Optimization of SiGe HBT VCOs for Wireless Applications

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Abstract — **This paper describes the optimization of phase noise performance in fully integrated SiGe HBT differential LC-tuned voltage-controlled oscillators (VCOs) for wireless applications. An accurate expression for phase noise in SiGe HBT LC-tuned VCOs is presented which takes the nonlinear operation of the oscillator into account. Design methods are shown which minimize the different sources of phase noise toward the intrinsic limit set by the resonator quality factor. A set of 2 GHz SiGe HBT VCOs have been implemented in a 0.5 μm, 47 GHz SiGe BiCMOS process to provide experimental verification of the benefits of the design methods presented in this paper.**

I. Introduction

The present growth in wireless communication demands more available channels. This in turn leads to stringent requirement on the frequency stability of the local oscillators in wireless systems. Thus in recent years phase noise in fully integrated VCOs have been a topic of active research [1]-[2]. Due to their relative good phase noise performance, differential LC-tuned oscillators, are usually the preferred topology for fully integrated VCOs [3].

This paper presents the analysis and design of a fully integrated 2 GHz SiGe HBT LC-tuned VCO with very low phase noise implemented in a 0.5 μm, 47 GHz, SiGe BiCMOS process. SiGe HBT technology was chosen due to the excellent noise performance of the transistors in this technology [4]. First an accurate expression for phase noise in SiGe HBT differential LC-tuned VCOs is presented which takes the non-linear operation of the oscillator into account. Next the insight gained from the analysis is used to design a SiGe HBT VCO with very low phase noise.

The experimental results for a set of 2 GHz SiGe HBT VCOs shows the benefits of the design methods presented in this paper. The VCOs demonstrates very low-phase noise performance at low power consumption compared with other recent published results for SiGe HBT VCOs.

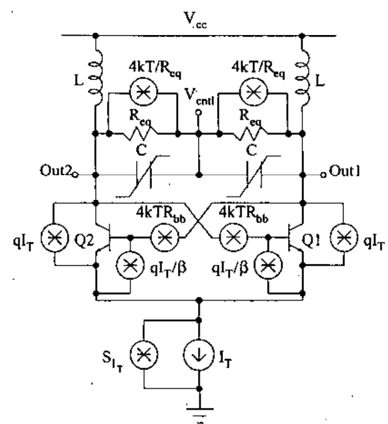

Fig. 1. SiGe HBT differential LC-tuned VCO with noise sources.

II. Phase Noise in SiGe HBT Differential LC-Tuned VCOs

A simplified schematic of a SiGe HBT differential LC-tuned VCO with its most important noise sources is seen in Fig. 1. The cross-coupled differential transistor pair presents a negative resistance to the resonator due to positive feedback. This negative resistance cancel the losses from the resonator enabling sustained oscillation. Frequency variation is achieved with a reverse-biased pn-junction diode or MOS varactor. The resonator losses are modeled as an equivalent resistance R_{eq} in parallel to the resonator.

A. Non-linear Phase Noise Analysis

The close-in phase noise behavior at an offset f_m from the carrier frequency f_0 in the differential LC-tuned VCO is found from Leeson's model [5]

$$\mathcal{L}(f_m) = \frac{2kTR_{eq}F}{A_0^2}\left(\frac{f_0}{2Qf_m}\right)^2\left(1+\frac{\Delta f_{1/f^3}}{f_m}\right) \quad (1)$$

where k is Boltzmann's constant, T is the absolute temperature, A_0 is the amplitude of oscillation, Q is the resonator loaded quality factor, $\Delta f_{1/f^3}$ is the corner frequency where $1/f$ device noise no longer predominate and F is the excess noise factor.

