A Periodically Time-Varying Inductor Applied to The Class-D VCO for Phase Noise Improvement

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Abstract — This paper proposes a periodically time-varying (PTV) inductor applied to a class-D oscillator resulting in a 4.7VDD output swing that effectively improves the phase noise in both 1/f² and 1/f³ regions. Without extra switches, the PTV inductor is realized by reusing the original cross-coupled transistors in the class-D oscillator that semi-periodically cancel the magnetic flux between two coupled inductors. Fabricated in 65nm LP CMOS, the oscillator prototype operating at 0.35V measures FoMs of 192.5±1dBc/Hz at 1MHz offset frequency and 194.6±0.3dBc/Hz at 10MHz offset frequency, over a frequency tuning range from 3.67 to 4.43GHz.

Keywords—Voltage-controlled oscillator (VCO), class-D oscillator, phase noise, flicker noise upconversion, common-mode resonance, periodically time-varying inductor

I. INTRODUCTION

The emerging wireless communication systems demand local oscillator (LO) signals with low phase noise to support denser modulation schemes. As a critical component in the phase-locked loop (PLL), the voltage-controlled oscillator (VCO) not only dominates the out-of-band phase noise at a large offset frequency but also substantially contributes to the phase noise at the middle offset frequency close to the PLL bandwidth. Thus, it is critical to minimize VCO’s phase noise in both 1/f² and 1/f³ regions.

The conventional class-B VCO suffers from a reduced output voltage swing due to the voltage drop VSD across the tail current source or resistor, limiting its phase noise performance. Since the supply voltage VDD keeps decreasing as technology scales down while VSD remains relatively constant, the output swing limitation is exacerbated in the advanced CMOS technology. The class-D VCO [1] provides a simple yet effective and stable solution to break the output swing limitation at low VDD at the GHz frequency band, achieving low phase noise stably across a wide tuning range. Taking advantage of better switches with small turn-on resistance and parasitic capacitance offered by the advanced CMOS technology, the class-D VCO favored by the low-voltage applications removes the current control circuitry and reaches a peak oscillation amplitude of ~3.3VDD. Since the class-D VCO is only composed of an LC tank and two large-size cross-coupled transistors mimicking ideal switches, it also benefits a compact chip area and a simple design procedure.

To minimize the phase noise in the 1/f² region, the original class-D VCO [1] as shown in Fig. 1(a) adopts a single-turn inductor with a small inductance and a high-quality factor (Q). Although an excellent phase noise at the 1/f² region is recorded, the class-D VCO employing a single-turn inductor is sensitive to the flicker noise upconversion, especially when oscillating at the highest frequency. As reported by [1], the measured 1/f³ phase noise corner frequency is 2.1MHz at the highest oscillation frequency (fosc=4.8GHz). Since the large parasitic capacitors Cgs from the two cross-coupled transistors affect the tank common-mode (CM) impedance, employing a tail filter resonant at 2fosc becomes less effective to suppress the flicker noise upconversion [2]. The measurement shows that the 1/f³ phase noise corner frequency is still 1.5MHz with the help of a tail filter [1], which is higher than a typical bandwidth of an LC-VCO-based PLL.

On the other hand, using an LC tank with implicit CM resonance can help to suppress the flicker noise upconversion in the class-D VCO. To locate the tank CM resonant frequency at two times of the oscillation frequency fosc, [3] proposed an F2 inductor with a CM inductor to be one-fourth of the differential-mode (DM) inductance LDM, i.e., LCM = LDM/4. Resonant with the single-ended tank capacitors Cc, the F2 tank achieves a CM resonant frequency at 2fosc, which effectively reduce the 1/f³ corner frequency of the class-D VCO to below 100kHz. However, the F2 inductor relying on the magnetic flux cancellation to reduce LCM must be realized by a 2-turn inductor with small trace space, which inevitably results in a large LDM and contradicts the goal of low phase noise in the 1/f² region. Typically, the LDM of a 2-turn inductor with a small trace space can be three times larger than a single-turn inductor if both inductors have a similar radius. Assuming the Q is the same, using the F2 inductor will degrade the phase noise in the 1/f² region by 4.8dB according to Leeson’s equation.

Fig. 1. Conventional class-D oscillators using (a) a single-turn inductor [1] and (b) a 2-turn F2 inductor [3].
Aiming to improve the phase noise at both $1/f^2$ and $1/f^3$ regions for the class-D VCO, this paper proposes a periodically time-varying (PTV) inductor to boost the output voltage swing, compensating the phase noise degradation due to the use of an FET inductor. The PTV inductor can be realized by reusing the original cross-coupled transistors in the class-D VCO that semi-periodically cancel the magnetic flux between two coupled inductors without the need for extra switches. Compared with the class-D VCO using a single-turn inductor and a tail filter [1], our design using the proposed PTV inductor reduces the phase noise at 10MHz by 2.5dB at the highest frequency.

