A Varactor Configuration Minimizing the Amplitude-to-Phase Noise Conversion in VCOs

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Abstract — Amplitude-to-phase-noise conversion due to varactors can severely limit the close-in phase noise performance in LC-tuned oscillators. This work proposes a rigorous analysis of this phenomenon, which highlights the fundamental limitations of single-ended tuned and differentially-tuned diode varactor configurations. The back-to-back varactor topology is identified as a suitable solution to linearize the tank capacitance. The amplitude to phase noise conversion is greatly attenuated and the $1/f^3$ phase noise is drastically reduced, without impairing the achievable tuning range. These results are validated through circuit simulations in an existing 0.35-μm CMOS technology.

Index Terms — Voltage Controlled Oscillator, phase noise, flicker noise up-conversion, diode varactor.
I. INTRODUCTION

Phase noise as well as tuning range is a critical issue in the design of a Voltage-Controlled Oscillator (VCO) for wireless transceivers. For instance, in cellular and wireless LAN applications the required tuning range of the local oscillator is typically ±10% of its central frequency. Such range is supposed to cover not only the operating frequency band, but also process spreads, temperature variations and incorrect estimation of parasitics. The wide tuning range, together with the supply reduction, suggest the use of a varactors featuring a steep C(V) characteristic.

However highly non-linear varactors make the output frequency dependent on oscillation amplitude [1] and so, any amplitude noise translates into phase noise. Figure 1 presents a conventional topology of integrated LC-tuned VCO. In this circuit, flicker noise associated to the current source generates flicker amplitude-modulation (AM) noise. The varactors convert this AM noise into frequency-modulation (FM) noise. This effect may be quantified by a conversion coefficient:

\[ K_{AM\rightarrow FM} = |\frac{\partial \omega}{\partial A}| \] 

which represents the sensitivity of the oscillation frequency to variations of the oscillation amplitude \( A \). The resulting phase noise spectrum, expressed in terms of Single-Sideband-to-Carrier Ratio (SSCR or \( \mathcal{L} \)), is [2]:

\[ \mathcal{L}(\omega_m) = \frac{K_{AM\rightarrow FM}^2 \cdot S_{AM}(\omega_m)}{2\omega_m^2} \] 

where \( S_{AM}(\omega_m) \) is the power spectral density of the amplitude noise at \( \omega_m \) offset from the carrier. 1/\( f \) amplitude noise can therefore causing 1/\( f^3 \) phase noise, thus limiting the VCO performance [3].
The insets in Fig. 1 show three varactor configurations, that can be used in the VCO. The most conventional topology is shown in inset (a). We will refer to this topology as single-ended-tuned VCO (SE-VCO). In order to improve the immunity to common mode disturbances, as ground bounces, in some cases the differential varactor configuration depicted in Fig 1(b) (D-VCO) is adopted [4-5]. A third possible configuration [see Fig 1(c)], typically employed to reduce distortion in RF tuned filters [6-7], uses varactors in back-to-back series connection (BB-VCO). In this paper, those three topologies are compared in terms of VCO flicker noise up-conversion and tuning range. In particular, in Section II the AM-to-PM conversion mechanism is analytically described for the single-ended tuned varactor configuration and for the differentially-tuned one. The back-to-back varactor VCO is introduced in Section III and the AM-to-PM conversion for this new oscillator topology is discussed. We demonstrate a substantial reduction of amplitude-to-phase noise conversion respect to the other two configurations, which leads to a great reduction of 1/f noise up-conversion. Simulation results are presented in Section IV; last, conclusions are drawn in Section V.

