2 R_1] \omega_0 = 0. \quad (6)$$

Provided $g_m L_1 \ll R_2 (C_1 + C_2)$ in typical design cases, (5) can be simplified as

$$L_1 L_2 C_1 C_2 \omega_0^4 - [L_1 C_1 + L_2 (C_1 + C_2)] \omega_0^2 + 1 = 0. \quad (7)$$

From (7), the oscillation frequency can be approximated by

$$\omega_0 \approx \sqrt{\frac{1}{L_2 C_2} + \frac{C_1 + C_2}{L_1 C_1 C_2}}. \quad (8)$$

Based on (6) and (8), the required transconductance g_m to sustain the oscillation is given by

$$g_m = \frac{L_1 L_2 (R_1 C_1 + R_2 C_1 + R_2 C_2) \omega_0^2 - L_1 R_2 - L_2 R_1}{R_1 R_2 (L_2 - L_1) + L_1 L_2 R_1 R_2 C_2 \omega_0^2}. \quad (9)$$

Assuming that $L_1 = L_2 = L_p$ and $R_1 = R_2 = R_p$, (8) and (9) can be simplified as

$$\omega_0 = \sqrt{\frac{1}{L_p} \left(\frac{1}{C_1} + \frac{2}{C_2} \right)} \quad (10)$$

$$g_m R_p = 1 + \frac{4(C_1/C_2)^2}{1 + 2(C_1/C_2)}. \quad (11)$$

From (10), it is clear that the variation of C_2 has more influence on ω_0 than C_1 does. As a result, the capacitance C_2 is realized by varactors in this design, leading to a reasonable tuning range for the VCO under ultra-low-voltage operations. Note that a large capacitance value for C_1 is desirable to enhance the frequency tuning range. However, the required transconductance to sustain the oscillation increases with the ratio of C_1 and C_2 . Therefore, a design tradeoff is established between the frequency tuning range and the power consumption of the VCO by the circuit parameter C_1/C_2 . Once the capacitance ratio is determined, the required transconductance of the cross-coupled

Fig. 4. Required transconductance to satisfy the startup conditions for various inductor Q factors.

Fig. 5. Conceptual illustration of the drain and source voltage due to the capacitive feedback.

transistors can be estimated by (11). Assuming an inductance value of 2 nH and an oscillation frequency of 5 GHz, the required transconductance as a function of C_1/C_2 for various inductor quality factors are shown in Fig. 4, which provides useful design guidelines for the proposed circuit topology.

B. Output Voltage Swing

In the proposed VCO circuit, a capacitive feedback is formed by capacitors C_1 and C_2 . Due to the in-phase relationship provided by the capacitive feedback and the use of on-chip inductors, the drain and source voltage can swing above the supply voltage and below the ground potential, as illustrated in Fig. 5. Consequently, the close-in phase noise benefits from the enhanced voltage swing at the VCO output. To evaluate the performance enhancement of this technique, the output voltage swing of the VCO is derived from the time-domain waveform of the drain current $I_1(t)$, as shown in Fig. 6. For simplicity, the periodic drain current is modeled by a square wave with a period of $T_0 = 2\pi/\omega_0$ and an amplitude of I_0 . Note that, at the quiescent point, the transistors are biased at $V_{D1} = V_{D2} = V_{DD}$ and $V_{S1} = V_{S2} = 0$. The maximum drain current $I_1(t)$ occurs when the gate voltage V_{G1} reaches its peak value. Assuming that the amplitude of output oscillating signal is A , the gate voltages V_{G1} and V_{G2} are $(V_{DD} + A)$ and $(V_{DD} - A)$, respectively, while the

Fig. 6. Actual and modeled drain current waveform of the transistor M_1 .