II. PROPOSED CLASS-D VCO USING A PTV INDUCTOR

A. Concept

The class-D VCO experiences two different working modes during one oscillation period [1]. During the first period ($T_1$), the switch $SW_a$ is on, and $SW_b$ is off [Fig. 2(a)]. Assuming that the tank inductor and $SW_a$ are lossless, the inductor current $I_{L1}$ ramps up linearly with time at a slope of $V_{DD}/L_{T1}$, where $L_{T1}$ is the tank inductance during $T_1$. Entering the second semi-period ($T_2$), $SW_a$ is off, and $SW_b$ is on. The tank inductor $L_{T2}$ and capacitor $C$ form a series resonant tank [Fig. 2(b)]. This series resonance renders $I_{L2}$ like a portion of sinusoid waveform with a frequency of $\omega_{tank} = 1/\sqrt{L_{T2}C}$. The amplitude of the sinusoid $I_{L2}$ during $T_2$ is [1]:

$$I_{pk} = \frac{2.27V_{DD}}{\omega_{tank}L_{T1}}$$  \hspace{1cm} (1)

By integrating $I_{L2}$ on the tank capacitor $C$, we obtain a sinusoidal output voltage $V_O$, with an amplitude:

$$V_{pk} = \frac{I_{pp}}{\omega_{tank}C}$$  \hspace{1cm} (2)

In the original class-D VCO, the tank inductance is the same during both $T_1$ and $T_2$ ($L_{T1} = L_{T2}$), resulting in $V_{pk} = 2.27V_{DD}$. Thus the output voltage swing can theoretically reach $V_{DD} + V_{pk} = 3.27V_{DD}$. The above analysis also reveals that $I_{pp}$ depends on $L_{T1}$ while $\omega_{tank}$ depends on $L_{T2}$. For the same $\omega_{tank}$, if we can make $L_{T1} < L_{T2}$, a large sinusoid current $I_{L2}$ will be integrated into a small tank capacitor $C$, resulting in a boosted $V_{pk}$:

$$V_{pk} = 2.27V_{DD} \cdot \frac{L_{T2}}{L_{T1}}$$  \hspace{1cm} (3)

Here, the $L_{T2}/L_{T1}$ represents the amplitude boosting factor compared with the original class-D VCO.

B. Implementation of the PTV Inductor

In practice, we do not want to employ extra switches to vary the inductance periodically since they will inevitably introduce more noise and in turn impairs the phase noise improvement. In the implementation, the function to change the inductance is absorbed by the original class-D switches $SW_a$ and $SW_b$ and the periodical inductance variation is realized by adding another inductor $L_d$ that magnetically couples ($k_1$) with the tank inductor $L_d$ as shown in Fig. 3(a).

Using the equivalent model of the coupled inductors as shown in Fig. 3(b), we can derive the equivalent tank inductance in $T_1$ and $T_2$. When $SW_a$ is on, and $SW_b$ is off, the left $L_d$ is in the semi-period $T_1$ while the right $L_d$ is in the semi-period $T_2$ [Fig. 4(a)]. During the class-D operation, the current flows through the switch mainly during a short time when the switch transits between on- and off-states. During the rest of the time, most of the tank current circulates between the tank inductors and capacitor, i.e., $I_{L1} \approx -I_{L2}$. Thus, we can assume no current flowing into $L_d$ through $SW_a$ during most time of $T_1$. Since we do not intend to add any capacitors at the $V_O$ node, the capacitor connected to $L_d$ that comes from the parasitic capacitors of the switch is small, and the current from it can be ignored. In this case, the induced
current in the $L_d$ can be ignored ($I_{md} \approx 0$) and $I_{ms}$ just equals to $-(M/L_d)I_{L1}$, where $M = k_1\sqrt{L_d}L_a$ is the mutual inductance induced by the magnetic coupling. This $I_{ms}$ serves to pull down $V_C$ to a negative voltage of $-sM/L_{L1}$, which increases the voltage drop across the left $L_d$ to $V_{DD} + sM/L_{L1}$ and boosts the current flowing through the left $L_d$ during $T_1$ to $I_{L1} = (V_{DD} + sM/L_{L1})/sL_d$. Then $I_{L1}$ can be calculated as:

$$I_{L1} = \frac{V_{DD}}{s(L_d - M)}$$

(4)

It indicates that the equivalent tank inductance $L_{T1}$ is reduced to $L_d - M$ with the help of the coupled inductor $L_a$.

