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Phase Noise Measurement

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Abstract: *The analysis of the phase noise is presented. The phase fluctuation can be described of either standard deviation of the phase or time jitter or in detail as signal's spectrum. The digital receiver is proposed for the phase noise measurements.*

1. Introduction

Phase noise and the noise floor are fundamental properties of signals. Harmonic signals can be described with their amplitude and frequency. However, both the amplitude and the frequency are not stable in real life. A more accurate description of real signals is given by $A(t)\cdot\sin(\phi(t))$, where A(t) is a fluctuating amplitude with AM noise and $\phi(t)$ describes both the center frequency and a phase fluctuation noise. Practically, AM noise is very low in comparison, thus the first approximation to be made describes the signal as $A\cdot\sin(\phi(t))$. The instant phase $\phi(t)$ is described as $\omega_0\cdot t+\theta(t)$, where the phase fluctuation $\theta(t)$ is random and can be described in statistical terms. The phase fluctuation can be described of either standard deviation in radians rms or time jitter in seconds rms or in detail as the signal's spectrum.

The phase fluctuation $\theta(t)$ can be approximated as $m\cdot\sin(\omega_m\cdot t)$, where m is the modulation index and it is assumed that $m << 1$. The spectrum of so phase modulated signal is well described by means of Bessel functions. The spectral components related to Bessel coefficients of orders greater than 1 can be neglected due to small modulation index. Thus the spectrum contains just the carrier and two sidebands, whose energy has relative power to the carrier for a total of $m^2/2$. This results in conclusion that square of the phase deviation rms in radians^2 is equal to the relative power of the sidebands. The extension to random modulating signal is that the phase jitter rms in radians^2 is equivalent to the noise spectrum's power [1].

$$\theta j_{rms}{}^2 = \frac{1}{SNR} \qquad [rad^2] \tag{1}$$

Moreover, the phase jitter $\theta(t)$ may contain both the random and deterministic fluctuation. In this case the spurious signals add phase jitter exactly as random noise.

The spectrum of the signal is usually defined by the function $L(f_m)$ in dBc/Hz, which

$$\theta j_{rms}{}^2 = 2\cdot\int_{f_1}^{f_2} L(f_m)\,df \qquad [rad^2] \tag{2}$$

indicates the single-sideband (SSB) noise contribution in a 1 Hz bandwidth relative to total power at an offset f_m from the center frequency. To obtain the contribution of the sideband noise from the frequency range of interest <f1; f2> to the phase jitter θj_{rms}, the integral should be solved.

A more frequent description is time jitter τj_{rms}, which is usually given in picoseconds.

$$\tau j_{rms} = \frac{\theta j_{rms}}{\omega_0} \qquad [s] \tag{3}$$

2. Phase noise measurement

Three main methods are used to measure phase noise. The first method involves measuring phase noise ($L(f_m)$) directly on a spectrum analyzer. This measurement can be done as long as the analyzer has better phase noise than the measured source. In the second method, a better source is phase locked to the same frequency with the 90° offset. The mixed product of the unit under test and the reference signal is measured using a FFT analyzer. The measurement of signal total power and calibration of the system is done by means of the premixing of input signal with a 1 kHz offset. This system has the best sensitivity (in the range of -175 dBc/Hz), however, the reference synthetizer must be of supreme quality. The third method uses a discriminator and compares the signal to itself delayed in time. This measurement is limited for close-in offsets, because a large part of close-in noise is canceled. The big advantage of this system is that it does not require an excellent reference.

Besides these methods, the phase noise can be measured by using a digital receiver. The demand on the excellent reference remains, however, it can be fixed. The sensitivity is limited with the dynamic range of the contemporary receivers, which is not higher than 150 dBFS/√Hz. Also the frequency range is limited to the range of 100 MHz.

3. Phase noise and jitter

The jitter can be measured indirectly with a digital receiver from its dynamic range degradation due to jitter. A harmonic signal of sufficiently high frequency f_{in} is put at the receiver input, while the unit under test generates the clock signal for the receiver (Fig. 1a).

$$\frac{1}{SNRj} = \left(2\cdot\pi\cdot f_{in}\cdot\tau j_{rms}\right)^2 \qquad [-] \qquad (4),$$

where the SNRj is the limitation of the signal-to-noise ratio due to jitter of clock signal τj_{rms}. This equation is in good relation to equation (1) and (3) besides the frequency, which is different.

