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# Precision Close-to-Carrier Phase Noise Simulation of BAW Oscillators

Maxim Goryachev and Serge Galliou

**Abstract**—Based on a commercial simulation tool, the influence of BAW resonator noise on the resulting oscillator phase noise is revisited. The parametric model of the resonator uses experimental data, and includes an  $f^{-2}$  noise not often considered in measurements, in addition to its flicker noise.

## I. INTRODUCTION

MODERN commercial electronic simulators are necessary tools for oscillator analysis [1]. Such engineering tools, including Advanced Design System (ADS) from Agilent Technologies, Santa Clara, CA, cover the whole oscillator design procedure, including phase noise analysis. They can also clarify some intrinsic oscillator processes [2]. Indeed, not all important oscillator parameters are available for direct measurements. For example, the true phase noise of the oscillator loop alone is difficult to measure and partially hidden by the output-amplifier noise. Thus, the influence of the different loop components, including the resonator itself, on the resulting noise spectrum is not clearly distinguishable. So, in dielectric resonator oscillators (10–100 GHz) the active component noise dominates the resonator noise. For oscillators using acoustic resonators [bulk acoustic wave (BAW) up to 100 MHz, surface acoustic wave (SAW), and film bulk acoustic resonators (FBAR) up to 10 GHz], resonator noise cannot be ignored, as is usually done. At best, it is added manually by the designer after simulation. For example, for BAW [3], [4] and FBAR [5], resonator-caused noises are measured, but not included in resonator simulations. Obviously, this simplification reduces the validity of the close-to-carrier simulation results. In this paper, the noise of ultra-stable RF quartz crystal oscillators is covered. The resonator noise model, ignoring the noise origins (which are still not clearly identified [6]), is developed in the first section. This is followed in the next section by the implementation of the resonator model in oscillator simulation to determine the impact on the oscillator noise spectrum.

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## II. QUARTZ RESONATOR PHASE NOISE MODELING

To get a realistic phase noise model of the resonator, recent precision phase noise measurements have been used [3], [4]. Measurements are made with an interferometric bench. For all resonators, a double-oven temperature control (having the same quality as that in oscillators) is implemented [7]. The typical phase noise spectrum of a quartz crystal resonator consists of the following four zones: 1)  $f^{-2}$  is a random-walk process caused by remaining temperature fluctuations. Even though it is not usually taken into account, it always exists [8], [9] and can limit the phase noise performances at low frequencies. 2)  $f^{-1}$  is the intrinsic crystal noise within its bandwidth [10]–[12]; 3)  $f^{-3}$  is the intrinsic crystal noise outside the bandwidth; and 4)  $f^0$  is the noise floor of the measurement bench. For an unambiguous description of the resonator noise, three spectrum parameters can be extracted: 1)  $f_T$  is the corner frequency between  $f^{-2}$  and  $f^{-1}$  regions; 2) the phase noise level at 1 Hz; and 3)  $f_L$  is the corner frequency between  $f^{-1}$  and  $f^{-3}$  regions.

This type of measured spectrum has been implemented using the circuit shown in Fig. 1. The quartz crystal resonator is modeled with its conventional RLC branch (doubled here because a parasitic mode is close to the working mode for doubly-rotated crystal cuts) and its parallel parasitic capacitor, but including an additive current-controlled inductance, whose value depends on the current of two noise sources SRC1 and SRC2. Because the device resonance frequency depends not only on parameters of the motional branches but also on the variable inductor, the resulting system has a parametric noise with a spectral density proportional to the squared current of the noise sources. Thus, the first source with a power spectral density proportional to  $f^{-1/2}$  is used to represent  $f^{-1}$  and  $f^{-3}$  noises for medium and high Fourier frequencies. The second source, with a PSD proportional to  $f^{-1}$ , gives an  $f^{-2}$  slope for low frequencies. The parameters of the sources ( $I_{\text{Noise}}$ , which is the noise current magnitude per hertz<sup>1/2</sup>) are adjusted to simulate the phase noise of real 10-MHz doubly-rotated BVA-type resonators measured at the FEMTO-ST Institute. Fig. 2 presents the simulation results, i.e., the phase noise PSDs of the signal at the node out (see Fig. 1), which reproduce the measurement data [3], [4]. Noise spectra of two different samples are shown to demonstrate that all resonator samples have different parameters and noise characteristics, even if they are of the same type and made by the same manufacturer. It is also important to note that the quality of the temperature controller influences the recorded noise in the  $f^{-2}$  region.

