# High Frequency and Low Frequency Noise in Microwave Oscillators

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*Abstract* — **This paper is a summary of our research in the field of noise in oscillators. An original approach has been developed to check the validity of the theoretical models and also, to investigate some techniques of noise reduction. It is based on open loop (or residual) phase noise measurements of microwave transistors. It allows us to identify the different noise sources contributing to the phase noise in a high frequency oscillator. The oscillator phase noise can then be simulated and optimised.**

## I. INTRODUCTION

Understanding and optimising phase noise in microwave oscillators has always been a difficult problem for microwave engineers [1-3]. Which device, or which technology, has to be chosen for low phase noise operation ? Can the classical device noise models be used in nonlinear regime ? Which circuit parameter could improve the phase noise performance ? All these questions have been asked many times, and a precise answer is no easy to give.

We will present our approach, which is based on investigations on RF active devices using residual phase noise data. These experiments have shown that, depending on the device and on the frequency offset from carrier which is of interest, the phase noise is either originated by the device low frequency excess noise (often improperly designed as 1/f noise) or from the device high frequency noise. In the first case, it is a multiplicative process, which means that the noise is following the carrier power (or, equivalently, the increase of the carrier power has no direct influence on phase noise). In the second case, it is an additive process, which means that an increase of the carrier power improves the carrier to noise ratio, and thus the phase noise. Both types of noise can be simulated using CAD. However the result will largely rely on the simulation techniques involved and, above all, on the device model, which must be as close as possible from device physics. In case such a model is too difficult to extract, it is still possible to optimise an oscillator with classical transistors models, but taking into account the nonlinear effects on the equivalent noise sources, and with a full knowledge of the model or the computing techniques limitations. The case of noise in synchronized oscillators is also discussed at the end of the presentation. The difficulties to simulate these circuits are pointed out but, once again, it is possible to get reliable results using simple analytical models, or by mixing these models with nowadays CAD tools.

## II. PHASE NOISE IN A MICROWAVE AMPLIFIER

The most famous model to describe the noise in an oscillator is probably Leeson's one [1]. Actually, this model is not really a phase noise modelling approach, it is more the description of the fundamental effect which appears in an oscillator loop : the transformation of the phase fluctuations in the loop into frequency fluctuations. This occurs of course only at the frequency offsets where the loop is closed, i.e. inside the resonator half bandwidth f/2Q. This effect is shown in Figure 1 for a bipolar transistor amplifier and an oscillator realised with this amplifier. Inside the loop bandwidth, the amplifier phase noise is converted into frequency noise, which results in an increase of 20 dB/dec of the spectrum slope.

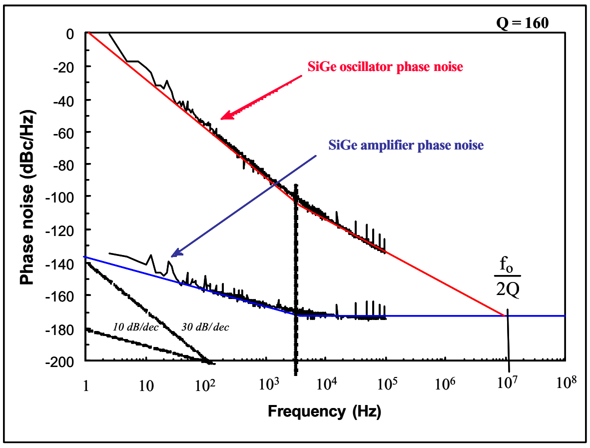


Fig. 1 : Phase noise to frequency noise conversion in a 4 GHz oscillator realized with an SiGe HBT (loaded Q : 160)

Two important cut-off frequencies can be noticed in this figure : the resonator half-bandwidth, where the two phase noise intersects each other, and the cut-off between the low frequency noise (in this case, featuring a 1/f shape) and the white phase noise floor of the amplifier. This last cut-off frequency can be much different from the transistor low frequency (LF) noise cut-off frequency, which is generally measured on the device output current fluctuations. The first reason is that both noises (1/f and white noise) are not necessarily originated at the same location in the device. This results in different conversion coefficients for both noise sources, and thus different cut-off frequencies between amplifier phase noise and transistor LF noise. But in the case of the device investigated in Figure 1, there is a different reason : the white noise floor is an (additive) HF noise and the 1/f component is an upconverted (multiplicative) LF noise. This has been demonstrated by varying the carrier power and measuring the amplifier phase noise [4]. As shown in Figure 2, the residual phase noise of this transistor, loaded onto 50 Ω at RF frequencies and using a 2 GHz carrier, features a strong change versus the input power with respect to the white noise floor, while the 1/f phase noise is almost no sensitive to the input power level.