II. AM-TO-PM CONVERSION

In a previous work [2], the authors have described a method to derive a closed-form expression of the AM-to-FM conversion factor, given the varactor C-V characteristic, for the traditional SE-VCO. In the following, we will recall briefly this calculation. This problem can be studied referring to the simplified circuit in Fig. 2. The voltage $V(t)$ is not sinusoidal since when $V(t)$ increases, the capacitance grows and the instantaneous frequency slows down; when $V(t)$ decreases, the instantaneous frequency speeds up. Therefore, $V(t)$ is intrinsically not harmonic and it is well approximated accounting only for the first two harmonics:
\( V(t) = A \cos(\omega t) - B \cos(2\omega t) \). The quasi-sinusoidal approximation used in [8] neglects the presence of the second-order harmonic and leads to less accurate results. Since \( V(t) \) is periodic, the capacitance can be expanded in Fourier series as:

\[
C[V(t)] = c^{(0)} + \sum_{n=1}^{\infty} 2c^{(n)} \cos(n\omega t) \quad (3)
\]

Substituting the expressions of \( V(t) \) and \( C(t) \) into the differential equation describing the tank,

\[
\frac{1}{L} \int V(t) \, dt = C \frac{dV(t)}{dt} \quad (4)
\]

the frequency of oscillation can be expressed by means of an effective capacitance \( C_{\text{eff}} \) [2, 8]:

\[
\omega = \frac{1}{\sqrt{LC_{\text{eff}}}}
\]

\[
C_{\text{eff}} = c^{(0)} - c^{(2)} - 2\left(c^{(1)} - c^{(3)}\right)(B/A) \quad (5)
\]

For a p-n junction, the capacitance is usually written as:

\[
C_{V}(V_{R}) = \frac{\partial Q}{\partial V} = C_{J0} \left(1 + \frac{V_{R}}{V_{J}}\right)^{-m} \quad (6)
\]

where \( V_{R} \) is the reverse voltage across the diode. This capacitance expressed as a function of the voltage \( V \) across the inductor is: \( C_{V}(V_{R}) = C_{V}(V_{\text{TUNE}} - V) \) and it is represented in Fig. 2. By changing \( V_{\text{TUNE}} \) the capacitance curve shifts horizontally.
In order to derive a closed-form expression of the effective capacitance, it is convenient to approximate the $C(V)$ characteristic around the bias voltage $V = 0$ by means of a quadratic curve: $C(V) = C_0 + C_1 V + C_2 V^2$, where $C_0$ is the capacitance at $V_{TUNE}$ and $C_1$ and $2C_2$ are the first-order and the second-order derivative of Eq. (6) at $V_{TUNE}$. Inserting the expression of $V(t)$ in the quadratic $C(V)$, the coefficients $c^{(e)}$ of the Fourier series in Eq. (3) can be written as a function of $C_0$, $C_1$ and $C_2$. Thus, from Eq. (5) the effective capacitance results:

$$C_{eff} = C_0 + \frac{1}{4} C_2 (A^2 + B^2) - \frac{1}{2} C_1 B$$

(7)

Expressing $C_0$, $C_1$ and $C_2$ in terms of diode varactor parameters, the oscillation frequency is:

$$\frac{\omega}{\omega_0} = 1 - \frac{m(3-m)}{48 (V_T + V_{TUNE})^2} A^2$$

(8)

where $\omega_0 = 1/\sqrt{LC_0}$.

The $K_{AM-FM}$ factor can be obtained by differentiating Eq. (8):

$$K_{AM-FM} = \frac{\partial \omega_0}{\partial A} = \omega_0 \left[ \frac{m(3-m)}{24 (V_T + V_{TUNE})^2} A \right]$$

(9)

Let us apply the same procedure to the differential-tuned VCO in Fig. 1(b), whose tank is schematically depicted in Fig. 3. The tank capacitance is made of two p-n junctions connected in anti-parallel mode. Thus, the overall $C-V$ characteristic is given by summing the characteristics of both varactors, as shown in Fig. 3:

---

1 For any value of $V_{TUNE}$, the bias voltage is set by the inductor at $V = 0$. This means that when $V(t) = 0$ the reverse voltage across the diode is $V_{TUNE}$. 