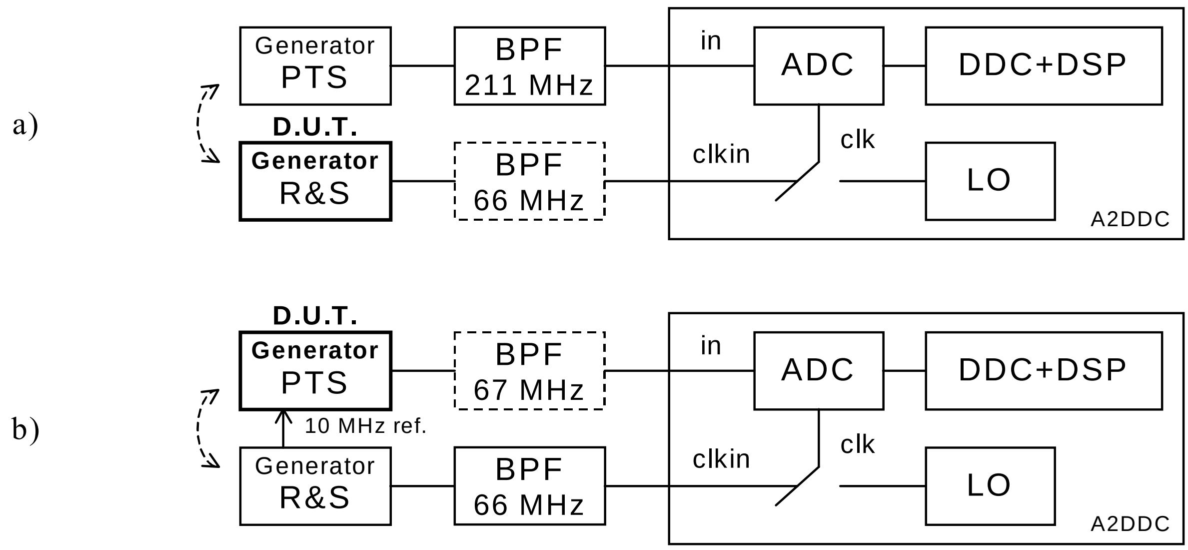

Figure 1. The high-frequency jitter measurement (a) and low-frequency phase noise measurement (b).

The auxiliary frequency f_{in} of 211 MHz is chosen for the best measurement sensitivity. The auxiliary signal has to be filtered otherwise the noise floor of the auxiliary generator is measured instead of the jitter effect.

The results of the jitter measurements are in Table 1. The various configurations of generators allow the jitter measurement of all three generators: the on-board crystal oscillator, the Rohde & Schwarz SMS generator and the PTS 500 generator. The resulting jitter of the both measurements with onboard oscillator should be the same and actually it is. This limit of 0.27 ps represents not only the jitter of oscillator itself, but in addition the jitter of the ADC and the jitter of the clock distribution.

	Clock source generator of 66 MHz					
	Onboard crystal osc.		R&S (211 MHz from PTS)		PTS (211 MHz from R&S)	
	(211 MHz from PTS)	(211 MHz from RS)	direct	with BPF	direct	with BPF
SNR_{max} [dBFS/√Hz]	143.9	144.1	126.3	141.6	114.5	142.4
τj_{rms} [ps]	0.28	0.27	2.1	0.36	8.2	0.33
θj_{rms} [rad . 10^{-3}]	0.12	0.11	0.87	0.15	3.4	0.14

Table 1. High-frequency time and phase jitter acquired from the dynamic range measurement.

However, those methods utilise the noise measurement at distant offset from the carrier and thus measure just the high-frequency parts of the overall jitter. To measure the close-in noise, the measurement according to Figure 1b is necessary. The digital receiver is tuned on the carrier and the noise distribution is obtained using FFT. Unfortunately, the phase noise of the clock source has influence on the resulting phase noise. The Figure 2 represents the phase noise at various configurations for the analog input and the clock generation. The lowest phase noise is achieved, when the PTS generates the analog input signal and the onboard oscillator generates clock. Whenever the Rohde & Schwarz generator was employed, the resulting phase noise is degraded. The measured phase noise of the R&S SMS generator of -107 dBc/Hz at 2 kHz offset is in good relation to the generator specification of -106 dBc/Hz.

The measured phase noise distribution can be translated to the equivalent time jitter according to (2) and (3). The jitter contribution from the range of 10 to 100 Hz is greater (PTS 14 ps / R&S 45 ps) than from the range of 100 to 1000 Hz (PTS 0.75 ps / R&S 14 ps).

