

LOW-NOISE BIASING OF VOLTAGE-CONTROLLED OSCILLATORS BY MEANS OF RESONANT INDUCTIVE DEGENERATION

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ABSTRACT

Many of the existing theories on phase-noise generation in oscillators are concerned with the noise originating from the active part and the resonator, while the contribution of the biasing noise is usually neglected. However, some recent theories have enlightened that the contribution of the biasing noise to the overall phase noise of the oscillator, actually, overwhelms all the other contributors together. Therefore, the concept of biasing to minimize the noise contribution of the oscillator's tail-current source is described in this paper. The procedure followed in the noise optimization of the inductively-degenerated low-noise amplifier scheme is here applied to the tail-current source. Noise at close to double the oscillating frequency is identified as the main contributor of biasing-noise to the over-all oscillator phase noise, and design procedure is developed accordingly. The presented analysis gives full insight into the performances of the low-noise biasing scheme, i.e., the resonant inductive degeneration of the current source, proving that this topology is an excellent trade-off between design effort and phase-noise performance.

1. INTRODUCTION

For the last few decades, in the field of the voltage-controlled oscillators (VCOs) and especially in the field researching their noise behavior, there have been introduced many theories, all of them competing to explain how the phase noise appears in the oscillators. In an LC-oscillator, consisting of an active part, acting as a negative resistance, an LC-tank, being a parallel resonating circuit, and the biasing network, those are the noise sources of the active part and the LC-tank to have been considered as the ones mainly contributing to the over-all phase noise, while, at the same time, the biasing is considered to be noise-free.

However, it has been recently determined from the mismatch between the measurement and the simulation results, that the phase noise is drastically worse than predicted. Therefore, new theories [1,2] have appeared, identifying the noise sources that actually have the largest impact on the generation of the phase noise. Accordingly, these theories have shown that the biasing tail-current noise can not be neglected from the calculation of the of the oscillator phase-noise, and moreover, it has been shown that

its contribution even overwhelms all the other contributions together. Following from these results, considerable attention has been paid to the elimination of the tail-current noise-source.

Therefore, a low-noise biasing scheme based on *resonant inductive degeneration* (RID) of a tail-current source (TCS) is introduced in this paper, where both the role of the degenerative inductance and the concept of noise reduction are completely different from all the nowadays known techniques of the tail current noise reduction [1,2,3,4]. Here, the concept of noise optimization for an inductively-degenerated low-noise amplifier (LNA) [5] is applied, where the optimization procedure is performed at double the resonating frequency. That is due to the fact that only the white noise of the biasing current source around the even harmonics of the resonating frequency is downconverted [2] into oscillator phase-noise, where the contribution of the second harmonic is the largest.

The paper is divided into five sections. A phase-noise model is presented in Section 2. An approach leading to the most comprehensive low-noise biasing scheme is presented in Section 3. Section 4 discusses our findings, while Section 5 summarizes the conclusions resulting from the presented analysis.

2. PHASE-NOISE MODEL

For the purpose of the analysis to come we will refer to a bipolar VCO, shown in Fig. 1, as to a quasi-tapped VCO [6], where the biasing is not completely shown. It is only the tail current source of the differential pair that is depicted, as our elaboration will be related only to that part of the whole oscillator. The parameters of the oscillator, as used in the SpectreRF simulation tool, are: the oscillating frequency $f_0=2.4\text{GHz}$, the tank inductance $L=1\text{nH}$, the tank capacitance $C=4.5\text{pF}$, the equivalent tank resistance $R_{TK}=400\Omega$, the quasi-tapping capacitances $C_A=0.8\text{pF}$ and $C_B=1\text{pF}$, the base-emitter capacitance C_{TB} , the quasi-tapping ratio $n=1+(C_A+C_B)/C_B=2$, the power supply $V_{CC}=2.5\text{V}$ and the tail current $I_{TAIL}=4\text{mA}$.