Moreover, as mentioned previously, the resonator model consists of two resonance branches that model both the desired and spurious crystal modes. For example, an LD-cut quartz crystal resonator designed to work at 10 MHz

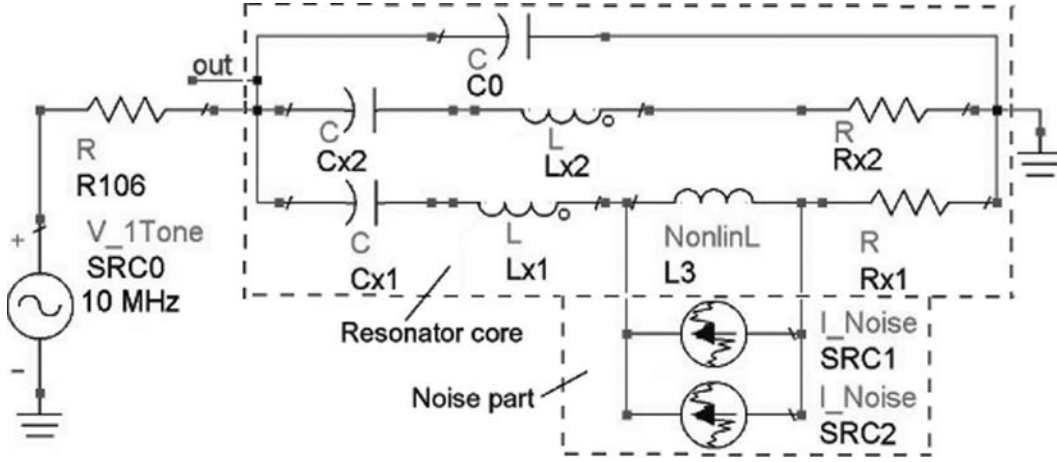


Fig. 1. Quartz resonator noise modeling circuit. Noise sources SRC1 and SRC2 have spectral noise densities proportional to  $f^{-1/2}$  and  $f^{-1}$  correspondingly.

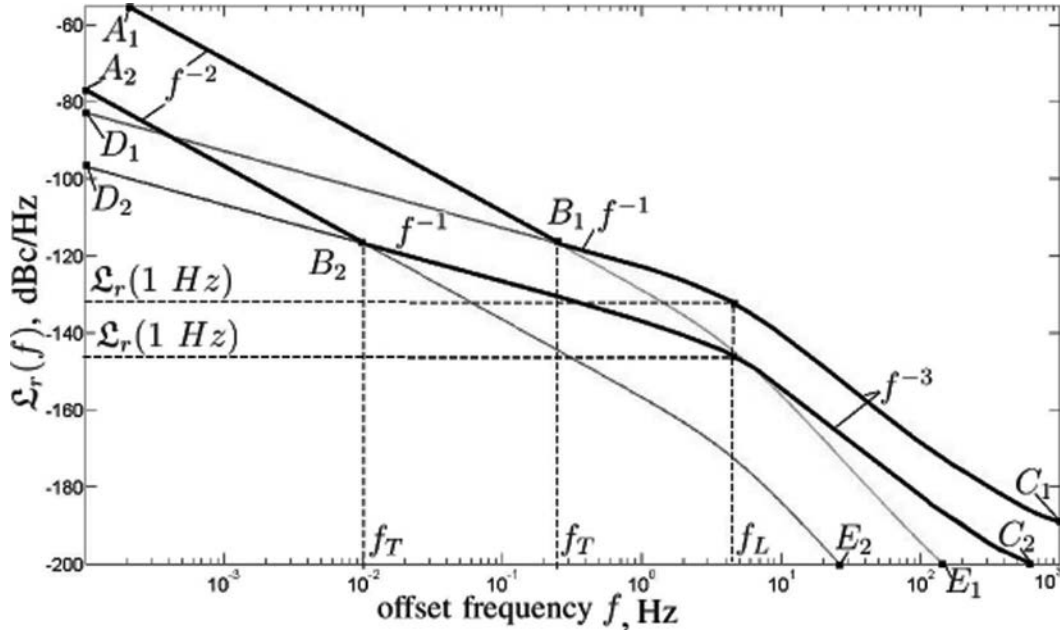


Fig. 2. The typical phase noise spectra of 10-MHz quartz crystal resonators with high ( $A_1B_1C_1$ ) and medium ( $A_2B_2C_2$ ) noise levels. Each spectra is an overlay of two independent PSDs, which correspond to sources SRC1 and SRC2: curves  $A_1B_1E_1$  and  $A_2B_2E_2$  originate from the source SRC2 (residual temperature fluctuations), and curves  $D_1B_1C_1$  and  $D_2B_2C_2$  are results of the source SRC1 (intrinsic noise).

on its C-mode has a parasitic resonance at 10.7 MHz (B-mode). This means that the spurious mode is inside the usual analysis band (which is defined here as close-to-carrier: typically 1 MHz from the carrier at 10 MHz) of the working mode. The parasitic branch does not have a considerable effect on the close-to-carrier phase noise, but it is used in simulations to verify that the oscillator operates on the expected mode. In some cases, additional elements are required to cancel the parasitic effects (see Fig. 3).