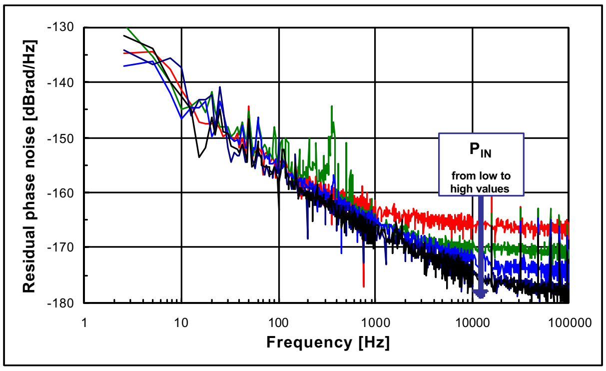


Fig. 2 : 2 GHz phase noise of an SiGe HBT, loaded onto 50 Ω,for different input power (up to 5 dB compression)

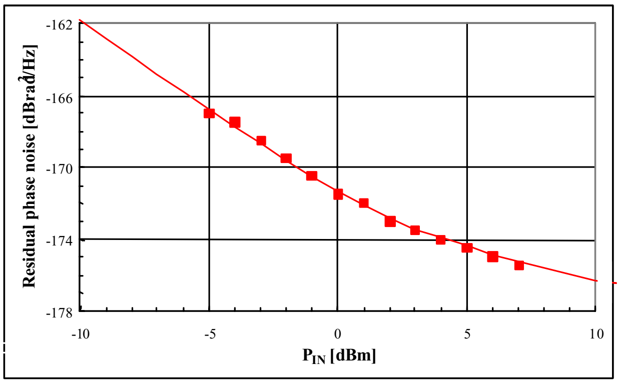


Fig. 3 : Evolution of the 100 kHz phase noise (same device and conditions than in Fig. 2) [4]

Actually, the white phase noise of this transistor is inversely proportional to the input power, at least before the compression. This agrees very well with the old phenomenological model for phase noise, already used by Leeson's, which involves the amplifier noise figure F :

$$S_\phi(f) = \frac{K_c}{f} + \frac{FkT}{P_{in}} \qquad (1)$$

$P_{in}$ being the amplifier input power and $K_c$ a arbitrary coefficient. However, not all the devices phase noise can be modelled by this simple equation. Firstly, in III-V devices and particularly in FET devices, the LF excess noise is high and generally only the first term in this equation is concerned at the offset frequencies of interest for many applications. Secondly, the noise figure of the transistor changes with device compression [4], in a very similar way than what happens with the LF noise sources under non-linear regime [5 to 9]. Figure 3 represents the 100 kHz phase noise of this amplifier, plotted versus the device input RF power. The measurement data are compared to a phenomenological model, extracted from noise figure measurements made under non-linear behaviour. The first linear part of the curve represents the power range on which the linear noise figure can effectively predict the phase noise. The second part requires a more sophisticated modelling, like in the case of the LF noise conversion.

## III. CONVERSION NOISE – PROBLEMS AND SOLUTIONS

The conversion of LF noise into phase noise (i.e. the multiplicative phase noise) is probably the most difficult problem to solve to compute the oscillator phase noise. The reason is that the low frequency noise sources in a device are very difficult to locate in the device model. Moreover, active devices electrical models are very often far from real device physics, and finding the right location for a noise source is, in this case, almost impossible.

Trying to locate a noise source in a nonlinear device model is necessary because, unlike in the linear case, the equivalent noise source approach is not valid. Indeed, as an example, a noise source which affects a nonlinear element will be converted, and another noise source which is at the transistor output can be very weakly converted [9]. Moreover, the noise source itself can be affected by the RF large signal, even if the time constants involved in the LF noise generation are very slow compared to the signal period. In other words, the noise does not depend only on the transistor DC conditions (even if these conditions includes the rectification of the microwave large signal) [10]. The way to take into account these effects in an equivalent circuit model is still a disputed subject. The noise source can be associated to a nonlinear element of the equivalent circuit [9], or considered itself as nonlinear. The last choice has been proposed very early [5], and is used by different researchers in this field [6,7]. However, an equivalent model is by no means a rigorous representation of devices physics, and may be the only accurate solution is in physical approaches [10] or microscopic models. But these models cannot be used directly to compute the noise in a complex system like an oscillator.