A simple, but also, rather intuitive phase-noise model for the oscillator shown in Fig. 1, can be found in [6] as:

$$\mathcal{L} = \frac{\bar{V}_{N,TOT}^2}{V_S^2/2} \quad \bar{V}_{N,TOT}^2 = KT \frac{G_{TK}}{(\omega_0 C_{TOT})^2} (1 + A_{QT}) \left(\frac{\omega_0}{\Delta\omega} \right)^2 \quad (1)$$

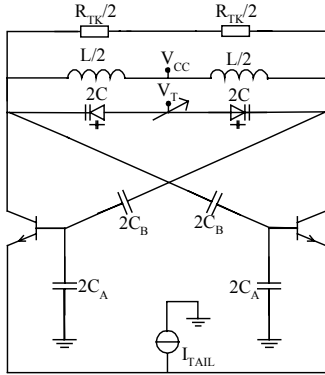


Fig. 1 Quasi-tapped LC-oscillator (biasing not shown).

$$A_{QT} = 2n^2 r_b G_{TK} + g_m / 4G_{TK} \quad V_s = \frac{2g_m V_T}{\pi G_{TK}} \quad (2)$$

where \mathcal{L} stands for the phase noise, $V_{N_{TOT}}$ for the equivalent voltage noise spectral density over the LC-tank, V_s the voltage swing over the tank, A_{QT} the noise factor of the active part, G_{TK} the equivalent tank conductance, ω_0 the angular resonant frequency, C_{TOT} the total tank capacitance, K Boltzmann's constant and T the absolute temperature. Inspecting Eq. (1), one can easily notice, that the effect of the biasing noise, i.e., tail-current noise, is not modeled since it was believed that there were no mechanisms converting it into the oscillator's phase-noise.

However, it has recently been shown [2,3] that the omission of the biasing-noise from the calculation of the over-all phase noise can over-estimate the oscillator performances to the extent that they are dramatically different from the actual ones. Namely, if the ideal tail-current source is replaced with the real one, as shown in Fig. 2a, the equivalent voltage-noise power over the tank equals:

$$\overline{V}_{N_{TOT}}^2 = KT \frac{G_{TK}}{(\omega_0 C_{TOT})^2} (1 + A_{QT} + A_B) \left(\frac{\omega_0}{\Delta\omega} \right)^2 \quad (3)$$

where A_B is the *biasing noise factor*. Even though there are many efforts to analytically express the effects of biasing on the phase noise [2], the mechanism of its transformation into the phase-noise is still not fully tracked. However, from the latest simulation tools as well as measurements results, the contribution of the tail-current noise can be fairly quantified, being larger than the contribution of all the other noise sources together. That is to say, the *tail-current noise* (TCN) may contribute to the equivalent phase-noise of the oscillator with 50% or more.

If we introduce the *phase-noise difference* factor as:

$$PND = 1 + A_B / A_{QT} \quad (4)$$

it can then simply be calculated to what extent the biasing-noise affects the phase-noise of the oscillator – Eq. (5).

$$A_B[\%] = 100 / [1 + 1 / (10^{0.1 PND[\text{dB}]} - 1)] \quad (5)$$

For example, supplying the SpectreRF simulation tool with the already given parameters of the oscillator, the following results are obtained: \mathcal{L} (ideal) = -131 dB@1MHz and \mathcal{L} (real) = -127 dB @1MHz from 2.4GHz oscillating frequency. Referring to Eq. (5), it is obtained, for the example under consideration, that the TCN contributes with 60% ($A_B[\%] = 60$) to the over-all phase noise, indeed confirming its largest impact on the noise performance.

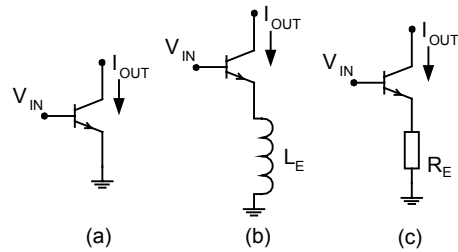


Fig. 2 Current sources (a) Real, (b) Inductively-degenerated, (c) Resistively-degenerated.

Even though, the mechanism of the conversion of the biasing-noise into the oscillator's resonator might not be well understood, it doesn't prevent us from developing a technique for its efficient removal. Simply, by reducing the output noise of the current source, its contribution, whatever it is, to the phase noise will be reduced accordingly. Therefore, our attention will be fully focused on the noise behavior of the tail-current source and not on the oscillator core or its tank, as their impact on the noise performance is well known and understood. Finally, once the biasing-noise is made negligible, all the performances of the oscillator will again depend only on its active part and resonator.

3. LOW-NOISE BIASING

Nowadays, it is known that both the low-frequency and the high-frequency noise sources of the tail-current source, have the most detrimental effect [1][2][3] on the phase-noise performance of the oscillator. In particular, it is the tail-current noise at twice the oscillation frequency that, after being down-converted by the switching action of the oscillator active part, has the largest impact.