---
\[ C(V) = \frac{1}{2} \left[ C_V (V_{TUNE} - V) + C_V (V + V_{TUNE}) \right] \] (10)

The factor 1/2 comes from the fact that the two varactors are halved with respect to varactor used in the single-ended tuned configuration in Fig. 2 to retain the same oscillation frequency. The resulting characteristic is an even function, which can be approximated by a parabola:

\[ C(V) = C_0 + C_2 V^2. \]

Since \( C(V) \) is an even function, also \( V(t) \) is an even function and contains no even-order harmonics. The presence of high-order odd harmonics can be neglected without any appreciable errors, thus: \( V(t) = A \cos(\omega t) \).

The effective capacitance follows again from Eq.s (5) with B=0:

\[ C_{\text{eff}} = c^{(0)} - c^{(2)} = C_0 + C_2 \frac{A^2}{4} \] (11)

where \( C_0 \) is the capacitance \( C_V \) at \( V_{\text{TUNE}} \), as in the previous case, while the parameter \( 2C_2 \) is the second-order derivative of the function \( C(V) \) in Eq. (10). The resulting oscillation frequency and the AM-to-FM conversion factor are:

\[ \frac{\omega}{\omega_0} = 1 - \frac{m(m+1)}{16(V_j + V_{\text{TUNE}})^2} A^2 \] (12)

\[ K_{\text{AM-FM}} = \omega_0 \left[ \frac{m(m+1)}{8(V_j + V_{\text{TUNE}})^2} A \right] \] (13)

The AM-to-FM conversion of the two different varactor configurations can be now quantitatively compared by evaluating the relative \( K_{\text{AM-FM}} \) factors. As a benchmark case, we considered an existing 0.35um CMOS technology. P+/n-well junctions, which can be used as varactors, features
a \( C-V \) curve described by Eq. (6) with \( m = 0.43 \) and \( V_f = 0.8 \text{V} \). The oscillation frequencies for the single-ended tuned VCO and for the differentially-tuned one, calculated from Eq. (8) and Eq. (12) respectively, are plotted as lines in Fig. 4(a). In both cases, the varactors are biased at 0 V (i.e \( V_{TUNE} = 0 \text{V} \)), which is the most critical situation for linearity, and \( (\omega_0/2\pi) \) is 1.6 GHz. Figure 4(b) shows that the conversion factor of the differential-tuned configuration is higher than that of the single-end tuning. This can be also analytically achieved by calculating the ratio between the conversion factors in Eqs. (9) and (13), which results: \( 3(m+1)/(3-m) = 1.7 \). Note that if the second harmonic of the voltage \( V(t) \) were not considered in Eq. (5), the same expression of the oscillation frequency and of the conversion factor would be obtained for the traditional tank and for the differentially-tuned one. In fact, \( B \) in the Eq. (7) has to be neglected and \( C_0 \) and \( C_2 \) results the same in both cases.

From circuit simulation, another drawback of the differentially-tuned VCO arises. The oscillation amplitude is limited by the clamping action of the diodes; in this case, for \( V_{TUNE} = 0 \), the single-ended oscillation amplitude is limited at 0.75V. In fact, considering again Fig. 3, \( V(t) \) is clamped by varactor \( C_{v1} \) for positive excursion and by varactor \( C_{v2} \) for negative excursion. That doesn’t happen in S-VCO where only when \( V(t) \) grows the clamping occurs; nevertheless, \( V(t) \) is not limited when it becomes negative. However, this implies a second harmonic distortion with a positive peak different from the negative one.