Fig. 7. Predicted and simulated peak-to-peak amplitude at the VCO output.

source voltages V_{S1} and V_{S2} can be obtained by the voltage divider of C_1 and C_2 . Thus, I_0 is approximated by the maximum drain current with the transistor operating in the nonsaturated region

$$I_0 \approx \mu_n C_{ox} \frac{W}{L} \left[(V_{DD} + A + nA - V_t)(V_{DD} - A + nA) - \frac{1}{2}(V_{DD} - A + nA)^2 \right] \quad (12)$$

where $n = C_1/(C_1 + C_2)$ and V_t is the threshold voltage of the MOSFET. From the Fourier series of $I_1(t)$, the fundamental current component is given by

$$I_1(t)|_{\text{fundamental}} \approx \frac{2}{\pi} I_0 \sin(\omega_0 t) \quad (13)$$

and the fundamental voltage amplitude is

$$A \approx \frac{2}{\pi} I_0 R_p. \quad (14)$$

where R_p is the load resistance. From (12) and (14), a simplified expression of the VCO output swing is given by

$$A \approx (1 + C_1/C_2) \cdot V_{DD}. \quad (15)$$

According to (15), the predicted VCO output amplitude as a function of C_1/C_2 is depicted in Fig. 7 along with the values obtained from the circuit-level simulation. It is obvious that an output swing significantly larger than the supply voltage can be achieved by the proposed capacitive feedback.

Fig. 8. Simulated phase noise at various temperatures with an offset frequency of 1 MHz.

Fig. 9. I - V characteristics of the MOSFET with and without FBB.

Fig. 10. Simulated threshold voltage and drain current of the MOSFET with FBB.

C. Phase Noise

Since the proposed VCO is operated under an ultra-low dc voltage, the cross-coupled transistors are potentially biased in the weak-inversion region. Therefore, the drain noise of the transistors is no longer dominated by the thermal noise as [14]

$$\frac{i_{n,d}}{\Delta f} = 4kT\gamma g_{d0} \quad (16)$$

where g_{d0} is the channel conductance with $V_{GS} = 0$, and γ is the thermal noise coefficient. Instead, the drain noise is expressed

Fig. 11. Proposed methodology for the design of the low-voltage VCO.

TABLE I
CIRCUIT PARAMETERS OF THE VCO

Devices	Design Values	
M_1, M_2	35 μm / 0.18 μm	
L_1, L_2	L_p	1.5 nH
	Q	12
C_1	0.68 pF	
C_2	1~2 pF	
C_1/C_2	0.34~0.68	
g_m	8 mA/V	

as [15]

$$\frac{i_{n,d}}{\Delta f} = \frac{4kT\gamma g_{d0}}{1 + \exp\left(-\frac{V_{GS} - V_t}{2mV_T}\right)} + \frac{2qI_D}{1 + \exp\left(\frac{V_{GS} - V_t}{2mV_T}\right)} \quad (17)$$

where I_D is the drain current, V_T is the thermal voltage, and m is the weak inversion slope factor. From (17), $i_{n,d}/\Delta f$ includes not only the contribution from the thermal noise, but also that of the shot noise since the drain current consists of both drift and diffusion components as the gate overdrive $V_{GS} - V_t$ approaches V_T . Due to the fact that $2qI_D$ is generally greater than $4kT\gamma g_{d0}$ for low-power operation, the cross-coupled transistors may contribute more noise to the LC tank as the supply voltage decreases. Therefore, it imposes a fundamental limitation on the phase noise of the VCO for ultra-low-power and ultra-low-voltage applications.

Fig. 12. Microphotograph of the fabricated VCO.

Fig. 13. Measured and simulated tuning characteristics of the VCO with a supply voltage of 0.6 V.

Fig. 14. Measured close-in output spectrum of the VCO with a supply voltage of 0.6 V.

For the conventional LC -tank VCOs operating at a reduced supply voltage, a significant degradation in the phase noise is inevitable due to the limited signal power at the output. In the proposed VCO topology, the output swing can be effectively enhanced by the capacitive feedback, leading to an improved close-in phase noise. Furthermore, due to the similarity to the Colpitts oscillator, the VCO circuit also benefits from the cyclostationary noise effect [16]. To further investigate the influence of the capacitive feedback, the simulated phase noise as a function of the capacitance ratio C_1/C_2 is shown in Fig. 8. It is noted that lower phase noise can be achieved by increasing the capacitance ratio. However, as shown in (11), the performance improvement in the phase noise is accompanied by an increase

Fig. 15. Measured and simulated VCO characteristics with a supply voltage sweeping from 0.5 to 0.7 V.