4. **Conclusion**

The digital receiver can be well used for the phase noise measurement on condition that the phase noise of the clock generator is lower than of the device under test. The advantage of this measurement is that the clock generator is fixed for the wide frequency range of measured signals.

The time jitter values of clock sources should be specified at a given frequency range of the jitter. If not specified, the high-frequency jitter is supposed and it is related to the noise floor of the clock source. Because the contribution to the overall jitter decreases with the offset from the carrier, it is reasonable to specify the jitter for frequencies greater than a given frequency. For example, ValpeyFischer specifies the jitter of 1 ps at fj > 1 kHz for their crystal clock oscillator VFAC170.

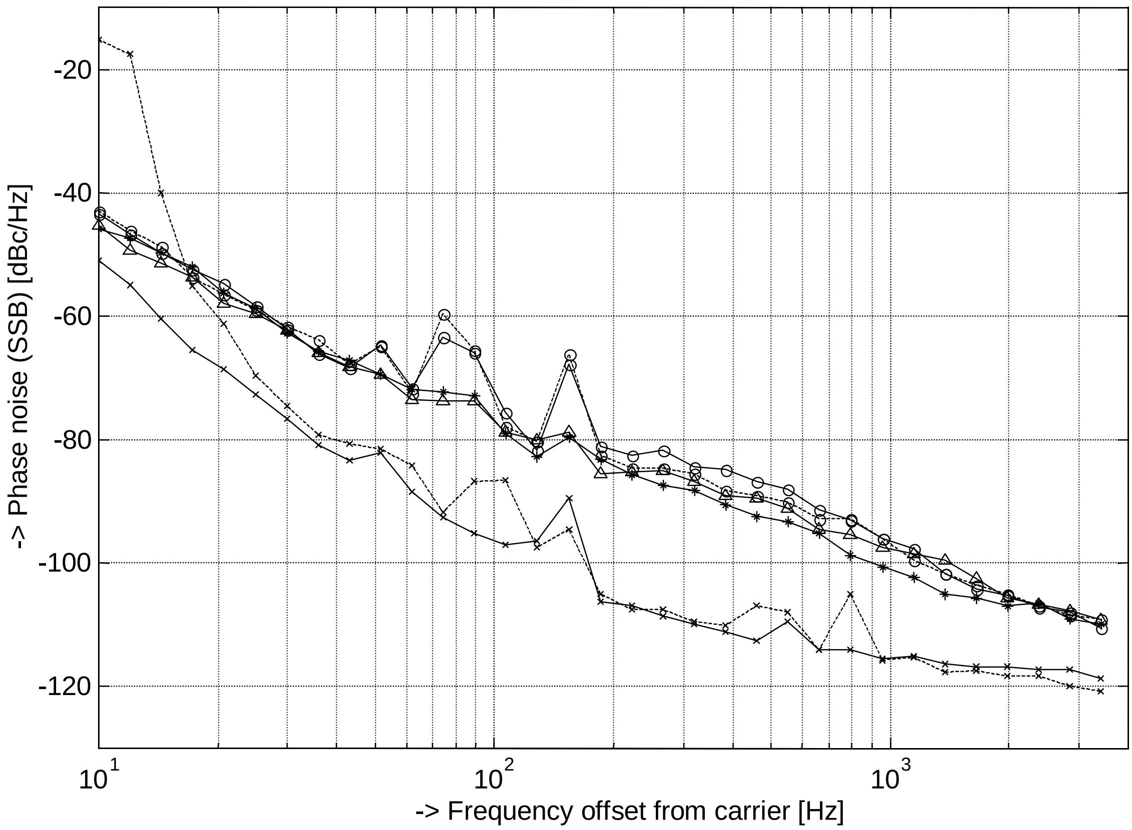

Figure 2. The low-frequency phase noise at various generator configurations. The analog input is 67 MHz and the clock is 66 MHz. The R&S and PTS (bundled) generator configuration represents '∗' mark, the PTS (bundled) and R&S 'Δ', the PTS and LO '×', the R&S and LO 'o'. The filtered analog input is represented with a dashed line.

References:

[1] B. Goldberg, Phase Noise Theory and Measurements: A Short Review, *Vitcom Corp., Tutorial,* 2000

[2] B. Brannon, Aperture Uncertainty and ADC System Performance, *Analog Devices, AN-501,* 1998