Optimization of SiGe 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 non-linear operation of the oscillator into account. The expression is derived using conversion-matrix analysis and include the transfer of wide-band noise sources in the LC-tuned VCO into close-in phase noise. 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 differential 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 where 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 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.

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.

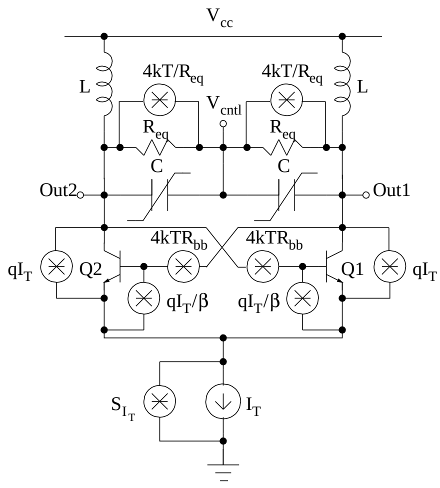

Fig. 1. SiGe HBT Differential LC-tuned VCO with noise sources.

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]

$$\mathcal{L}(f_m) = 10\log\left\{\frac{2kTR_{eq}F}{A_0^2}\left(\frac{f_0}{2Qf_m}\right)^2\left(1+\frac{\Delta f_{1/f^3}}{f_m}\right)\right\} \tag{1}$$

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 wideband noise from the cross-coupled differential transistor pair and the tail current source taking the non-linear operation of the oscillator into account

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

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. 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. Assuming instantaneous switching the amplitude modulation of the carrier is inherent cutoff in the analysis and the resulting contribution from wideband thermal noise due to resonators losses becomes identical to Leeson's model for $F = 1$. Thermal noise due to the base spreading resistance R_{bb} around odd harmonics of f_0 are modulated into the resonator passband with equal weights. The contribution to the excess noise factor becomes $\frac{R_{bb}}{2R_{eq}}\frac{f_t}{f_0}$ where a conservative estimate of the bandwidth limitation of the cross-coupled transistor pair has been used. This contribution can also be found by assuming sampling of a wideband noise spectrum with an infinite impulse train at twice the oscillation frequency [2]. The shot noise contribution only takes place in a small time-interval during the zero-crossing of the large-signal voltage waveform. The shot noise source when refered to the input of the transistors adds a factor of $\frac{qI_TR_{eq}}{4kT}(\frac{\Delta V}{\pi A_0})^2(1+sinc^2(\frac{\Delta V}{2A_0}))$ to the excess noise factor. The contribution from the input refered current noise source which may become important at higher current levels is neglected in this analysis. Low-frequency noise from the tail current source result in amplitude modulation of the carrier only. Noise around even harmonics of the oscillation frequency however result in both amplitude and phase noise modulation. The noise from the tail current source with spectral density S_{I_T} contribute a factor of $\frac{S_{I_T}R_{eq}}{8kT}$ to the excess noise factor in agreement with [2].

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

$$\mathcal{L}(f_m) = 10\log\left\{\frac{2kTR_{eq}F}{A_0^2}\left(\frac{f_0}{2Qf_m}\right)^2 + \frac{|K_{I_T}|^2}{2f_m^2}S_{I_T} + \frac{|K_{AM}|^2}{2f_m^2}\left(\frac{2}{\pi}\right)^2 R_{eq}^2 S_{I_T} + \frac{|K_{VCO}|^2}{2f_m^2}S_{R_v}\right\} \quad (3)$$

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 is seen in Fig. 2 and will serve as the basic for the following discussion. The dc current in this circuit is adjusted by varying the voltage V_{bias}. The capacitive coupling between the collector and base of the differential pair allows an amplitude of oscillation of approximately $V_{cc}/2$ before the transistors enter saturation. 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.

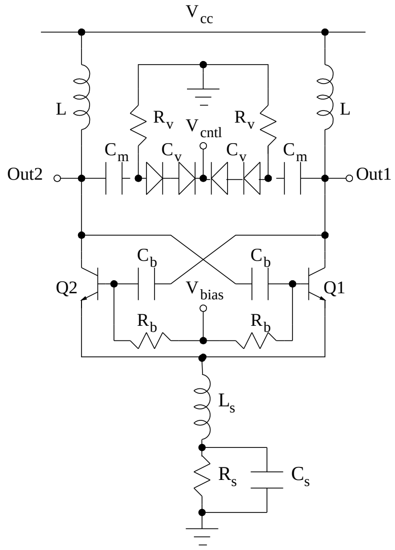

Fig. 2. SiGe HBT VCO with noisefilter and stacked varactors.

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 2. 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 1. order low-pass case with the capacitor C_s alone as seen in Fig. 3. The degradation

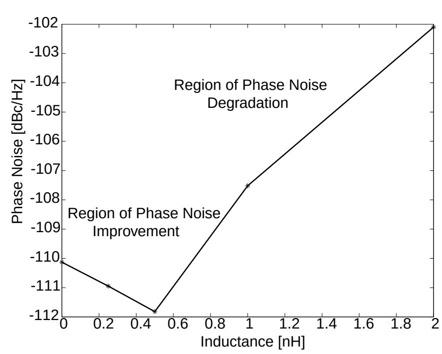

Fig. 3. Phase noise at 100 KHz offset from 2 GHz carrier versus inductance L_s. The best phase noise is found for $L_s = 0.5nH$ at a bias current of $I_T = 3.0mA$.