Therefore, the noise-optimization procedure, introduced in this paper, is performed at double the resonant frequency, at the same time adapting the idea that is used in a design of the inductively-degenerated low-noise amplifiers [5]. To make the following analysis complete, we will undertake the comparative study of three different biasing schemes, being *simple* current source (Fig. 2a), the *resonant inductively-degenerated* (RID) current source (Fig 2b) and the *resistively-degenerated* (RD) current source [7] (Fig 2c).

However, let us first outline the new procedure behind the resonant inductive degeneration of the current source. The basic idea lies in *resonant-matching* at the frequency $2f_0$ of the inductor in the emitter of the biasing transistor to the transistors reactive part, being the base-emitter capacitance C_{IT} . In such a case, the effective transfer of the equivalent input voltage-noise power to the output of the biasing transistor is considerably reduced as being inversely proportional to a very large input impedance $Z_{IN} = 2\pi f_T L_E$, where f_T is the transition frequency and L_E the matching inductor.

In order to get insight into how the equivalent output-noise current density is reduced, and accordingly the "noise-portion" to be transferred to the resonator, we will introduce in the following sub-sections a gain model, a noise model and finally, a current-noise power ratio, for the three aforementioned current sources.

3.1. Gain model

From the literature it is well known that the input impedance of an inductively-degenerated transistor [5] equals:

$$Z_{IN}(\omega) = \omega_T L_E + j \left[\omega L_E - \frac{\omega_T}{\omega} \frac{1}{g_m} \right] \quad (6)$$

Relying on the *resonant-matching*, i.e., the imaginary part of the impedance is set to zero at the frequency $2f_0$, the following condition must be satisfied:

$$R_{IN} g_m = \left(\frac{\omega_T}{2\omega_0} \right)^2 \quad R_{IN} = \omega_T L_E \quad (7)$$

with the equivalent input impedance reduced to $R_{IN} = 2\pi f_0 L_E$.

Now, the effective transconductance, at the frequency $2f_0$, of the RID [8] current-source, shown in Fig. 2b, is simply obtained as:

$$G_{EFF}(sL_E, 2\omega_0) \cong -\frac{1}{R_{IN}} \frac{\omega_T}{2\omega_0} \quad R_{IN} \gg r_B \quad (8)$$

where r_B is the base resistance of the transistor.

On the other hand, the effective transconductance of the RD [7] current source, again at the frequency $2f_0$, can simply be calculated, from the circuit of Fig. 2c, as:

$$|G_{EFF}(R_E, 2\omega_0)|^2 = \frac{1}{R_{IN}^2} \frac{1}{1 + (2\omega_0/\omega_T)^2 (1 + r_B/R_E)^2} \cong \frac{1}{R_E^2} \quad (9)$$

where it was reasonably assumed that $g_m R_E \gg 1$ and $2\omega_0/\omega_T \ll 1$.

3.2. Noise model

In order to find the expression for the output current-noise spectral density, the equivalent input voltage-noise spectral density of the transistor, shown in Figs. 2a, 2b and 2c, must be calculated first. For that purpose, the amplifier noise model, with corresponding input noise sources is shown in Fig. 3,

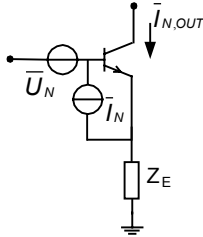


Fig. 3 Current source noise model.

where Z_E stands for either the feedback resistor R_E or the feedback inductor L_E .

Applying the Blakesley transformation to the voltage noise sources and splitting the current noise sources, at the same time keeping their orientation, the equivalent voltage noise at the input of the transistor is found to be:

$$\bar{U}_{N,EQ} = \bar{U}_N + Z_E \bar{I}_N + \bar{U}_{Z_E} \quad \bar{U}_{L_E} = 0 \quad \bar{U}_{R_E} = \sqrt{4KTR_E} \quad (10)$$

where \bar{U}_N and \bar{I}_N are the equivalent input-referred noise sources of the common-emitter transistor, expressed as:

$$\bar{U}_N = \bar{U}_B - B_N \bar{I}_C + r_B (\bar{I}_B - D_N \bar{I}_C) \quad \bar{I}_N = \bar{I}_B - D_N \bar{I}_C \quad (11)$$

$$\bar{I}_B^2 = 2qI_B \quad \bar{I}_C^2 = 2qI_C \quad \bar{U}_B^2 = 4KTr_B \quad (12)$$

with \bar{I}_B the base current shot-noise, \bar{I}_C the collector current shot-noise, \bar{U}_B the base resistance thermal noise, $D_N = -(1/\beta_F + j\omega/\omega_T)$

and $B_N = -1/g_m$ transistor transmission parameters. Yet, \bar{U}_{Z_E} represents the noise of the degenerative element.

