Advanced Phase noise modeling techniques of nonlinear microwave devices

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ABSTRACT — **In this paper we present a coherent set of tools allowing an accurate and predictive design of low phase noise oscillators. Advanced phase noise modelling techniques in non linear microwave devices must be supported by a proven combination of the following :**

- **Electrical modeling of low-frequency noise of semiconductor devices, oriented to circuit CAD . The local noise sources will be either cyclostationary noise sources or quasistationary noise sources.**
- **Theoretically-based design rules are needed in order to define the most appropriate circuit architecture in function of the main characteristics to be optimized. For phase noise evaluation in feedback oscillators, the modified Leeson formula is a useful tool.**

Finally, results obtained with a 2 GHz HBT push-push VCO based on this methodology are presented.

I. INTRODUCTION

Designing oscillator circuits at RF and microwaves requires specific knowledge in extremely varied fields of electronics.

The following items will be the core of the presentation:

- The characterization method of electrical noise in order to extract realistic models that accurately describe the observed phenomena
- Behavior of low-frequency noise current sources in large signal operating conditions will be detailed, with a special emphasis on low-frequency noise modeling of HBTs
- The last section details the design rules leading to the optimization of the transistor oscillator phase noise, which is one of the prime characteristics in the design RF.

Theoretical conditions to be fulfilled by the circuit will be detailed on the basis of the Leeson analysis revisited.

II. LOW FREQUENCY NOISE MEASUREMENT

One of the main representations of a linear noisy two-port consists a noiseless two-port associated to two current noise sources placed at the two-port input and output access. These noise sources being correlated.

The measurement technique consists in to directly measure these noise currents sources at the input and the output of the two-port by using two amplifiers connected at a FFT spectrum analyzer.

Different setup can be used for this measure dependent on the amplifier used: voltage amplifier (VA) or transimpedance amplifier (TA). The choice of the kind of amplifier is done following several parameters :

- the impedance presented by the two-port : for high impedances, the transimpedance amplifier is chosen rather than for low or medium impedances for which the voltage amplifier is preferred.
- the second parameter is the noise current level : transimpedance amplifiers have bandwidth which varies with the sensitivity but to allow an amplification of very low noise currents which cannot be done with voltage amplifiers that have higher bandwidth and more moderate gain, insofar as we have compared them.
- the third parameter is the bias current needed by the transistor : the transimpedance amplifier generally amplifies the DC current flowing through the access and then rapidly saturates. It is a drawback for transistors with input bias currents of few mAmps. A capacitor must be inserted in series with the amplifier, but the lowest frequency measurements is then increased.

The setup we have developed is based on the diagram of Fig. 2 and Fig. 3. [1]

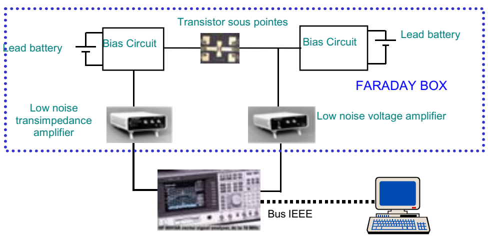

Fig 2 : Setup for measurement of noise currents sources of the transistor accesses

The spectral density of the noise current source of the transistor input i_1 is measured through a transimpedance amplifier connected at one channel of the FFT spectrum analyser. The transistor output is loaded at low frequencies by a low resistance R_m (compared to the transistor output resistance). The voltage generated by the noise current source i_2 is amplified by a voltage amplifier connected to the other channel of the FFT spectrum analyser.

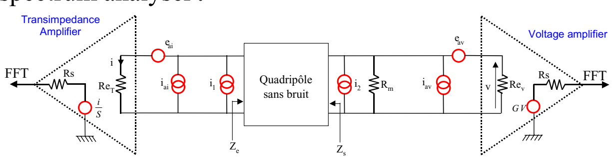

Fig 3 : electrical equivalent diagram of the measurement setup

In this diagram, S represents the transimpedance amplifier sensibility (in Siemens), G the voltage amplifier gain, Z_e and Z_s are respectively the input and output impedances of the two-port, Re_T and Re_v are the input resistance of the amplifiers. E_{ai} and i_{ai} , e_{av} and i_{av} are respectively the noise voltage and current sources of the amplifiers. The FFT spectrum analyser is the Hewlett-Packard HP89410A, the voltage amplifier is the PERKIN ELMER 5184 and the transimpedance amplifier is the PERKIN ELMER 5182.