Fig. 16. Measured and simulated tuning characteristics of the VCO with a supply voltage and body bias of 0.4 V.

Fig. 17. Measured close-in output spectrum of the VCO with a supply voltage and body bias of 0.4 V.

in the required transconductance, which implies higher power consumption of the VCO circuit. In addition to the capacitance ratio, the temperature also influences the VCO phase noise. The simulated results at various temperatures are illustrated in Fig. 8 as well, indicating an average 4-dB degradation in phase noise as the temperature increases from 25 °C to 50 °C.

D. FBB

For deep-submicrometer MOSFETs, the threshold voltage V_t is no longer constant, but influenced by circuit parameters such as gate length, channel width, and drain-to-source voltage due

TABLE II
PERFORMANCE SUMMARY OF THE LOW-VOLTAGE AND LOW-POWER VCOs

	Unit	This Work		[7]	[19]	[20]	[8]	
Technology	—	0.18- μm CMOS		0.18- μm CMOS	0.18- μm CMOS	0.18- μm CMOS	0.18- μm CMOS	
Frequency	GHz	5.6		5.8	4.5	4.8	1.4	3.8
Tuning Range	%	8.1	6.4	8.9	4.4	4.3 ³	0	3.0
Phase Noise @1-MHz	dBc/Hz	-118	-114	-97	-100 ²	-120	-129	-119
V _{DD}	V	0.6	0.4	0.6	0.8	1.5	0.35	0.5
DC power	mW	3.0	1.1	0.7	—	3.0	1.5	0.6
Output Power Level	dBm	-1	-9	-10	—	1	—	—
FOM ¹	dBc/Hz	189	189	174	—	189	190	193

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 offset frequency is 500 kHz. ³ From 0 to 1.5 V.

to the short-channel and narrow-channel effects [17]. Typically, transistors with a large channel width and a minimum gate length exhibit a reduced V_t , which is preferable for low-voltage operations. In this VCO topology, the fundamental limitation on the supply voltage is imposed by the threshold voltage of the cross-coupled transistors. To further reduce the supply voltage, the FBB technique is adopted as shown in Fig. 9. For a MOSFET device, the threshold voltage is governed by the body effect as [18]

$$V_t = V_{t0} + \left(\sqrt{2qN_A \varepsilon_s / C_{ox}} \right) \cdot \left(\sqrt{|2\phi_F + V_{SB}|} - \sqrt{|2\phi_F|} \right) \quad (18)$$

where V_{t0} is the threshold voltage for $V_{SB} = 0$ V, ϕ_F is a physical parameter with a typical value of 0.3 V, N_A is the substrate doping, and ε_s is the permittivity of silicon. By applying a forward bias voltage to the body through a current-limiting resistor R_B , the effective threshold voltage is thus reduced while maintaining a minimum forward junction current between the body and the source terminals. The simulated effective threshold voltage and the drain current of a MOSFET with $W = 35 \mu\text{m}$ and $L = 0.18 \mu\text{m}$ are demonstrated in Fig. 10, indicating a threshold voltage reduction more than 100 mV due to the FBB technique.

III. CIRCUIT DESIGN

Fig. 11 shows the design procedure of the high-performance VCO. The circuit design starts with the inductors L_1 and L_2 . For simplification, both inductors are chosen to be identical as L_p , which are optimized for a high- Q factor at the frequencies of interest. Once the inductance is determined, the required value of $(1/C_1 + 2/C_2)$ can be obtained from the designated oscillation frequency ω_0 by (10)

$$1/C_1 + 2/C_2 = \omega_0^2 L_p. \quad (19)$$

By defining the capacitance ratio $k = C_1/C_2$, the values of C_1 and C_2 are given by