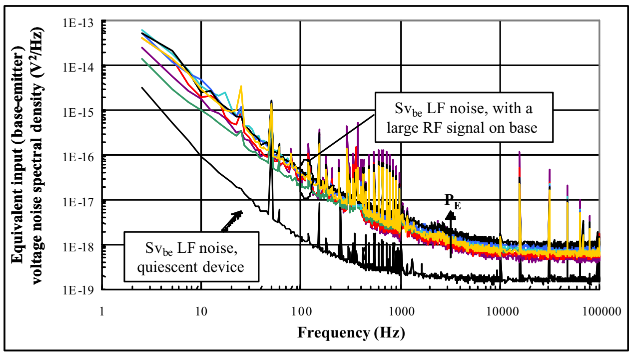


Fig. 4 : LF voltage noise at the input of an SiGe HBT

Our approach of the problem at LAAS these last years has been more based on a simplified technique, in order to be able not only to simulate, but also to optimise our circuits. In some cases, it is still possible to use an equivalent noise source to model the LF noise. As an example, in a bipolar transistor oscillator, most of the 1/f frequency noise comes from the LF fluctuation of the base emitter voltage. If an equivalent spectrum of these fluctuations can be measured, at different compression levels of the transistor, it can be used to compute the phase noise. Figure 4 shows the measured equivalent base-emitter voltage noise spectral density of an SiGe HBT, in quiescent operation mode and using a large pump signal at 3.5 GHz with different power levels (the RF signal source must be particularly stable for this measurement, mainly versus AM fluctuations). When the large signal is imposed, the noise climbs suddenly to an higher value, even for a compression level lower than 1 dB. Then, at higher compression levels, the climb is still going, but it is moderate. An average curve has been chosen in this set of spectra, in order to model the LF noise $V_{be}$ fluctuations of this device.

The amplifier or the oscillator is then optimised versus its phase noise performance using a CAD software. We use mainly Agilent ADS, which allows the direct simulation of the oscillator phase noise spectrum using two different and powerful techniques : a modulation technique (pnmf) and a conversion technique (pnmx). The first one is actually a quasi-static perturbation technique which can be implemented on any software by performing two successive large signal simulations with a small DC perturbation between the two, and examining the difference on an output parameter (frequency in case of an oscillator, or phase in case of an amplifier). The second technique (pnmx) is more powerful because it is able to take into account a low frequency dispersion in the device model. It should also be able to compute the additive noise floor or even the slope change between the phase and frequency fluctuations at the limit of the cavity half bandwidth.

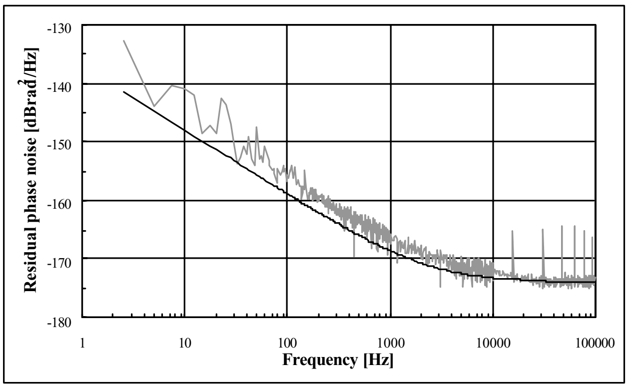


Figure 5 : Simulation and measurement of the phase noise of an SiGe HBT amplifier at 3.5 GHz [11]

Figure 5 depicts a comparison between theory and experiment for the phase noise of an SiGe HBT amplifier. In this case, the two contributions (1/f and white noise) have been computed separately. This amplifier was dedicated to an application for which the close to carrier (1 Hz to 100 Hz) phase noise specifications where the more stringent [11]. It has thus been optimised versus the 1/f phase noise performance using the quasi-static modulation technique. A systematic search for the lower conversion coefficient versus the input/output RF load has been performed, but keeping the input signal low (linear regime). It is indeed very difficult to optimise an oscillator versus all the parameters : RF load and power level. We have made, in most cases, the assumption that the conversion phenomena is not changing too much between the linear and nonlinear regimes [12] and that a minimum of this conversion can be found in linear operation mode (which is of course much easier and faster to compute, particularly in case of an amplifier).

