Design and measurement of low phase-noise X-band oscillator

R. Boudot, Y. Gruson, N. Bazin, E. Rubiola and V. Giordano

The design and measurement of a low-noise 9.5 GHz oscillator based on a high-Q whispering-gallery sapphire resonator operating at room temperature is described. The oscillator features a phase noise of -65 dBrad^2/Hz at $f=10$ Hz off the carrier (slope $1/f^3$), of -145 dBrad^2/Hz at $f=10$ kHz (slope $1/f^2$), and of -160 dBrad^2/Hz at $f \gtrsim 100$ kHz (white). Owing to the low phase noise, the measurement requires cross-correlation and some unusual solutions.

Introduction: Because of the excellent $\nu_0 Q$ (resonant frequency × merit factor) product, the microwave whispering-gallery-mode sapphire resonator (WGMSR) is an excellent reference for oscillators. A phase noise of -167 dBrad^2/Hz at 10 kHz offset from the carrier has been demonstrated at 10 GHz [1], yet using a cumbersome noise degeneration technique. At microwave frequencies, the sustaining amplifier is the main source of phase noise. Owing to the parametric nature of flicker ($1/f$) noise, the phase noise tends to be proportional to gain. Therefore, the gain phase-noise trade-off is a major issue in oscillator design. Our solution, inspired by the 5 GHz oscillator described in [2], consists of using multi-stage SiGe HBT amplifiers, which show lower phase noise than the traditional GaAs devices. In this Letter we present the oscillator design and the phase noise measurement.

Oscillator design: The resonator is a low-cost Verneuil sapphire rod of 34.1 mm diameter and 17.05 mm height enclosed in a cylindrical duralumin cavity. The latter helps to preserve the high Q of the sapphire, and also warrants temperature uniformity. The Al cavity is stabilised to 40°C. The control reduces the temperature sensitivity of the resonant frequency from the open-loop value of -7×10^{-5} K^{-1} to -5×10^{-8} K^{-1} [3]. The main drawback of the outer cavity is that it interacts with the sapphire rod, giving rise to a number of low-Q spurious resonances at unpredictable frequencies. This is corrected with Cr lines printed on one end of the sapphire, which sets a boundary condition on the electric field, and in turn selects one resonant mode [4]. The modal filter, still not sufficient to prevent oscillation at spurious frequencies, is used in conjunction with a microstrip bandpass filter based on dual-behaviour resonators (DBRs) [5]. Such a filter shows a bandwidth of 2×10^{-2}, insertion loss of 4.6 dB at 9.51 GHz and rejection of 20 dB at 9.340 GHz.

The sustaining amplifier is a SiGe AML812PNB1901, which has a gain of 22.5 dB and a 1 dB compression power of +17 dBm, followed by two isolators and a bandpass filter. The resonator mode is the quasi-transverse magnetic field WGH$_{8,0,0}$ at $\nu_0 = 9.5$ GHz. Owing to machining tolerances, the difference between the two units is 6 MHz. The resonator merit factor is $Q \simeq 8 \times 10^4$ in actual load conditions, thus the Leeson (cutoff) frequency $f_L = \nu_0/2Q$ is 60 kHz. For low phase noise, the sustaining amplifier is operated in soft compression, close to the compression point. In fact, the oscillator white phase noise $S_{\varphi 0}$, which is additive, is ruled by

$$S_{\varphi 0} = \frac{FkT}{P_{\text{in}}} \qquad (1)$$

where F is the amplifier noise figure, kT the thermal energy, and P_{in} the power at the amplifier input.

Phase noise performance measurement: The expected phase noise is lower than that of tunable sources, such as microwave synthesisers, and also lower than the background noise of sophisticated instruments, such as the dual-channel delay-line homodyne [6]. In addition, for technical reasons it is virtually impossible to manufacture two resonators with overlapping passbands. Consequently, we cannot phase-lock two oscillators to measure the fluctuation of the error signal, nor we can design a homodyne system that makes use of a resonator of the same type as the discriminator. Thus, we chose to compare two nearly equal oscillators and to measure the phase noise of the beat note, which occurs at $\nu_b = 6$ MHz. Low phase noise is easier to obtain at this lower frequency because the time jitter is stretched by a factor ν_0/ν_b. A side benefit of the 6 MHz difference is that oscillators never lock to one another by injection.

The phase noise of the beat note is still too low for it to be measured by direct comparison with a synthesiser. This difficulty is mitigated by using a synthesiser (HP8662A) set to a higher frequency (384 MHz), divided by a suitable ratio (÷64) with binary counters. We then lock two equal synthesisers to the beat note and measure the cross-spectrum of the error signals (Fig. 1). It is to be made clear that correlation is necessary to reject the synthesiser noise, while the mixer noise is low enough. Conversely, in most cases found in the literature a low-noise reference is available, and the correlation method serves to reject the mixer noise. This explains the unusual practice of single-mixer beating before correlating.