**III. BACK-TO-BACK VARACTOR CONFIGURATION**

Let us consider the varactor configuration represented in Fig. 1(c), where two p-n diode varactors are now connected back-to-back in series on each side of the tank.
The $C-V$ characteristic of the overall tank capacitance can be derived referring to the simplified circuit in Fig. 5. At the oscillation frequency the bias resistor is an open circuit and the two varactor capacitances may be considered in series, since $1/\omega C_v << R$. Note that the two varactors are biased at the same voltage, $V_{\text{TUNE}}$, being $V = 0 V$. Let us now consider the oscillation waveform $V(t)$ across the inductor and the varactors series. Since $V(t) = V_1 + V_2$, where $V_1$ is the direct bias voltage of varactor $C_{v_1}$ and $V_2$ is the reverse voltage of $C_{v_2}$, when $V(t)$ is positive, $C_{v_1}$ is driven forward biased while $C_{v_2}$ is driven reverse biased. The overall capacitance is dominated by the smaller one, which is mildly dependent on the applied voltage. Moreover, another effect contributes to make the overall capacitance constant: the residual non-linearity given by the curve $C_{v_1}(V_1)$ is mitigated, because $C_{v_2}$ is much lower than $C_{v_1}$ and then $V_1$ is only a small fraction of the voltage $V$. Figure 5 is an intuitive picture to explain the almost constant $C(V)$ characteristic. As $V_{\text{TUNE}}$ grows the two $C(V)$ curves departs from each other. In that figure we consider $V_1 \approx V_2 \approx V/2$ for sake of simplicity. That is true only when the amplitude of the applied voltage $V$ is small. Instead, for large amplitude of $V(t)$ and, for example, for $V(t)$ positive, it results $V_1 \gg V_2$ since $C_{v_2} >> C_{v_1}$. Nevertheless in Fig. 5 it’s intuitive that the tank topology helps in reducing the overall capacitance non-linearity.

In order to evaluate analytically the capacitance $C(V)$ of the capacitors series, it’s convenient to approximate the single varactor characteristic with a second order curve:

\[
\begin{aligned}
C_{v_1}(V_1) &= \frac{dQ_1}{dV_1} = a_0 + a_1 V_1 + a_2 V_1^2 \\
C_{v_2}(V_2) &= \frac{dQ_2}{dV_2} = a_0 - a_1 V_2 + a_2 V_2^2
\end{aligned}
\]  

(14)
The sign minus in the second of Eqs. (14) is due to the fact that $V_2$ is the reverse voltage across the varactor $C_{v2}$, while $V_1$ is the direct voltage across $C_{v1}$. Since the varactors are in series, they share the same charge. Integrating the voltage across the two varactors, $Q_1$ and $Q_2$ are given by:

$$
\begin{align*}
Q_1 &= \int_0^{V_1} C_{v1}(V_1)\,dV_1 = a_0 V_1 + \frac{1}{2} a_1 V_1^2 + \frac{1}{3} a_2 V_1^3 \\
Q_2 &= \int_0^{V_2} C_{v2}(V_2)\,dV_2 = a_0 V_2 - \frac{1}{2} a_1 V_2^2 + \frac{1}{3} a_2 V_2^3
\end{align*}
$$

(15)

Using the reverse series [6] and equating the two charges ($Q_1 = Q_2 = Q$), the voltages $V_1$ and $V_2$ in terms of the charge $Q$ are:

$$
\begin{align*}
V_1 &= \frac{Q}{a_0} - \frac{1}{2} a_1 \frac{Q^2}{a_0^3} + \frac{1}{2} a_0 a_2 \left( \frac{a_1^2}{2} - \frac{a_0 a_2}{3} \right) Q^3 + \\
V_2 &= \frac{Q}{a_0} + \frac{1}{2} a_1 \frac{Q^2}{a_0^3} - \frac{1}{2} a_0 a_2 \left( \frac{a_1^2}{2} - \frac{a_0 a_2}{3} \right) Q^3 + 
\end{align*}
$$

(16)

Since $V = V_1 + V_2$:

$$
V = \frac{Q}{a_0} + \frac{Q^2}{a_0^3} \left( \frac{a_1^2}{2} - \frac{a_0 a_2}{3} \right) Q^3 + 
$$

(17)

Using again the reverse series, the charge $Q$ is given by:

$$
Q = \frac{a_0}{2} V + \frac{a_2}{24} \left( \frac{a_1^2}{16 a_0} - \frac{a_0 a_2}{2} \right) V^3 + 
$$

(18)

Since $C = dQ/dV$, from Eq. (18) the total capacitance results:

$$
C = \frac{a_0}{2} + \frac{a_2}{8} \left( 1 - \frac{3 a_1^2}{2 a_0 a_2} \right) V^2 + 
$$