The excess noise factor is determined by the wide-band noise from the cross-coupled differential transistor pair and the tail current source taking the non-linear operation of the oscillator into account

$$F \approx 1+\frac{R_{bb}}{2R_{eq}}\left(\frac{f_t}{f_0}\right)+\frac{S_{I_T}R_{eq}}{8kT}+ \frac{qI_TR_{eq}}{4kT}\left(\frac{\Delta V}{\pi A_0}\right)^2\left(1+sinc^2\left(\frac{\Delta V}{2A_0}\right)\right) \quad (2)$$

where ΔV is the signal level required to make the cross-coupled differential transistor pair switch completely to one side, f_t is the unity current gain frequency, I_T is the dc tail current and S_{I_T} is tail current source noise spectral density. This expression is derived using conversion-matrix analysis [6] on the oscillator circuit assuming the only important non-linearity is the transconductance of the cross-coupled transistor pair.

B. *Up-conversion of Low-Frequency Noise*

The excess noise factor in Leeson's model only includes noise injected into the feedback path of the VCO. A different phase noise mechanism is the up-conversion of low-frequency noise sources due to the modulation of non-linear elements in the oscillator. Taking this up-conversion into account a more accurate expression for phase noise becomes

$$\mathcal{L}(f_m)=\frac{2kTR_{eq}F}{A_0^2}\left(\frac{f_0}{2Qf_m}\right)^2+\frac{|K_{I_T}|^2}{2f_m^2}S_{I_T}+ \frac{|K_{AM}|^2}{2f_m^2}\left(\frac{2}{\pi}\right)^2R_{eq}^2S_{I_T}+\frac{|K_{VCO}|^2}{2f_m^2}S_{R_v} \quad (3)$$

where $|K_{I_T}|$ and $|K_{AM}|$ are defined as the sensitivity of the frequency of oscillation on low-frequency tail current variations due to the indirect stability effect and varactor AM-to-PM conversion respectively [7]-[8]. The sensitivity of the frequency of oscillation on control voltage noise with spectral density S_{R_v} is described by the varactor gain $|K_{VCO}|$ [9].

III. Low Phase Noise SiGe HBT VCO Design

The implementation of a low phase noise differential LC-tuned VCO with noise filter and fully integrated resonator is seen in Fig. 2 and will serve as the basis for the following discussion. The variable capacitance for frequency tuning is implemented with a series connection of pn-junction diodes with capacitance C_v and a MIM-capacitance C_m. This configuration linearizes the overall capacitance variation with control voltage V_{cntl} at the expense of reduced tuning range. In the present design a quality factor $Q \approx 10$ is predicted for the fully integrated resonator at 2 GHz.

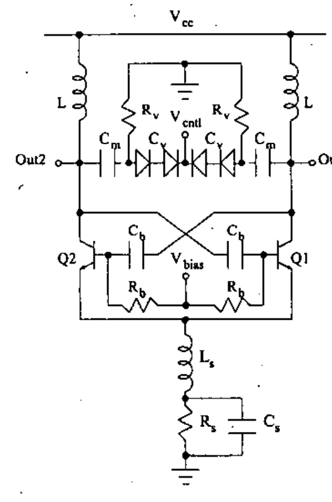

Fig. 2. SiGe HBT VCO with noise filter and fully integrated resonator.

A. *Improvement by Noise Filtering the Current Source*

In SiGe HBT differential LC-tuned VCOs the excess noise factor F is dominated by the noise from the tail current source near even harmonics of the carrier frequency. In order to improve phase noise this contribution has to be minimized. An efficient way of doing this is to use a noise filtering technique [10]. In Fig. 2 inductor L_s and capacitor C_s forms a 2nd order low-pass filter which prevents noise at even harmonics from being injected into the feedback path of the oscillator. Inclusion of different size inductors shows regions of both phase noise improvement and degradation over the first order low-pass case with the capacitor C_s alone. This degradation of phase noise is explained by the increasing inductance driving the oscillator into the saturation region as illustrated in Fig. 3. In this figure the simulated phase noise performance versus tail current is compared to calculated phase noise using (3) and neglecting the contributions from shot noise and tail current source noise. It is seen that the phase noise has been reduced toward the intrinsic limit set by the resonator quality factor. The reason for this is that inclusion of a large tail capacitance C_s makes the transistors in the cross-coupled differential pair conduct in pulses at the peak of the oscillation waveform. Due to the cyclostationary property of the shot noise sources this will be the ideal operation condition with respect to