A. Setup Calibration

The calibration procedure consists in the measurement of the amplifier parameters (gain, input resistance, noise sources) used to extract the transistor noise sources from the voltages spectral density measured with the FFT spectrum analyser. Some original procedures have been developed to measured these characteristics not generally given by the amplifier manufacturer. Figures 4 and 5 present the noise current and voltage spectral density of the transimpedance amplifier Si_{ai}, Se_{ai} and their correlation $Si_{ai}e_{ai}$. Its input impedance, measured for the 1µA/V sensibility is shown Fig. 6.

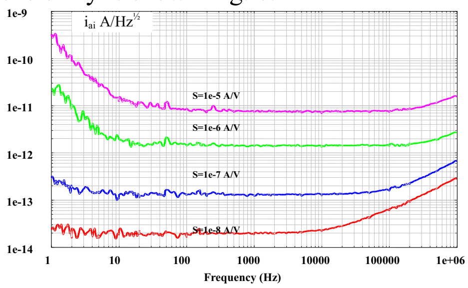

Fig 4 : input noise current of the transimpedance amplifier

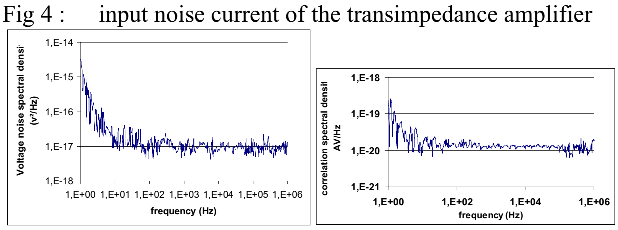

Fig 5 : Voltage noise and Correlation spectral densities of transimpedance amplifier

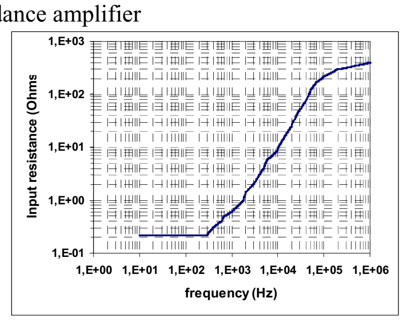

Fig 6 : transimpedance amplifier input impedance

This calibration procedure allows to accurately measure the input noise current source of the bipolar transistor for a large range of frequency and bias points.

B. Characterisation of bipolar transistors

Using this setup and the calibration procedure described behind, we show the input and output noise current spectral density of a HBT manufactured by UMS. It is a transistor with one emitter finger of 2 by 30 µm² size. As an example, Fig. 7 and 8 show the measurements results of Si_1 and Si_2 , in function of frequency, and for different emitter currents.

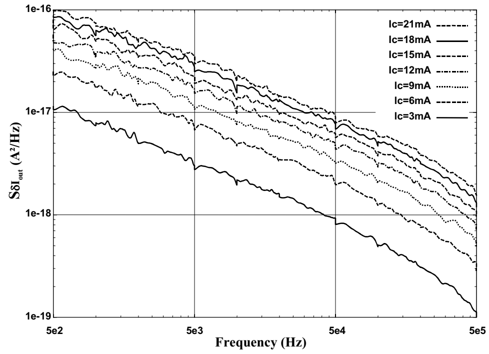

Fig 7 measured Si_1 for different emitter current bias

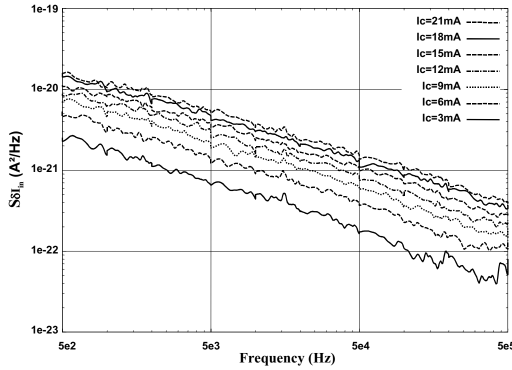

Fig 8: measured Si_2 for different emitter current bias

III. LOCAL FREQUENCY NOISE MODELLING

A. Low frequency noise sources in large signal operating conditions.[2]

In a simple one dimensional model the current flowing into a device can be written as:

$$I = q.A.n_0.v \qquad (1)$$

where v is the carrier velocity, n_0 the carrier density.
The LF current density fluctuation is due to fluctuations of either carriers, either velocity or mobility. Obviously, in a bulk ample A cannot fluctuate.
In semiconductor junction, the current flowing can be separated into two parts:

- a surface generation-recombination current
- a volume current

Then fluctuations of traps occupancy give rise to two kinds of low-frequency noise currents:

- Surface noise current due to recombination velocity fluctuations
- Volume noise current due to fluctuations of minority carrier density.