Now, with the aid of Eqs. (10)-(12), the expressions for the input voltage-noise power, for the simple ($Z_E=0$), the resonant-inductive ($Z_E=sL_E$) and the resistive ($Z_E=R_E$) degenerated current source are, respectively:

$$\bar{U}_{N,EQ}^2(0) = \frac{4KT}{2g_m} [1 + 2r_B g_m + (r_B g_m)^2 \delta] \quad \delta = \frac{1}{\beta_F} + \left(\frac{\omega}{\omega_T} \right)^2 \quad (13)$$

$$\bar{U}_{N,EQ}^2(sL_E) \cong \bar{U}_{N,EQ}^2(0) \quad (14)$$

$$\bar{U}_{N,EQ}^2(R_E) = \bar{U}_{N,EQ}^2(0) + 4KTR_E \Psi \quad \Psi = 1 + r_B g_m \left(1 + \frac{\delta R_E}{2r_B} \right) \quad (15)$$

Finally, combining Eqs. (7)-(9) and (13)-(15), the equivalent output current-noise power, of the simple, RID and RD current source, at the frequency $2f_0$, equals:

$$\bar{I}_{N,OUT}^2(0, 2\omega_0) = g_m^2 \bar{U}_{N,EQ}^2(0) \quad (16)$$

$$\bar{I}_{N,OUT}^2(sL_E, 2\omega_0) = \left(\frac{2\omega_0}{\omega_T} \right)^2 g_m^2 \bar{U}_{N,EQ}^2(0) \quad (17)$$

$$\bar{I}_{N,OUT}^2(R_E, 2\omega_0) = \frac{1}{R_E^2} \left[\bar{U}_{N,EQ}^2(0) + 4KTR_E \Psi \right] \quad (18)$$

3.3. Current-noise power ratio

In order to compare the noisiness of the different biasing schemes, let us define the *ratio* of the *output current-noise power* of the simple and the RID TCS as well as the ratio of the output current-noise power of the RD and the RID source as follows:

$$INR(0, sL_E) = \frac{\bar{I}_{N,OUT}^2(0, 2\omega_0)}{\bar{I}_{N,OUT}^2(sL_E, 2\omega_0)} = \left(\frac{\omega_T}{2\omega_0} \right)^2 \gg 1 \quad (19)$$

$$INR(R_E, sL_E) = \frac{\bar{I}_{N,OUT}^2(R_E, 2\omega_0)}{\bar{I}_{N,OUT}^2(sL_E, 2\omega_0)} = \left(\frac{\omega_T}{2\omega_0} \right)^2 \frac{1 + 2g_m R_E \Psi / \Xi}{(g_m R_E)^2} \quad (20)$$

$$\Xi = 1 + 2r_B g_m \left(1 + \frac{g_m r_B \delta}{2} \right) \quad (21)$$

Inspecting Eq. (19), it becomes obvious to what extent the tail-current noise contribution to the phase noise can be reduced, applying the inductive degeneration to the source, with the simultaneous input matching at double the resonant frequency. As $\omega_T/2\omega_0$ is typically around 10, the $INR=100$ indicates that the biasing-noise contribution can be reduced up to 100 times with RID, making the biasing almost noiseless.

On the other hand, recognizing that the resistive degeneration [7,9] can also be very useful in reducing the output current-noise power of a tail current source, we will compare the performances of the RID TCS and the RD TCS, by transforming Eq. (20) into a more insightful form.