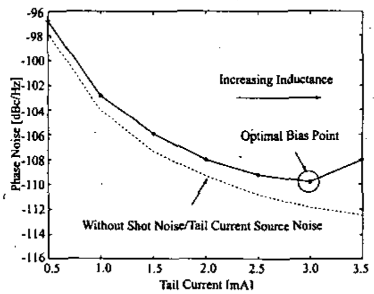

Fig. 3. Phase noise at 100 KHz offset from 2 GHz carrier versus tail current. Solid line: simulation, dashed line: calculated from (3) without shot noise and tail current source noise contribution.

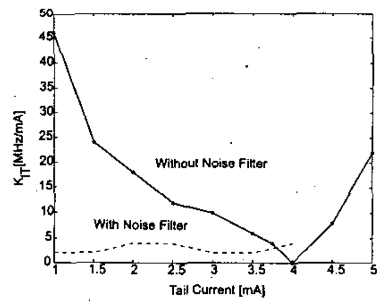

Fig. 4. Oscillation frequency sensitivity on tail current variations. Solid line: without noise filter, dashed line: with noise filter.

phase noise for the oscillator [1].

B. Indirect Stability and Effect of Varactor Non-linearity

The noise filter leaves low-frequency noise from the tail current source unaffected. Two mechanisms by which this gets transformed into close-in phase noise is by the indirect stability effect and the AM-to-PM conversion of the varactors. The indirect stability effect is due to the modulation of the phase shift in the feedback loop caused by low-frequency variations in the tail current. According to Barkhausen's criterion this results in a variation of the frequency of oscillation [8]. An open-loop gain analysis on the differential LC-tuned VCO without noise filter gives the frequency of oscillation as

$$f_0 \approx \frac{1}{2\pi\sqrt{LC}}\left(1 - \frac{1}{2\pi\sqrt{LC}}\,\frac{R_{bb}}{2Qf_tR_{eq}}\right) \qquad (4)$$

and predicts a rise in the frequency of oscillation at low current levels due to the current dependence of f_t. Once driven into saturation the frequency of oscillation drops due to the corresponding reduction of f_t in this region. When the noise filter is present the open-loop gain depends on the capacitor C_s which tend to reduce the variation of the frequency of oscillation with tail current. It is found however that this reduction depends on the device sizes with significant effect only for relative large devices. This is believed to be due to the fact that f_t is lower for larger devices at a given current level and therefore the reduction with the noise filter present is more significant for larger devices. From (3) it is seen that the contribution to phase noise due to the indirect stability effect depends on the sensitivity of the frequency of oscillation on tail current variations $K_{I_T} = \frac{\partial f_0}{\partial I_T}$. This sensitivity is evaluated for the SiGe HBT VCO with and without noise filter as seen in Fig. 4. The small sensitivity observed with noise filter present is due to the use of relative large devices which leads to reduced variation of the frequency of oscillation with tail current according with the above explanation.

Low-frequency noise from the tail current source is also up-converted to the carrier as amplitude modulation. Due to the non-linear $C - V$ characteristic of the varactors this amplitude modulation results in phase modulation. The contribution to phase noise depends on the frequency sensitivity on the oscillation amplitude variations $K_{AM} = \frac{\partial f_0}{\partial A_0}$. It can be evaluated by biasing the VCO where $K_{I_T} \approx 0$ and injecting a low-frequency current tone through the tail current source and observing the power in the up-converted sidebands. A sensitivity of only $|K_{AM}| = 6.4\frac{MHz}{V}$ at $V_{cntl} = 0$ is calculated for the SiGe HBT VCO because of the good linearity of the MIM-varactor configuration used.