The volume noise current itself can proceed from : space-charge regions where the drift current prevails and quasineutral regions where the diffusion current prevails.
In all cases the noise current resulting from low-frequency fluctuations of a parameter P can be written as:

$$\delta i = I.\frac{\delta p}{p} \qquad (2)$$

where p is the parameter taken for carriers, velocity or mobility and δp the parameter fluctuating part. The product I by $\frac{\delta p}{p}$ give rise either to cyclostationnary noise sources, either to linear noise sources. The generation process leading to local low frequency GR noise current sources is synthesized in Fig. 9

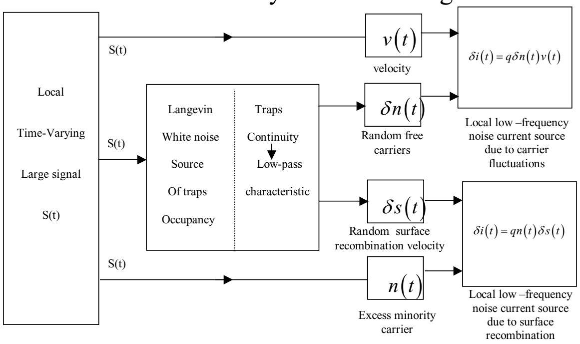

Fig. 9: Generation of low frequency GR noise current sources following the modulation scheme : S(t) , low frequency filtering, S(t).

The electrical modeling of low frequency noise sources of HBT will be done following this principle.

B Low-frequency noise modeling of HBT

As a general rule, two main contributors are sufficient to account for the low-frequency noise characteristics of a HBT. Figure 10 shows a realistic compact modeling of a given GaInP-GaAs HBT technology process, with two main low-frequency noise sources: i_{RB} which represents the noisy recombination diode of the base junction and i_B which represents the noisy volume diode of the base junction. There are inserted in the nonlinear model of HBT used for circuit CAD. [3-4]

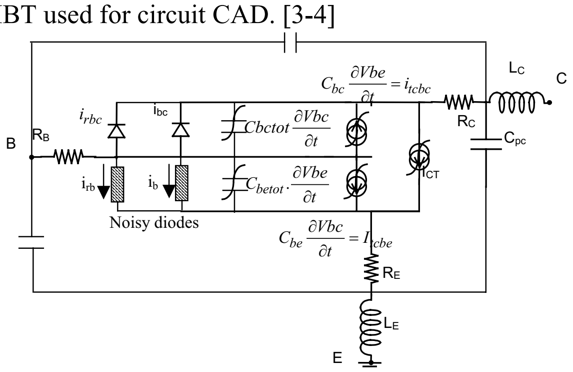

Fig 10: Nonlinear HBT model including LF noise sources

IV DESIGN RULES OF LOW PHASE-NOISE FREE RUNNING OSCILLATORS

A. Necessary rules for optimizing the design

A well-designed oscillator requires the optimization of the circuit behavior as follows:

- At the oscillation frequency ω_o. (large-signal)
- At the intermodulation frequencies near the carrier resulting from the interaction between (small-signal) noise sources and (large-signal) steady state.
- At low frequencies near DC, when 1/f and generation recombination noise sources are present.

We can infer that phase noise spectral density in a free-running oscillator circuit will be optimal [5]:

- when the nonlinear active element provides the maximum output power at the oscillation frequency ω_o.
- when this power is mainly dissipated in the resonator and, when the energy stored at the active device port reaches its maximum. This condition is achieved by tuning the transistor load line.

First optimizations conditions for low noise operation thus appear:

- optimization of the operating conditions at ω_o in terms of active device available power.
- optimization of the operating conditions around the oscillation frequency ω_o in stored energy terms. Note that this energy stored in a linear parallel resonant circuit is proportional to the susceptance slope versus frequency around ω_o.

B. A useful tool: the modified Leeson formula [6]

When choosing the architecture, designers must be able to evaluate the phase noise quickly and sufficiently accurately, in order to determine the most convenient architecture for the given application. Then numerical simulations will confirm the first evaluation, and will allow to fine tune the circuit. For a first evaluation of phase noise in feedback oscillators, the Leeson formula is a useful tool. Nevertheless a direct application of this formula to oscillator circuits with complex feedback tanks can lead to erroneous results because the formula contains coefficients which are not clearly defined. A successful application requires a modification of the parameters included in the formula.