As the ratio Ψ/Ξ reduces to $\Psi/\Xi=1/2$, after simplifications $\delta \ll 1$ and $g_m R_E \gg 1$, the corresponding current-noise power ratio results into:

$$INR(R_E, sL_E) \cong \left(\frac{\omega_T}{2\omega_0} \right)^2 \frac{1}{g_m R_E} \quad (22)$$

Finally, from Eq. (22) the condition for the RID to be a better solution than RD can be written as:

$$g_m R_E < \left(\frac{\omega_T}{2\omega_0} \right)^2 \quad (23)$$

If typical value of $\omega_T/2\omega_0=10$ is assumed, then for the RID to be the best solution, the loop-gain $g_m R_E$ of the corresponding RD TCS must be lower than 100, which is for the supply voltages below 3V and VCO topology, as the one shown in Fig. 1, always satisfied. For example, for $g_m=0.2S$, corresponding to a mA tail-current range, there will be no voltage headroom for the integration of the resistance $R_E>500$, that would make the RD preferable. Therefore, the *resonant inductive degeneration* of the tail current source, allowing for the *largest suppression* of the biasing noise, is considered to be *always better* solution than the resistive degeneration of the current source.

For example, supplying the SpectreRF simulator tool with the, in Section 2, already given parameters of the oscillator, shown in Fig. 1, with the addition of the biasing scheme, shown in Fig. 2b (resonating inductance $L_E=4nH$), all for an operating condition characterized by $f_i=48GHz$ and $I_{TAIL}=4mA$, the equivalent phase-noise is $\mathcal{L}=-130.6dB@1MHz$ for a 2.4GHz resonant frequency. Now, from Eq. (5), it appears that the contribution of the tail current noise reduces from 60% to only $A_B=10\%$, indicating the possibility of almost approaching the *noiseless* biasing case.

4. DISCUSSION

For the last few years, considerable attention has been paid to the elimination of the contribution of the tail-current noise source. Some of the typical solutions are briefly discussed here.

In [1], a filtering technique is applied, where a capacitor placed in parallel with the current source is meant for the removal of the white noise over a range of frequencies. However, as it might be successful in improving the net phase-noise of the oscillator, there are two fundamental drawbacks of this approach. First, “*killing*” of the basic property of the current source, being a high output impedance, and second, a biasing node is at a *low-impedance level*, additionally degrading the oscillator phase-noise performance [3].

Another approach [3], proposes filtering again, but unlike previous solution, here, by the insertion of an inductor in between the current source and the oscillator core, it is solved the problem of high impedance in a common node at double the oscillating frequency. However, rather large values of the corresponding low-pass filter elements do not promote this solution into a favorite one from the *integration* point of view.

Analyzing the results of the previous section, it is apparent that the resistive degeneration [9] can also be considered as a concurrent of the resonant inductive degeneration. However, in nowadays, low-voltage low-power systems, it is likely that there is *no “voltage-room”* for the integration of the resistor in the emitter of the tail-current source. Simply, there is hardly enough voltage-room for the oscillator’s active part and the current-source themselves, directly discarding the possibility of use of any additional “voltage-hungry” element.

A more comprehensive solution can be found in [4], where tail-current noise suppression is achieved by combination of noise filtering and inductive degeneration of the current source. However, even though the idea of the inductive degeneration is also applied to a tail-current source in [4], there is rather fundamental difference with respect to our approach, regarding the role of the degenerative inductance as well as the noise reduction procedure. In [4], *it is given neither insight into how to choose for a certain inductance value, nor it is given any qualitative argumentation*, but crude simulation results, in favor

of the proposed technique. Also, a *large discrete* inductor in order of μH , as being used is in [4], may easily pick some external noise and by injecting it into the oscillator additionally degrade its phase-noise performance. Finally, in [4] it is not recognized that an inductance value in order of nH , as implied by the *resonant inductive degeneration* proposed in this paper, that can be easily designed and integrated on chip, allows for almost complete removal of the tail current noise, making the use of any larger inductor wasteful.

5. CONCLUSIONS

Recognizing, that the contribution of the biasing tail-current noise source to the over-all phase-noise of an oscillator is larger than all the other contributions together, an avalanche of research activities directed towards the elimination of this phenomenon, has been initiated.

In this paper, a novel low-noise biasing scheme, based on resonant inductive degeneration of the tail-current source, is introduced, leading to a reduction of the noise contribution of the oscillator’s tail-current source to a minimum. The effectiveness of this method is in the input resonant matching of the current source, which is performed at twice the oscillating frequency, as the largest part of the biasing noise is downconverted into the phase noise exactly from that frequency.

A comparative study has shown that the proposed noise-optimization procedure is by far the best solution as it allows for the largest reduction of the tail-current noise contribution to the over-all phase noise of the VCO.

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