It should be noted that using a current-controlled inductance combined with current noise sources is not the only method available to modulate noise at the resonator. For example, a voltage-controlled capacitance across  $C_{x1}$ , linked with voltage noise sources, is also relevant [13]. In this case, the noise voltage magnitude per hertz<sup>1/2</sup> of the sources has to be adjusted accordingly.

### III. OSCILLATOR SIMULATION

The described model has been implemented for simulation of an actual 10 MHz ultra-stable quartz crystal oscillator designed at FEMTO-ST (see Fig. 3). This unit has a one-transistor Colpitts-type oscillation loop (the sustaining amplifier + the resonator), and a common-collector stage followed by an output cascode amplifier (i.e., two buffer stages are used to isolate the loop from the load, to match impedances and to purify the output signal).

Fig. 4 shows the simulated phase noise PSD of the oscillator in different configurations with or without crystal noise, and before or after the buffer stages to determine the influence of the various oscillator parts. In addition to the resonator noise, thermal noise sources of all component and transistor flicker noise sources are enabled.

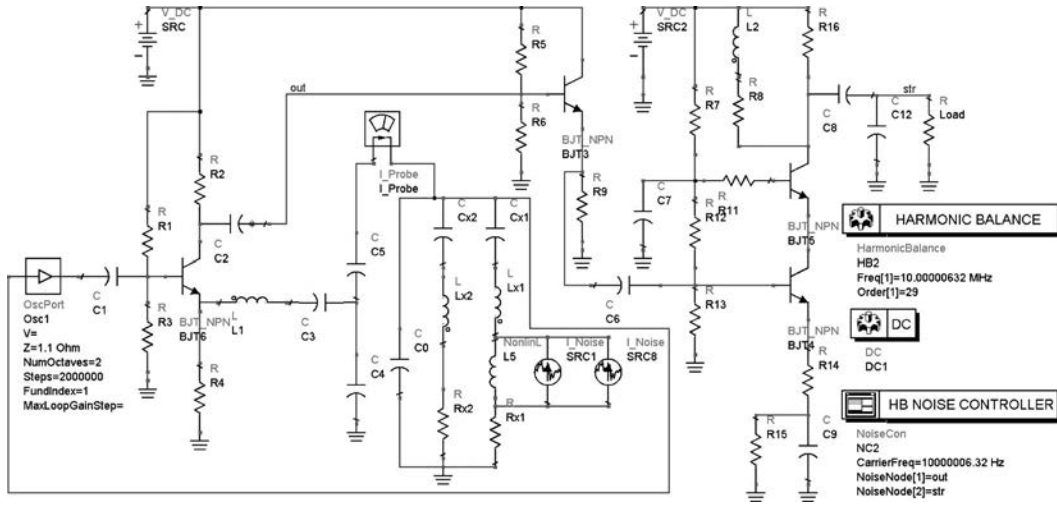


Fig. 3. Advanced Design System schematics of the modeled Colpitts oscillator with two output buffers. Note that capacitor C3 and inductor L1 are used to suppress the spurious mode (i.e., the B mode here).