(19)
In order to achieve the same oscillation frequency as in the previous two cases, the varactors capacitance is twice the capacitance in the single-ended tuned VCO. It results:

\[
\begin{align*}
    a_0 &= 2C_0 = \frac{2C_J}{\left(1 + \frac{V_{TUNE}}{V_J}\right)^m} \\
    a_1 &= \frac{m}{(V_J + V_{TUNE})}a_0 \\
    a_2 &= \frac{m(m+1)}{2(V_J + V_{TUNE})^2}a_0
\end{align*}
\]  
(20)

As in the case of differential tuning, \(C(V)\) is even and the voltage \(V(t)\) may be considered harmonic. The effective capacitance may be evaluated from Eq. (11):

\[
C_{eff} = c^{(0)} - c^{(2)} = C_0 \left[1 + \frac{m-2m^2}{32(V_J + V_{TUNE})^2}A^2\right]
\]  
(21)

Thus, from Eq. (21) the oscillation frequency results:

\[
\frac{\omega}{\omega_0} = 1 - \frac{m(1-2m)}{64(V_J + V_{TUNE})^2}A^2
\]  
(22)

where \(\omega_0 = \frac{1}{\sqrt{LC}}\).

Differentiating Eq. (22), the AM-to-FM conversion factor is evaluated as:

\[
K_{AM-FM} = \omega_0 \left[\frac{m(1-2m)}{32(V_J + V_{TUNE})^2}A\right]
\]  
(23)
The curve calculated from Eq. (22) is plotted as line (BB-VCO) in Fig. 4(a), while the corresponding conversion factor is plotted in Fig. 4(b). Also for this configuration, the varactors are biased at 0 V and \( \omega_n/2\pi \) is 1.6 GHz. Note that, since the effective capacitance grows with the oscillation amplitude, the oscillation frequency decreases with A, as in the single-ended tuned VCO and in the differentially-tuned case. However, the dependence is weak and the \( K_{AM-FM} \) is less than 4 MHz/V for oscillation amplitude up to 1 V. With the technology parameters given in the previous section, the ratio between the AM-to-FM conversion factors in the single-ended tuning configuration and in the back-to-back one results: \( (4/3)(3-m)/(1-2m) = 25 \).

It’s interesting to note that the AM-to-FM conversion does not take place if \( m = 0.5 \). Furthermore it’s evident in Fig. 4(a) that the oscillation amplitude is not limited as in the differentially-tuned configuration. In fact, the voltage \( V \) across the series is distributed among the varactors but the forward biased component, which has the larger capacitance, has a small voltage drop across of its terminals.

**IV. CIRCUIT SIMULATION RESULTS**

This quantitative analysis has been validated against circuit simulations. Spectre has been run on the three oscillators employing the different varactor configurations in Fig. 1. The circuits have been designed in an existing 0.35-\( \mu \)m CMOS technology with 3-V voltage supply. The inductors are spiral coils giving 1.85 nH each and with a quality factor of 10 at 2 GHz, while the varactors have unit capacitance of 4.15 pF and a quality factor of 20 at the same frequency. The varactor \( C(V) \) characteristic is well described by Eq. (6) with \( m = 0.43 \) and \( V_j = 0.8V \); the \( C_{max}/C_{min} \) ratio is 2 considering a reverse voltage across the diode ranging from 0 to 3 V..
overall tank quality factor is about 7. For a fair comparison, each oscillator has been biased at the same current, 10 mA. The amplitude $A$ of the single-ended output is about 0.75 V at 1.75 GHz. The simulated tuning curves are plotted in Fig. 6(a). The same tuning range (about 32%, without load and output buffer) is achieved in each of the configurations. Figure 6(b) depicts the tuning constant $K_{VCO}$ calculated as the first derivative of the curves in Fig. 6(a). While the tuning voltage of SE-VCO and BB-VCO can range between 0 and $V_{DD}$, the differential tuning voltage\(^2\) of D-VCO can sweep between $-V_{DD}$ and $V_{DD}$. The differentially-tuned VCO features half the peak $K_{VCO}$ of the other two configurations, since it covers the same tuning range with twice the voltage range.