If S_{In} is the spectral density around the oscillation frequency ω_o of the equivalent noise current source at the input terminals of the transistor. If the feedback tank is represented by a chain matrix [6], a modified Leeson formula valid for complex feedback tanks can be derived. It relates the spectral density of the phase noise of the oscillator, with the spectral density of the input noise current source, the elements constituting the oscillator circuit, and the peak value of the carrier signal at the frequency of oscillation. The phase noise $S(\Delta\varphi_{out})_\Omega$ can be written as :

$$S(\Delta\varphi_{out})_\Omega = 10\log\left\{\frac{S_{In}}{|V_{o1}|^2}\left|\frac{A_o}{C_o}\right|^2\left(1+\left|\frac{C_o}{C_1'}\right|^2\frac{1}{\Omega^2}\right)\right\} \qquad (3)$$

where: Ω is the offset frequency from carrier

V_{o1} is the peak value of the carrier signal at the frequency of oscillation.

A_o and C_o are the chain elements of the matrix describing the feedback circuit, taken at the oscillation frequency.

$C_1^{'} = \left. \frac{d\left(\mathrm{Im}(C)\right)}{d\omega} \right|_{\omega_o}$ is the slope factor of the imaginary part of the C element, taken at the oscillation frequency. Note that $-\frac{C_1^{'}}{C_o}$ is proportional to the group delay of the feedback circuit. It is directly related to the phase-frequency relationship of the oscillator loop gain. A detailed derivation of (3) can be found in [5]. From (3) it can be easily shown that :

- the amplitude of the oscillation voltage, V_{o1}, must be maximized at the input terminals of the transistor
- the spectral density S_{In} of the input noise source must be minimized.
- The slope factor, $\frac{d\varphi}{d\omega} = -\frac{C_1^{'}}{C_o}$ must be maximized.

This slope factor [6] is proportional to the slope factor of the energy stored in the circuit at the oscillation frequency. The previous conditions must be remembered at a time of designing low phase noise oscillator circuits.

V. 2 GHz VCO

A VCO has been designed by UMS foundry and characterized on-wafer for chip selection and then in test fixture for full characterization. The frequency response is given in Fig. 11, the centre frequency is 2 GHz, the tuning range is around 280 MHz (14% band)

Phase noise measurements have been done on a millimetre wave test bench based on frequency transposition and on delay line at IF frequency. The phase noise spectrum versus frequency offset from the carrier is shown on Fig. 12. At 2 GHz, the phase noise is –95 dBc/Hz at 10 kHz from carrier, and –118 dBc/Hz at 100 kHz. Moreover, Fig. 13 shows the phase noise increase for some low tuning voltages. This results are in good accordance with the predicted simulation results.

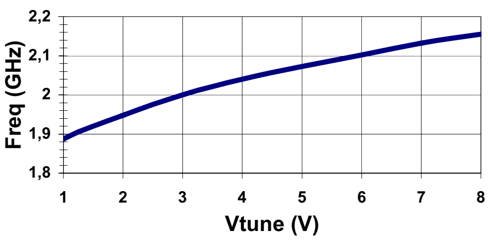

Fig 11: Frequency tuning characteristic versus tuning voltage

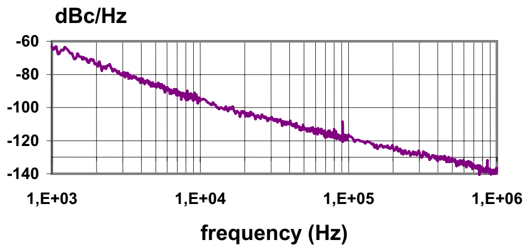

Fig 12: Measured phase noise for a tuning voltage at 4V versus frequency offset from carrier

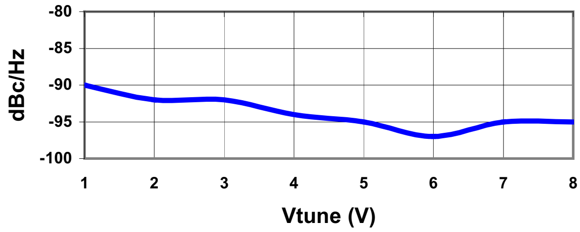

Fig 13: Measured phase noise at 10 KHz offset from carrier versus tuning voltage

VI. CONCLUSION

Designing oscillator circuits at RF and microwaves requires specific knowledge in extremely varied fields of electronics. We have presented in this paper several advanced topics which must be controlled by designers in order to reach optimized designs of low phase noise oscillators.

The major problems encountered today in the phase noise simulation of free-running oscillators are due to inaccurate modeling of the low frequency noise sources and their cyclostationary behavior.

Then some theoretically-based design rules are discussed in order to define the most appropriate circuit architecture in function of the main characteristics to be optimized.

Finally, 2 GHz HBT push-push VCO has been designed by UMS following this methodology. Its phase noise is –118 dBc/Hz @ 100 kHz from carrier for a center frequency of 2Ghz and 14 % tuning bandwidth.

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