One of the problem with this amplifier was related to its moderate gain performance. There is effectively a trade-off between gain and phase noise performance in microwave transistors, both for the additive or multiplicative noises. We are today designing a two stage low phase noise amplifier which should overcome this difficulty. Some results will be presented at the conference.

## IV. NOISE IN SYNCHRONIZED OSCILLATORS

In many cases, microwave oscillators are not used alone but are synchronized in some way, in order to obtain the mean term or long term stability required for the application. The reference oscillator is generally a quartz crystal oscillator, at lower frequencies. The synchronization is obtained through a PLL, using frequency dividers, multipliers, or both simultaneously. Phase noise in a PLL is generally calculated from the residual phase noise of all the elements of the loop and from the free running phase noise of the microwave oscillator.

A simpler configuration than the PLL is the injection locking. The injection of a small reference signal in an oscillator, even at a sub-harmonic or at an harmonic frequency, causes this oscillator to lock naturally on the reference. This simple behaviour (and circuit topology) hides some very complex nonlinear mechanisms which will determine the performance of the synchronized oscillator.

The two most famous papers on this problem are probably the references [13] and [2]. The first one gives a simple expression for the oscillator locking bandwidth, and the second one gives an expression of the noise behaviour. We found Kurokawa's expression very accurate to describe the injection locked oscillator phase noise, at least when the synchronising signal is not injected at the edge of the synchronisation bandwidth. On the other hand, Adler's locking bandwidth equation is, in our opinion, a bit too simple and difficult to use practically. However, the important inputs for Kurokawa's noise theory are : the oscillator free running phase noise (the modelling of which is described in paragraph II and III) and the locking bandwidth. It is therefore essential to be able to compute the locking bandwidth and, if possible, its dynamic behaviour.

However, modern CAD software do not provide a direct solution to this problem. A direct attempt to simulate an injection locked oscillator often fails because the oscillator start up behaviour is not taken into account, unless a special module has been implemented to this purpose. Fortunately, many research teams have worked on this subject, and have proposed some solutions. One of the most complete is the work of the teams of Limoges and Santander [14], which gives some techniques to simulate the locking bandwidth which may be used on a commercial harmonic balance software.

We have implemented one of these techniques to simulate the locking bandwidth and then the phase noise of different injection locked oscillators. Our application was the distribution of frequency reference signals using fiber optics. The goal is to replace in a system (in our case, a telecommunications satellite) the coaxial cables by fiber optics, which feature many advantages (losses, weight, size, signal isolation…). However, the decrease of the signal to noise ratio due to the distribution on many receivers (10 to 100) generates a relatively high phase noise floor at the system output. A solution to recover a spectrally pure signal is to synchronise an oscillator just after the optical receiver (sometimes, the amplifying device of the oscillator is optically sensitive and the oscillator and the optical receiver are the same device). This technique is known as the "photo-oscillator" technique, and is also used in optical data communications for clock recovery.

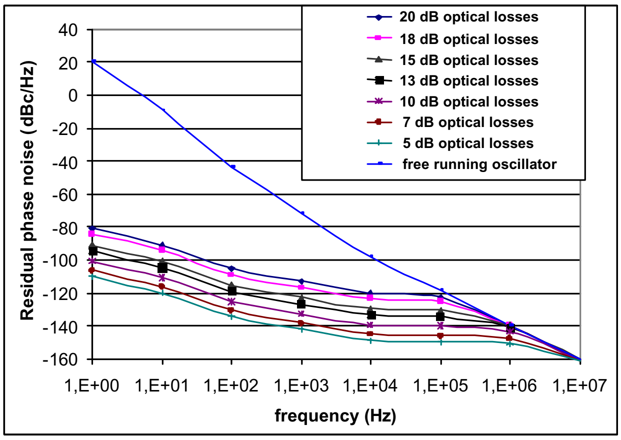


Figure 6 : Simulated phase noise of a fiber optics link using an injection locked photo-oscillator receiver at 874 MHz [15]

Figure 6 shows a simulated phase noise result for such a circuit (free running and synchronised noises, with different distribution factors, or equivalently different level of losses in the link). The gain and noise properties of the optical link are described through simple analytical expressions [15]. The phase noise is calculated from the (simulated) oscillator free running noise and from its (simulated) locking bandwidth using Kurokawa's equation. The results agrees very well with the final measurement (not shown here) performed on the whole system. The next step in such a design would be to optimize simultaneously the photo-oscillator free running phase noise and its locking bandwidth but, taking into account the complexity of the simulation techniques involved, this is a very difficult task.


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