The oscillation frequency (for $V_{TUNE} = 0V$) resulting from circuit simulations has been plotted versus the single ended oscillation amplitude (squares in Fig. 4). The solid lines for SE-VCO and D-VCO fit well the circuit simulations. A discrepancy occurs at high amplitudes since the quadratic approximation of the $C(V)$ curve begins to fail. In fact, the difference between theory and simulations reduces when $V_{TUNE}$ is increased and the non-linearity reduced. Instead, the results for the back-to-back configuration differ from the predicted ones. This discrepancy may be ascribed to the varactor bias resistors, which cause the varactors to be not rigorously in series. The resistor value was chosen to not introduce additional phase noise. In fact, the resistor voltage noise modulates the varactor capacitance and gives rise to phase noise but a resistor of small value may degrade the tank quality factor. The contribute of the bias resistor to the varactors’ series quality factor is $2\omega C_v R_B$; where $C_v$ is the single varactor capacitance and the factor 2 comes from the series of the varactors; choosing $R_B = 1 k\Omega$ it results a quality factor of 200! However, despite

\(^2\) Adopting the same bias scheme described in [4], the common-mode voltage of the varactors can be set to $V_{DD}/2$. The differential tuning can still vary between $-V_{DD}$ and $V_{DD}$, however doing so, the two tuning voltages are always between 0 and $V_{DD}$.
\(1/\omega C_\gamma \ll 1 \, k\Omega\), the asymmetry introduced by the bias resistor makes the tank slightly different from the one considered in the previous analysis.

Let us justify this statement. In Fig. 7(a), the normalized oscillation frequency is plotted as function of single-ended oscillation amplitude for different value of the varactor parameter \(m\) (see Eq. 6) and for bias resistors of 1 \(k\Omega\). Increasing the varactor non-linearity, i.e. increasing \(m\), the dependence of oscillation frequency on oscillation amplitude grows. Furthermore, oscillation frequency always decreases with amplitude, despite the theory predicts (see Eq. (22)) that for \(m>0.5\) the frequency should increase with oscillation amplitude.

The same simulations were performed biasing the varactors with resistor of 10 \(M\Omega\) instead of 1 \(k\Omega\). The results are depicted in Fig. 7(b) as symbols, instead solid lines refer to theory. For \(m=1\) and for \(m=2\) the theoretical analysis well predicts the simulated results, and the small discrepancy may be justified considering that the quadratic approximation is not so suitable as the non-linearity increases. Instead, for \(m<1\) the simulated curves are far from theoretical lines, but they are very close to the dashed curve that was obtained with \(m=0\), i.e. with constant capacitors. Since the varactor non-linearity has been practically eliminated, other effects cause a dependence of the oscillation frequency on oscillation amplitude. In particular, the residual dependence can be justified accounting for the non-linear reactance due to the transconductor and to the parasitic n-well substrate diode of the varactors. Therefore, the simulations confirm that the varactor bias resistor is the main responsible of the non-perfect concordance between theory and circuit simulation as a reasonable value of resistor is chosen to bias the back-to-back varactor configuration.

Nevertheless, the AM-PM conversion factor of the back-to-back VCO for an amplitude of 0.75 V is reduced at least by a factor of four (or 12 dB) respects to the other two configurations.
According to Eq. (2), the lower sensitivity to amplitude variations is beneficial to reduce up-conversion of the flicker noise coming from the tail generator. Figure 8 shows the three spectra, obtained by Spectre simulations, at $V_{TUNE} = 0 \, V$, that is the most critical case for AM-PM conversion. $1/f^2$ phase noise is limited by the tank losses and by the oscillation amplitude [9] and does not depend by varactor topology, the $1/f^3$ phase noise of the three VCOs is instead different. The BB-VCO has less close-in phase noise by almost 16 dB with respect to the SE-VCO and over 20 dB with respect to the D-VCO. The corner frequency lays at 40 kHz, 1 MHz and 2 MHz, respectively for the BB-VCO, SE-VCO and D-VCO. Fig. 9 shows the $1/f^3$ phase noise at 10 kHz offset from then carrier for the BB-VCO and the SE-VCO as a function of the tuning voltage. Note that these simulations were performed at the same bias current along the whole tuning range. This means that the oscillation amplitude increases as the tuning voltage grows since varactor capacitance decreases and the quality factor increases. The reduction of up-converted flicker noise varies from 16 dB at low tuning voltage, where the varactors non-linearity is the greatest, to 3 dB at high tuning voltage where the varactor capacitance is much more linear. The residual $1/f^3$ noise is mainly due to the transconductor pairs. The flicker noise associated to these transistors up-converts through the tail capacitance at the drain of the current generator [10], and does not depend on the AM-FM mechanism.