On closer inspection on Fig. 1, the error signals are $x = k_\varphi(\varphi_b - \varphi_1)$ and $y = k_\varphi(\varphi_b - \varphi_2)$, where $k_\varphi = 0.35$ V/rad is the gain of the saturated mixer. Actually, x and y are weighted by the transfer function of the phase locking, which in our case has a small effect accounted separately. As usual, the cross-spectrum is measured as the product of the Fourier transforms, i.e. $S_{yx}(f) = \langle Y(f) X^*(f) \rangle_m$, averaged on m realisations of the process. Thus, assuming that φ_1 and φ_2 are independent (they originate in separate synthesisers), it holds that $S_{yx}(f) = k_\varphi^2 S_{\varphi b}(f)$. Residual terms proportional to $S_{\varphi 1}$ and $S_{\varphi 2}$ still remain, proportional to $1/\sqrt{2m}$, and vanishing for large m.

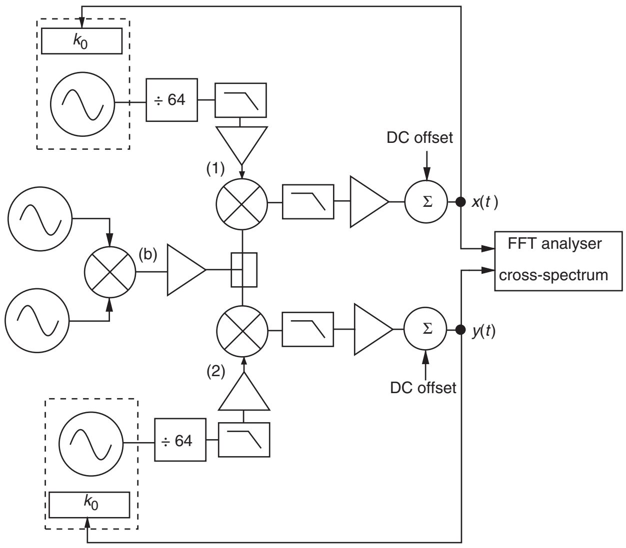

Fig. 1 *Cross-correlation measurement bench*

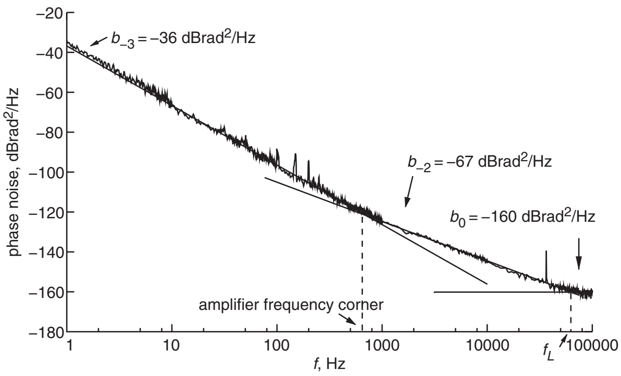

Fig. 2 *Phase noise of single oscillator*

The hypothesis of statistical independence deserves further attention. In fact the mixer offset is sensitive to microwave power, for the AM noise turns into DC noise in x and y. In the case of the beat note this effect is fully correlated, for it is not rejected. However a second-order effect shows up because the detection mechanisms get a small correlated random signal out of large noise. The literature [7, 8] suggests that mixers may be operated at a point of zero AM sensitivity, a few degrees off the quadrature, and to be determined experimentally. Yet our experience indicates that such a sweet point can be exploited only on microwave mixers. In fact, in the case of modern VHF mixers the sweet point turns out to be too far from the quadrature, where the DC offset is too high, and the system becomes noisy and unstable. We believe that this is a consequence of the high symmetry. In the end, the background noise turned out to be sufficiently small for the measurement, since we did not investigate further the AM noise.

Fig. 2 shows the phase noise of a single oscillator assuming that it is 3 dB lower than the oscillator pair noise. The spectrum is well described by the power-law $S_\varphi(f) = \sum_{i=-3}^{0} b_i f^i$, with $b_{-3} = -36$ dBrad^2/Hz, $b_{-2} = -67$ dBrad^2/Hz, and $b_0 = -160$ dBrad^2/Hz. Finally, the spectrum of Fig. 2 fits well the projection based on the Leeson model [9] and on the amplifier noise specifications [10]. This further validates the measurement.

Conclusions: Simple design makes possible the implementation of high spectral purity oscillators exhibiting a phase noise of -160 dBrad^2/Hz at 100 kHz offset frequency, and of -145 dBrad^2/Hz at 10 kHz. These oscillators can be used as reference sources in high sensitivity radar systems or in phase noise metrology and high-order frequency multiplication. Even lower noise can be obtained using HEMEX sapphire crystals, the unloaded Q factors of which can be up to 1.9×10^5 at room temperature [11], or by cooling the resonator.

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