$$C_1 = (1 + 2k) / (\omega_0^2 L_p) \quad (20)$$

$$C_2 = (1 + 2k) / (k \omega_0^2 L_p). \quad (21)$$

The circuit performance of the VCO is strongly influenced by the design parameter k . As indicated in (11) and (15), the required transconductance of the cross-coupled transistors and the associated output swing increase with k . Considering the design specifications such as the phase noise, output swing, tuning range, and power consumption, the optimum k is thus determined and the circuit parameters including the capacitance values and the transistor aspect ratio can be calculated accordingly. Finally, a global optimization is performed and design iterations may be needed to satisfy the required circuit performance. Following the design procedures, the parameters of the VCO design are tabulated in Table I. Based on the narrow-band approximation in (1) and (2), the calculated value of R_p is 633Ω in this particular case. The resulting values for $g_m L_p$ and $R_p(C_1 + C_2)$ are 1.2×10^{-11} and 1.06×10^{-9} , respectively, which validate the assumption $g_m L_p \ll R_p(C_1 + C_2)$ in the theoretical derivation of the proposed VCO.

IV. EXPERIMENTAL RESULTS

The proposed VCO is implemented in a standard 1P6M 0.18- μm CMOS process provided by a commercial foundry. With an optimum layout for the RF performance, the n -channel MOSFETs in the deep n-well exhibit a maximum oscillation frequency f_{max} up to 60 GHz. As for the passive components, top AlCu metallization layer of 2- μm thickness is available for on-chip inductance while metal-insulator-metal (MIM) capacitors with oxide intermetal dielectric are also provided.

Fig. 12 shows a microphotograph of the fabricated circuit with a chip area of $0.55 \times 0.9 \text{ mm}^2$ including the pads. In order to enhance the quality factor and to ensure the layout symmetry for fully differential operations, center-tapped spiral inductors

with a metal width of 15 μm are employed for the required inductance L_1 and L_2 . In addition, the accumulation-type devices are used as the varactors. For test purposes, the VCO outputs are buffered by open-drain nMOS transistors to drive the external 50- Ω load. To evaluate the circuit performance, the chip was mounted on an FR4 test board and measured with an Agilent E4407B spectrum analyzer.

When operating at a supply voltage of 0.6 V, the VCO core consumes a dc power of 3 mW. Fig. 13 shows the measured tuning characteristics. As the controlled voltage sweeps from 0 to 0.6 V, the 5.6-GHz VCO exhibits a frequency tuning range of 8.1%. The tuning range can be increased to 12.2% with a maximum controlled voltage of 1.8 V. Fig. 14 shows the close-in output spectrum of the VCO with a controlled voltage of 0.3 V. The measured output power and phase noise at 1-MHz offset are -1 dBm and -118.3 dBc/Hz, respectively. For ultra-low-voltage operations, the sensitivity to the supply voltage is also investigated. As V_{DD} sweeps from 0.5 to 0.7 V, the output characteristics of the VCO are illustrated in Fig. 15, exhibiting a reasonable performance deviation with respect to the variation in the supply voltage.

By employing the FBB technique, the VCO can operate at a supply voltage as low as 0.4 V. With a supply voltage and a body bias of 0.4 V, the VCO core consumes a dc power of 1 mW. Fig. 16 shows the measured tuning characteristics, indicating a 6.4% tuning range as the controlled voltage sweeps from 0 to 0.4 V. The close-in output spectrum of the VCO with a 0.2-V controlled voltage is shown in Fig. 17. The measured phase noise is -114.4 dBc/Hz at 1-MHz offset, while the output power is -9 dBm. The performance of the VCO is summarized in Table II along with the results from previously published low-voltage and low-power CMOS VCO circuits.

V. CONCLUSION

Using a 0.18- μm CMOS technology, a VCO operating at a reduced supply voltage is presented. The performance of the fabricated circuit is characterized with a supply voltage from 0.6 to 0.4 V. Due to the use of the capacitive feedback and the FBB technique, significant performance improvement in terms of output swing, frequency tuning range, and phase noise is demonstrated, making it extremely attractive for ultra-low-power and ultra-low-voltage RF applications.

ACKNOWLEDGMENT

The authors would like to thank the National Chip Implementation Center, Hsinchu, Taiwan, R.O.C., for chip fabrication, and K.-S. Chung, Realtek Semiconductor Corporation, Hsinchu, Taiwan, R.O.C., and Y.-C. Huang, National Taiwan University, Taipei, Taiwan, R.O.C., for technical support.

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