Fig. 10(a) shows the simulated voltage waveforms of the SE-VCO and BB-VCO, tuned at 0 V. Note that the single-ended tuned VCO waveform is clearly distorted and a strong second harmonic is present. In fact, as already shown in section II, for positive voltage the varactor is forward biased and the capacitance higher than for negative voltage when the varactor is reverse biased. That justifies the “slow” positive lobe of the voltage waveform and the “fast” negative one. Fig. 10(b) shows the harmonic distortion of the output signal, i.e. the ratio between the amplitude of the n-th harmonic and of the first one. The second harmonic amplitude in the BB-VCO is 40 dB under the
fundamental tone, and the second harmonic rejection is 25 dB better than in the SE-VCO. This result justifies why the second harmonic was neglected in the analysis reported in section III; instead, in SE-VCO the omission of the second harmonic causes a substantial error, as verified in section II.

V. CONCLUSION

In this work, the mechanism of amplitude-to-phase noise conversion in oscillators is investigated. The analysis is applied to three different diode varactor topologies: the single-end-tuned, the differentially-tuned and the back-to-back configuration. Compared to the conventional single-ended one, the differentially-tuned topology, which is sometimes adopted to guarantee immunity to common mode disturbances, features stronger AM-to-PM conversion and limits the oscillation amplitude. Instead, the back-to-back series configuration reduces considerably the up-conversion of flicker noise and causes no penalty in the achievable tuning range. Finally, it must be noted that the same topology can be applied to reduce the conversion effects also in MOS varactors. In this case it is not possible to obtain a closed form for the conversion coefficient, and one must rely only on simulations. When MOS technology are available the switched tuning approach, [1], is frequently adopted. A single varactor, p-n junction or MOS, is anyway necessary for a fine tuning, and the back-to-back topology helps in reducing any residual conversion effect.
REFERENCES


FIGURE CAPTIONS

Figure 1. The LC-tuned differential oscillator, with different varactor configurations: (a) single-ended, (b) differential, (c) back-to-back series.

Figure 2. Simplified tank schematic and equivalent C-V curve for single-ended-tuned VCO.

Figure 3. Simplified tank schematic and equivalent C-V curve for differentially-tuned VCO.

Figure 4. Frequency vs. amplitude (a) and $K_{AM-FM}$ factor (b) relative to single-ended tuned VCO, differentially-tuned oscillator and back-to-back configuration.

Figure 5. Simplified tank schematic and equivalent C-V curve for the back-to-back varactor configuration.

Figure 6. Tuning curve (a) and tuning constant (b) for single-ended, differential and back-to-back varactors configuration.

Figure 7. Frequency vs. amplitude for the back-to-back VCO with bias resistors of 1 kΩ (a) and of 1 MΩ (b).

Figure 8. Simulated phase noise spectrum for single-ended, differential and back-to-back varactors configuration at $V_{TUNE} = 0 \, V$.

Figure 9. Improvement of $1/f^3$ phase noise for the back-to-back oscillator respect to the single-ended tuned one.

Figure 10. Time waveforms (a) and harmonic distortion of the output voltage (b) for SE-VCO and BB-VCO.
Figure 1
Figure 2
Figure 3
Figure 4
Figure 5
Figure 6
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Figure 8
Figure 9
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