The equivalent tank inductance $L_{T2}$ can be founded by analyzing the left $L_d$ in Fig. 4(b) that stays in $T_2$. Since the $SW_d$ is off, the left $L_d$ and $L_a$ are electrically isolated. Again, by ignoring the current from the small parasitic capacitor connected to $L_a$, there is no induced current in the right $L_d$ ($I_{md} \approx 0$), resulting in $L_{T2} = L_d$. With the help of the $SW_d$ and $SW_b$, the coupled inductors function like a PTV inductor without the need for extra switches, offering two different equivalent tank inductances in every semi-period. The amplitude-boosting factor is now decided by the coupled inductor:

$$\frac{L_{T2}}{L_{T1}} = \frac{1}{1 - k_1\frac{L_a}{L_d}}$$

(5)

Although the class-D VCO with a PTV inductor looks schematically similar to the transformer-feedback VCO (TF-VCO) firstly proposed in [4], their working principles are different. Yet, the TF-VCO operating in the class-B mode suffers from a degraded tank $Q$ when the cross-coupled transistors enter into the deep triode region at a high output swing, nullifying the phase noise improvement obtained from the oscillation amplitude boosting. While in the class-D operation where the cross-coupled transistors with large sizes are adopted, a high output swing helps further to reduce the turn-on resistance of the cross-coupled transistors in $T_1$ and shorten the switch’s transition time between on- and off-states. These all aid in suppressing the noise contribution from the cross-coupled switches.

C. VCO and $L_d$ Tank Design

Fig. 5(a) shows the detailed schematic of the class-D VCO using a PTV inductor. A 5-bit binary switched-capacitor array (SCA) and AMOS varactors are employed for coarse and fine frequency tuning, respectively. A fixed differential capacitor of 130fF is entailed to improve the phase noise at the highest frequency [1].

For a given $L_d$, increasing either $k_1$ or $L_a$ helps to achieve a large amplitude-boosting factor according to (5). Since a large $L_d$ reduces the loop gain, which impairs the startup condition, we focus on maximizing $k_1$ through the transformer design by stacking $L_d$ atop of $L_a$ as shown in Fig. 5(b). According to the EM simulation, $L_d = 600\mu$H, $L_a = 192\mu$H and $k_1 = 0.87$, resulting in $L_{T2}/L_{T1} = 2$. According to (3), $V_{pk} = 4.54V_{DD}$, and the output swing is boosted to $5.54V_{DD}$. Compared with the original class-D VCO with an output swing of $3.27V_{DD}$, our design can theoretically reduce the phase noise by 4.6dB due to the enlarged output swing according to Leeson’s equation. At $V_{DD} = 0.35V$, the simulated output swing is $4.7V_{DD}$ as shown in Fig. 6(a), which is smaller than the prediction due to the loss from the LC tank and switches.

The stacked transformer also enables a larger coupling coefficient $k_2$ within the two turns of $L_a$, resulting in a small CM inductance $L_{d,CM}$ of 170nH due to the magnetic-flux cancellation. As verified by the simulation results [Fig. 5(b)], the routing path (AB) connecting the center tap of $L_a$ to the ground negligibly affects the DM or CM inductance of $L_d$. Since there is a small differential capacitor in the tank when all the switches in the SCA are turned off, the $L_{d,CM}$ is designed to be larger than one-quarter of $L_d$ to locate the CM resonance at the highest frequency. The low-Q CM inductance secures a low $1/f^2$ phase noise even when the ratio of CM-to-DM resonant frequencies departs from two.

The simulated $Q$s of $L_d$ and $L_a$ are 20 and 6 at 4GHz, respectively. The phase noise contributed from the loss of $L_a$ is less critical since $L_a$ is 3x smaller than $L_d$ and only contributes to the output phase noise during $T_1$.  

III. MEASUREMENT RESULTS

The class-D VCO using a PTV inductor is prototyped in 65nm LP CMOS technology with a standard supply voltage of 1.2V, as shown in Fig. 6(b). The VCO occupies a compact area of 0.12mm$^2$, including a decoupling capacitor. Fig. 7 plots the phase noise measured at 3.67GHz and 4.44GHz at $V_{DD} = 0.35V$, which meet the cellular standards with enough...
power consumption and degrade the FoM at low oscillation frequencies. However, the performance at the highest frequency will not be affected. Compared with [5] that can theoretically boost the output swing of the class-D VCO to 5.5 Vdd is presented. The boosted output swing and the F2 tank help to improve the phase noise in both 1/f^2 and 1/f^3 regions. The class-D VCO using the PTV inductor achieves low phase noise and high FoMs from 3.62 to 4.44GHz and occupies a compact area.

### IV. CONCLUSION

In this paper, a periodically time-varying (PTV) inductor that can theoretically boost the output swing of the class-D VCO is demonstrated. The boosted output swing and the F2 tank help to improve the phase noise in both 1/f^2 and 1/f^3 regions. The class-D VCO using the PTV inductor achieves low phase noise and high FoMs from 3.62 to 4.44GHz and occupies a compact area.

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