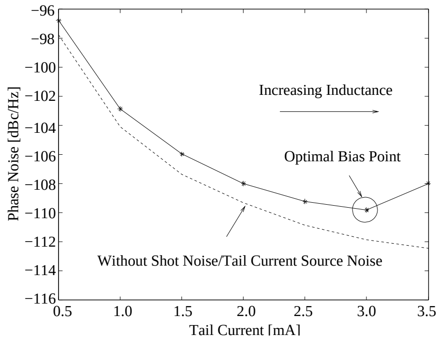

Fig. 4. Phase noise at 100 KHz offset from 2 GHz carrier versus tail current. Solid line: simulation, dashed line: calculated from (3) without shot noise and tail current source noise contribution.

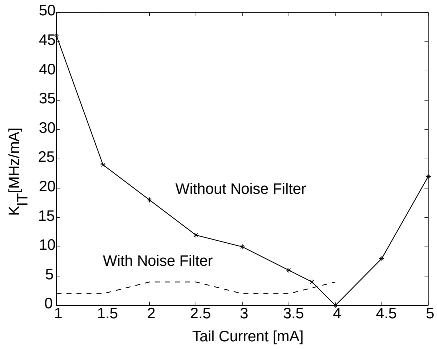

Fig. 5. Oscillation frequency sensitivity on tail current variations. Solid line: without noise filter, dashed line: with noise filter.

of phase noise is explained by the increasing tail current with increasing inductance driving the oscillator into the saturation region as illustrated in Fig. 4. 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 phase noise for the oscillator [1]. An equation for the conductance angle 2ϕ can be found by assuming that the voltage variation at the common-emitter node V_{common} is constant so that

$$2\phi \approx 2\arccos\left(\frac{V_{be,on} - V_{bias} + V_{common}}{A_0}\right) \tag{4}$$

where $V_{be,on}$ is the base-emitter turn-on voltage for the SiGe HBT. This serves as a design equation to choose the bias points in the oscillator to achieve the ideal operation condition.

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]. 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 $K_{I_T} = \frac{\partial f_0}{\partial I_T}$. Even though the noise filter leaves low-frequency noise unaffected it will reduce the frequency sensitivity on tail current variations to a negligible level as seen in Fig. 5.

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 of amplitude I_m at a frequency f_m through the tail current source. Then

$$|K_{AM}| = \left(\frac{P_{f_0 \pm f_m}}{P_{f_0}} \frac{2\pi f_m^2}{R_{eq}^2 I_m^2}\right)^{1/2} \tag{5}$$

where $P_{f_0 \pm f_m}$ is the power in the up-converted sidebands and P_{f_0} is the carrier power. A calculated sensitivity of only $|K_{AM}| = 6.4\frac{MHz}{V}$ at $V_{cntl} = 0$ resulted 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 reduce the varactor gain K_{VCO} at the lower tuning range which in turn reduce 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 is implemented with noise filter and stacked varactors. The second VCO is identical to the first except that the inductor in the noise filter have been removed. A third VCO was implemented using single 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 and batteries for the power supplies to avoid the low-frequency noise from regular power supplies. 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. 6. 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

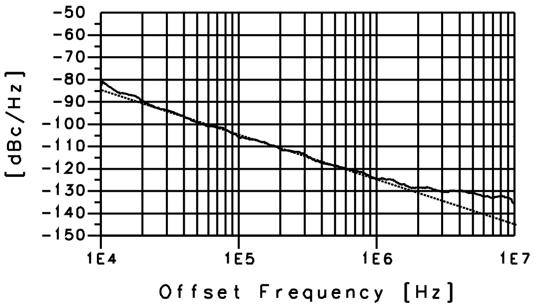

Fig. 6. Phase noise measurement between 10 KHz and 10 MHz offset from a 2.15GHz carrier for VCO1. The dashed reference line have a 20dB/decade slope.

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. As seen in Fig. 7 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.

	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.

The performance of fully integrated VCOs is affected by so many diverse factors that it is difficult to draw meaningful comparisons between various technology and circuit approaches. It is in general accepted to use a Figure-of-Merit (FOM) to compare different VCOs that normalizes the phase noise performance to the same frequency, offset and power consumption [11], where

$$FOM = \mathcal{L}(f_m) - 20\log\frac{f_0}{f_m} + 10\log\frac{P_{diss}}{1mW} \qquad (6)$$

and P_{diss} is the dc power dissipated by the VCO. Table. II compare recent published results for fully integrated Si/SiGe bipolar VCOs. 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.

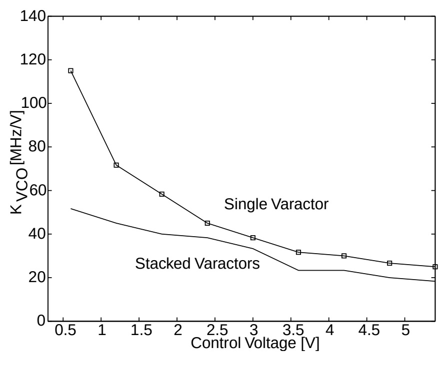

Fig. 7. Measured tuning gain K_{VCO} with single and stacked varactors.

Ref #	f_0 [GHz]	f_m [KHz]	P_{diss} [mW]	$\mathcal{L}(f_m)$ $[\frac{dBc}{Hz}]$	FOM $[\frac{dBc}{Hz}]$
[7]	2.56	100	14	-104	-180.7
[11]	5.05	100	15	-98	-180.3
[12]	1.5	25	25	-99.5	-181.1
[13]	1.35	3000	28	-142	-180.6
This Work	2.15	100	10.8	-105.7	-182.0

TABLE II

COMPARISON OF FULLY INTEGRATED SI/SIGE BIPOLAR VCO RESULTS.

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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