Low-frequency noise on the tuning line modulates the non-linear capacitance of the varactors giving rise to phase noise variation with control voltage. The contribution to phase noise due to this noise is dependent on the frequency sensitivity on control voltage variations $K_{VCO} = \frac{\partial f_0}{\partial V_{cntl}}$. The phase noise degradation due to control voltage noise is very significant at the lower tuning range where the varactors are most non-linear [9]. The stack of two varactors as seen in Fig. 2 reduces the varactor gain K_{VCO} at the lower tuning range which in turn reduces phase noise variation with control voltage.

IV. Experimental Results

A set of three fully integrated 2 GHz VCOs have been implemented in IBM's 0.5 μm SiGe BiCMOS 5AM process. This process offers a thick top Al metalization layer for high Q-factor inductors. The first VCO (VCO1) is implemented with the noise filter and stacked varactors. The second VCO (VCO2) is identical to the first except that the inductor in the noise filter have been removed. A third VCO (VCO3) was implemented using single

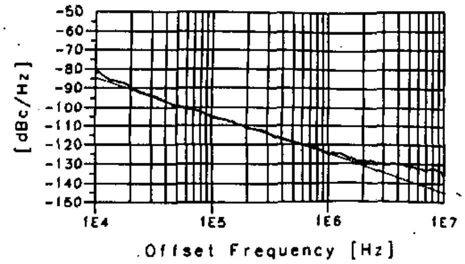

Fig. 5. Phase noise measurement between 10 KHz and 10 MHz offset from a 2.15 GHz carrier for VCO1. The dashed reference line have a 20dB/decade slope.

varactors scaled to provide the same overall capacitance variation as with the stacked varactors.

The VCOs were measured on-wafer using a Cascade probe station. Phase noise performance were measured using a HP 8563E spectrum analyzer with HP 85671A phase noise utility function. The wideband phase noise performance of the first VCO is shown in Fig. 5. Due to the very low $1/f$ device noise corner frequency in SiGe HBT technology the phase noise slope is 20 dB/decade until the noise floor of the measurement setup is reached. A summary of the measured performance for all three VCOs are shown in Table. I. The designed VCOs have identical phase noise performance; however the VCO without the inductor in the noise filter have higher power consumption than the other two. The phase noise variation over the tuning range is lowest in the VCOs with stacked varactors. This is due to the reduced tuning gain K_{VCO} at the lower tuning range. The tuning range however is lowered with stacked varactor which is explained by the larger parasitic capacitance present in this configuration. The single-ended output power for all VCOs are better than -5 dBm.

Different VCOs are compared using a Figure-of-Merit (FOM) that normalizes the phase noise performance to the same frequency, offset and power consumption [11] as

$$FOM = \mathcal{L}(f_m) - 20\log\frac{f_o}{f_m} + 10\log\frac{P_{diss}}{1mW} \quad (5)$$

	VCO1	VCO2	VCO3
Frequency [GHz]	2.150	2.161	2.156
Phase noise@100KHz [dBc/Hz]	-105.7	-105.3	-105.3
Tuning Range [MHz]	176	172	265
Phase noise variation [dBc/Hz]	2.47	3.00	3.84
Power consumption [mW]	10.8	12.2	10.8

TABLE I

SUMMARY OF VCO PERFORMANCE.

where P_{diss} is the dc power dissipated by the VCO. In this work we achieve a Figure-of-Merit of -182.0 dBc/Hz which is comparable to the best published results for Si/SiGe bipolar VCOs [7], [11]-[13].

V. CONCLUSIONS

Phase noise in SiGe HBT VCOs has been analyzed and design methods to reduce it toward the limit set by the resonator quality factor have been investigated. The design methods have been experimentally proven by measurement on a set of SiGe HBT VCOs showing low phase noise at low power consumption.

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