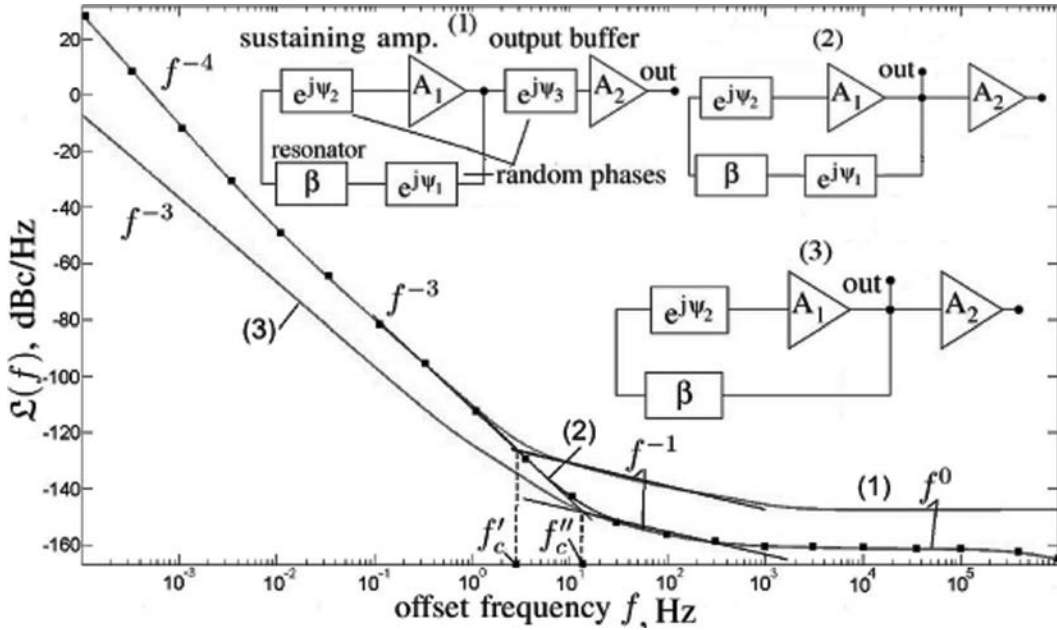


Fig. 4. PSDs of oscillator phase noise representing an actual 10-MHz oscillator in different configurations: (1) the oscillator PSD after the buffers, i.e., the PSD measured in reality, (2) the PSD before the buffers (also marked as a black square), (3) the PSD before the buffers without the resonator noise. Three generalized noise sources are present:  $\psi_1$ , resonator;  $\psi_2$ , sustaining amplifier;  $\psi_3$ , buffer phase noises.

From a practical point of view, the simulated curves correspond to actual measurements. The oscillator phase noise spectrum consists of  $f^{-4}$ ,  $f^{-3}$ ,  $f^{-1}$ , and  $f^0$  noises [14]–[16], which can also easily be identified in a typical RF oscillator phase noise spectrum. The  $f^{-4}$  spectrum slope is due to the Leeson effect [17] of the  $f^{-2}$  resonator noise.  $f^{-3}$  corresponds to the transformed  $f^{-1}$  resonator noise. The flicker and white noises of curve (1) are dominated by the noise of the output buffers, though the flicker and white noises of curve (2) are only due to the sustaining amplifier and resonator phase noise.

The first observation from these results is that white and flicker noise of the buffers overlap these types of noises of the oscillating loop. This leads to a decrease of the

corner frequency between  $f^{-1}$  and  $f^{-3}$  PSD slopes ( $f_c'$ ), which can be taken for the Leeson frequency by mistake [18], [19]. This error leads to an overly high estimate of the loaded quality factor. Actually, the Leeson effect takes place inside the oscillating loop, so the true Leeson frequency is the corner frequency between the  $f^{-1}$  and  $f^{-3}$  noises of the curve (2), or  $f_c''$ , which is hidden by the buffer noise. In conclusion,  $f_c'$  would give unjustified high values of the loaded quality factor. The Leeson frequency could also be found from curve (3), as the intersection between  $f^{-2}$  and  $f^0$  noise slopes, because the corner frequency of the sustaining amplifier is lower.

Second, as can also be concluded by comparing curves (1) and (3), the phase noise close to the carrier (both  $f^{-3}$

and  $f^{-4}$ ) is predominantly due to the resonator instability. Thus, for reliable ultra-stable oscillator noise simulation, the resonator phase noise model has to be included in the loop. The importance of the  $f^{-4}$  noise depends on the quality of the temperature control in the ovenized crystal oscillator as well as in the measurement bench of resonator noise. So, neglecting it (as is usually done both in measurements and modeling) leads to unrealistic simulation results close to the carrier. It should be noted that resonator aging can hide the effect of long-term temperature fluctuations, depending on their relative weights.

#### IV. CONCLUSION

The proposed resonator noise model is sufficient for precise simulations of ultra-stable quartz oscillators. Simulations are in good agreement with both resonator and oscillator phase noise measurements. This model implements remaining quartz crystal parametric noise, both caused by temperature and intrinsic fluctuations. This approach provides reliable results of phase noise close to the carrier. Also, the implementation of the  $f^{-2}$  resonator noise is necessary, because it limits the system performances in many applications. Our simulations confirm that actual oscillator phase noise performance close to the carrier is a strong function of the quartz-crystal resonator noise. In addition, far from the carrier ( $f^{-1}$  and  $f^0$  noises) the system stability is mainly limited by the buffer amplifying stages. The suggested approach may be very helpful for computer-aided ultra-stable crystal oscillator design.

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